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Ph.D. Thesis

AGU 2024

HF SIGNAL INJECTION AND SLIDING MODE ESTIMATOR-BASED SENSORLESS CONTROL OF IPMSM

Ph.D. THESIS
SUBMITTED TO THE DEPARTMENT OF ELECTRICAL AND
COMPUTER ENGINEERING
AND THE GRADUATE SCHOOL OF ENGINEERING AND SCIENCE
OF ABDULLAH GUL UNIVERSITY
IN PARTIAL FULFILLMENT OF THE REQUIREMENTS
FOR THE DEGREE OF
DOCTOR OF PHILOSOPHY

By
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ABSTRACT

HF SIGNAL INJECTION AND SLIDING MODE ESTIMATOR BASED SENSORLESS CONTROL OF IPMSM

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Ph.D. in Electrical and Computer Engineering

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December 2024

In this study, we introduce a sliding mode observer (SMO) strategy based on high-frequency (HF) voltage injection integrated with a phase-locked loop (PLL) for the real-time estimation of the extended electromotive force (EEMF), rotor position, and rotor velocity in an Interior Permanent Magnet Synchronous Machine (IPMSM). This approach is specifically designed to address the limitations of traditional SMO and PLL techniques, which often struggle to provide reliable rotor angular position and velocity estimations at extremely low speeds and standstill. These conditions present significant challenges in IPMSM control because the BEMF signals crucial for detecting the position of the rotor tend to weaken or become indistinct, leading to inaccurate estimations when using conventional methods. To address these issues, the proposed SMO leverages HF voltage injection to stimulate the machine and produce distinguishable EEMF signals, which are less dependent on the motor's actual speed. This innovation allows for more consistent and noise-resistant EEMF estimation, forming a foundation for accurate rotor position and velocity estimation even at zero or near-zero speeds. The PLL then refines these EEMF estimates, enabling precise extraction of the rotor's velocity and position information. By maintaining a stable phase-lock with the EEMF signal, the PLL is capable of filtering out noise and enhancing the precision of the rotor position and velocity measurements. Building on this foundation, we developed and implemented a sensorless speed control system for IPMSM using the proposed SMO-PLL combination. The real-time system was tested across various operational circumstances, including low-speed and standstill states. Experimental results from an 8-pole, 0.4 kW IPMSM motor validate the superior efficiency and robustness of the suggested method compared to conventional SMO and PLL techniques.

Keywords: Sensorless Control, Signal Injection, IPMSM, Sliding Mode Observer, PLL

ÖZET

IPMSM'İN HF SİNYAL ENJEKSİYONU VE KAYAN KIPLİ GÖZLEMCİ TABANLI SENSÖRSÜZ KONTROLÜ

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Aralık-2024

Bu çalışmada, gömülü mıknatıslı senkron makinede (GMSM) genişletilmiş elektromotor kuvveti (GEMK), rotor pozisyonu ve rotor hızının gerçek zamanlı kestirimi için faz kilitli döngü (FKD) ile birleştirilmiş yüksek frekans (YF) gerilim sinyali enjeksiyonuna dayalı bir kayma modlu gözlemci (KMG) tanıtılmaktadır. Bu yaklaşım, özellikle düşük hızlarda ve durma anında rotor pozisyonu ve hız kestirimlerinde zorlanan geleneksel KMG ve FKD tekniklerinin sınırlamalarını ele almak üzere tasarlanmıştır. GMSM kontrolünde bu durumlar, rotor pozisyonu tespiti için kritik olan zıt EMK sinyallerinin zayıflama veya belirsiz hale gelme eğiliminde olması nedeniyle önemli zorluklar ortaya çıkarır, bu da geleneksel yöntemler kullanıldığında yanlış kestirimlere yol açar. Bu sorunları çözmek için önerilen KMG, makineyi uyararak motorun gerçek hızına daha az bağımlı olan belirgin GEMK sinyalleri üreten YF gerilim enjeksiyonundan yararlanır. Bu yenilik, sıfır veya sıfıra yakın hızlarda dahi tutarlı ve gürültüye dayanıklı GEMK kestirimine olanak tanıyarak rotor pozisyonu ve hızının doğru şekilde kestirilmesi için bir temel oluşturur. FKD, bu GEMK kestirimlerini rafine ederek rotorun hız ve pozisyon bilgilerinin hassas bir şekilde elde edilmesini sağlar. GEMK sinyaliyle stabil bir faz kilidini koruyarak, FKD gürültüyü filtreler ve rotor pozisyonu ve hız ölçümlerinin doğruluğunu artırır. Bu temel üzerine, önerilen KMG-FKD kombinasyonunu kullanarak GMSM için sensörsüz hız kontrol sistemi geliştirdik ve uyguladık. Gerçek zamanlı sistem, düşük hız ve durma durumları dahil olmak üzere geniş bir çalışma aralığında test edilmiştir. 8 kutuplu, 0,4 kW'lık bir GMSM motorundan elde edilen deneysel sonuçlar, önerilen yöntemin geleneksel KMG ve FKD tekniklerine kıyasla üstün verimlilik ve sağlamlığını doğrulamaktadır.

Anahtar kelimeler: Algılayıcı Kontrol, Sinyal Ekleme, GMSM, Kayma Modlu Gözlemci, FKD

Acknowledgments

During my PhD, I had the privilege of being supervised by Assoc. Prof. Burak Tekgün. His unwavering academic support and guidance were invaluable, yet he also entrusted me with the independence to make key decisions and take ownership of my research.

I am grateful to Dr. Bilal Akin for graciously welcoming me into his team at the Power Electronics and Drives Lab (PEDL) at the University of Texas at Dallas, TX, USA for ten months. Working with him was a pleasure and provided me with invaluable professional experience.

I would also like to express my gratitude to my co-advisor, Prof. Murat Barut and the committee members, Assoc. Prof. Rıdvan Demir and Assoc. Prof. Samet Güler, for their time, insightful feedback, and valuable contributions to my work.

I am deeply grateful to TÜBİTAK for generously funding my research at the University of Texas at Dallas, TX, USA, through the 2214-A International Research Fellowship Program for Ph.D. Students (grant number 1059B142201586). This support was instrumental in enabling my work.

I would like to express my heartfelt gratitude to all my colleagues, whose direct and indirect support and encouragement have been invaluable throughout this journey.

I would like to express my profound gratitude to my family, especially my parents, Sabri and Remziye, for their constant support and encouragement throughout my life.

Finally, my deepest thanks go to the stars of my life: my wife Elif, and our unborn child Cihangir. With you, everything becomes easier and more meaningful.

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LIST OF ABBREVIATIONS

AC	Alternative Current
AF	Active Flux
BEMF	Back Electromotive Force
DC	Direct Current
EEMF	Extended Electro-Motive Force
EV	Electric Vehicles
FOC	Field Oriented Control
FW	Flux Weakening
HF	High Frequency
IM	Induction Machine
IPMSM	Interior Permanent Magnet Synchronous Machine
LPF	Low-pass filter
MRAS	Model Reference Adaptive System
MTPA	Maximum Torque Per Ampere
PLL	Phase Locked Loop
PWM	Pulse Width Modulation
SMC	Sliding Mode Control
SMO	Sliding Mode Observer
SMPMSM	Surface Mounted Permanent Magnet Synchronous Machine
SVPWM	Space Vector Pulse Width Modulation
VC	Vector Control
VSC	Variable Structured Control



To my family

Chapter 1

Introduction

The rapid depletion of conventional energy sources and conflicts among nations over reserves are leading to energy crises. In today's world, where energy is extremely important, replacing non-renewable energy sources, which have limited reserves, with renewable and sustainable energy sources, which are inexhaustible, is crucial. For this reason, it is necessary to increase the use of energy sources such as solar, wind, biomass, and geothermal. These sources not only provide long-term energy solutions but also contribute to a cleaner environment by reducing dependency on fossil fuels, which are responsible for a variety of ecological and social challenges. In addition to wars between countries, the use of conventional energy sources brings about environmental issues, including ozone layer depletion, global warming, and acid rain. Increasing the use of renewable energy sources provides a global advantage in this regard, fostering energy security, environmental preservation, and economic resilience.

One of the sectors dominated by conventional energy sources is the transportation sector. The use of fossil fuels in transportation and the significant increase in the number of vehicles today lead to not only global problems but also issues such as noise pollution, lung diseases, and cancer. To address these problems, replacing fossil fuel-powered vehicles in transportation with electric vehicles (EVs) would increase renewable energy sources usage, prevent the emission of harmful greenhouse gases, and help reduce environmental and social problems. This is because the energy required for electric vehicles can be obtained through renewable energy sources. Moreover, widespread adoption of EVs could revolutionize urban planning by creating quieter, cleaner cities while reducing dependence on foreign oil and mitigating economic vulnerabilities tied to fluctuating fuel prices.

While the ability of electric vehicles to use renewable energy is advantageous, it is also crucial to use the obtained energy efficiently. The high efficiency of motors used in electric vehicles ensures the correct use of energy. For this purpose, permanent magnet synchronous machines (PMSM), which are more efficient and more suitable for vehicle

technology than other motors, are widely preferred in electric vehicles (EV). Compared to other motor types used in EV, PMSMs offer superior features like as high-power density, low noise operation, and high efficacy. Furthermore, PMSMs enable compact and lightweight designs, which are vital for improving vehicle performance and extending driving ranges. Their durability and reduced maintenance requirements also make them an economical choice in the long run, accelerating the shift towards sustainable transportation systems.

The integration of renewable energy technologies with electric vehicles not only paves the way for decarbonizing the transportation sector but also stimulates innovation in energy storage and grid management. Advanced battery technologies, such as lithium-ion, lithium-polymer, and solid-state batteries, complement PMSMs by providing reliable energy storage and enabling renewable energy systems to support electric vehicle networks. As this symbiotic relationship between clean energy and transportation evolves, it becomes clear that investing in renewable energy infrastructure, such as charging stations powered by solar or wind energy, will be a cornerstone in creating an eco-friendly, sustainable future.

1.1 Problem Statement and Hypothesis

The primary problem in sensorless control of Interior Permanent Magnet Synchronous Machines (IPMSM) arises from the inherent difficulty in precisely predicting the position of the rotor and velocity, particularly at a standstill or extremely low speeds, where the signal to noise ratio (SNR) of the sensed parameters is significantly reduced, making traditional estimation techniques less reliable and prone to errors due to factors such as magnetic saturation, inverter nonlinearities, and system disturbances. Traditional back EMF-based sensorless methods fail in these speed ranges due to their negligible EMF signals. This limitation creates significant challenges in applications requiring precise control under low-speed or zero-speed conditions, such as robotics, elevators, and electric vehicles. Although advanced methods like sliding mode observers provide robustness against noise and parameter variations, they struggle in low-speed scenarios and encounter challenges in adapting to changes in parameters such as resistance, inductance, and flux linkages due to factors like temperature variations, magnetic saturation, or aging effects in motor components. These variations lead to inaccuracies in estimation, reducing the reliability and efficiency of the control system.

Saliency tracking methods, which rely on signal injections, have been introduced to overcome some of these issues. By injecting high-frequency (HF) signals and analyzing the resulting system response, saliency-based techniques can predict rotor position even at low speeds. However, these methods come with trade-offs, including increased system complexity, higher implementation costs, and potential interference with the overall motor operation. Furthermore, the reliance on additional hardware components, such as specialized sensors or signal processing units, makes these methods less attractive for cost-sensitive or compact applications. Therefore, there is a pressing need for an innovative sensorless control strategy that provides accurate estimation of rotor position across the vast range of speed, eliminates mechanical sensors, and avoids the drawbacks of additional signal injection.

The hypothesis of this research is that a HF voltage injection-based sliding mode combined with a PLL estimator can provide accurate rotor position and speed estimation across the full range of speeds of IPMSMs, including a standstill and low speeds. By leveraging the innate advantages of HF voltage injection, this approach captures rotor saliency without the need for external signal injection techniques, ensuring simplicity and cost-effectiveness. The SMO offers robustness against noise and parameter variations, while the PLL estimator enhances accuracy by providing a synchronized phase reference for reliable estimation.

This combined method, when designed and implemented effectively, not only maintains the robustness of SMO but also addresses the limitations of traditional saliency tracking techniques. It simplifies the control architecture by reducing hardware requirements and ensures compatibility with existing motor drive systems. Additionally, the integration of HF voltage injection and PLL enhances adaptability to varying operating conditions, such as changes in load, temperature, or motor aging, making it a practical solution for a extensive range of automotive and industrial applications. Ultimately, this approach could set a new standard for sensorless control of IPMSMs, enabling precise and reliable operation across all speed ranges while minimizing complexity and cost.

1.2 Research Motivation and Contributions

The motivation behind this research stems from the growing demand for efficient and reliable sensorless control methods for IPMSMs in various industrial applications, such as wind turbines, electric vehicles, and manufacturing systems. These applications

often operate in environments where mechanical sensors are prone to failure due to harsh conditions, including vibration, temperature extremes, and contaminants like dust and moisture. The elimination of mechanical sensors not only enhances system reliability but also reduces the overall cost, complexity, and maintenance requirements of these systems. This is particularly crucial in cost-sensitive industries like automotive manufacturing, where reducing production expenses while maintaining high performance is a top priority.

Moreover, many industrial and automotive applications require precise control of motor operation at extremely low speeds or even standstill, such as during startup, positioning, or low-speed maneuvering. Traditional sensorless methods struggle in these scenarios, creating a technological gap that must be addressed to ensure seamless operation in real-world environments. The ability to maintain precise rotor position and velocity estimation across the entire range of speed, regardless of environmental conditions or motor parameter variations, is essential for maximizing energy efficiency and operational accuracy.

Developing a robust, computationally efficient, and cost-effective sensorless control method for IPMSMs has the potential to revolutionize the control of modern AC machines. Such advancements would not only improve industrial efficiency by enabling smoother and more reliable machine operation but also reduce downtime and associated maintenance costs. Furthermore, these improvements would directly contribute to energy savings by optimizing motor performance and reducing unnecessary power losses, aligning with the global push toward more sustainable and energy-efficient technologies.

Beyond immediate industrial benefits, the findings of this research could pave the way for broader adoption of IPMSMs in emerging fields like renewable energy systems, robotics, and autonomous vehicles. In renewable energy applications, for example, precise control of motors in wind turbines can enhance energy capture and reliability, while in robotics, the ability to operate smoothly at low speeds can improve dexterity and accuracy. Thus, this research has the potential to impact not only existing industries but also future technologies, underscoring its significance in advancing the field of motor control and enabling the next generation of smart, efficient, and resilient systems.

This study presents several key contributions to the field of sensorless control for IPMSMs:

- Development of a Novel Observer:** A high-frequency (HF) injection-based SMO combined with a PLL estimator is developed. This model-based observer is designed to

operate without additional signal injection algorithms, significantly reducing the system complexity and computational burden.

•**Innovative Use of HF Voltage Injection:** The study leverages HF voltage injection to enhance the precision of rotor position and velocity estimation, particularly in the challenging low-speed and standstill regions. Unlike conventional approaches, this method does not require complex signal processing algorithms, streamlining the design and improving computational efficiency.

•**Extension of Prior Work:** This research builds on existing studies, by introducing stability analysis, real-time implementation, and a performance comparison between the proposed HF voltage injection-based SMO and PLL estimator and traditional methods. The proposed method demonstrates improved robustness and accuracy, particularly during zero-speed reversals, making it a more reliable solution for sensorless control of IPMSMs.

•**Parameter Adaptation Mechanism:** Incorporating a Model Reference Adaptive System (MRAS) enables dynamic adaptation to real-time parameter changes, enhancing the robustness of sensorless control techniques in industrial applications where such variations are prevalent.

•**Impact on Industrial Applications:** The suggested sensorless control can be directly applied to various industrial systems by eliminating the need for mechanical sensors and additional signal injection algorithms. This reduces system costs, improves reliability, and simplifies maintenance in applications ranging from electric vehicles to renewable energy systems.

This thesis introduces a novel approach to the sensorless control of IPMSMs, addressing the key challenges of rotor position estimation at a standstill and very low velocity. Through developing and validating an HF voltage injection-based SMO and PLL estimator, this research provides a robust, accurate, and computationally efficient solution for sensorless control in a vast range of speeds. The proposed method offers a promising alternative to traditional back EMF-based and saliency tracking-based approaches, with significant potential for practical industrial applications.

Chapter 2

Literature Review

2.1 Introduction

IPMSMs have become increasingly prevalent in industrial applications due to their excellent efficiency, high power factor, dynamic performance, and high power density [1, 2]. These machines are widely utilized in high-performance control applications, like as robotics, aircraft systems, and renewable energy systems [3, 4]. Their ability to operate efficiently across a vast range of speeds and load conditions makes them ideal for applications demanding precision, energy savings, and reliability. Additionally, IPMSMs' compact design and high torque density allow for their integration into space-constrained systems, further increasing their appeal across various industries.

One of these most effective methods for controlling IPMSMs is vector (or field-oriented) control, which achieves high dynamic performance by aligning rotor angle and stator excitation. This control strategy ensures optimal torque production and efficient energy utilization, making it a key factor in realizing the full potential of IPMSMs. However, this method relies heavily on precise rotor position feedback, traditionally provided by mechanical position transducers such as resolvers or encoders [5]. These sensors, while accurate, introduce several drawbacks. They increase the system's cost and complexity due to the additional components and wiring required. Moreover, they are prone to failure in harsh industrial environments, where factors such as vibration, dust, moisture, and temperature extremes can compromise their performance.

To overcome these challenges, sensorless control methods have been developed, which estimate the position of the rotor and speed without needing mechanical position sensors. Sensorless approaches also simplify system integration and improve robustness in environments where traditional sensors would be unreliable [6]. However, achieving

accurate and reliable rotor position estimation without mechanical sensors, especially at low speeds or during standstill, remains a significant technical challenge. Advancements in sensorless control methods are therefore crucial to fully unlocking the benefits of IPMSMs in modern industrial and automotive applications, ensuring their continued adoption and success in an increasingly competitive and demanding technological landscape. In the literature, sensorless estimation methods for IPMSMs are generally divided into two main categories: (i) model-based techniques and (ii) saliency tracking-based techniques.

2.2 Sensorless Control Techniques Based on Models

These methods use the inherent mathematical frames of the IPMSM to predict the position of the rotor and velocity. These strategies are either open-loop estimators, which rely on back electromotive force (BEMF), extended electromotive force (EEMF), or active flux (AF) estimation [3], [7], [8], or closed-loop observers, which use techniques such as the linear state observers, the MRAS [9], disturbance observers [10], flux observers [11, 12], or SMOs [13–17].

Open-loop estimators, particularly those based on BEMF, are effective at medium to high speeds but struggle at low speeds (below 10% of the rated speed) and zero speed due to the diminishing back EMF signal [4, 18, 19]. This limitation makes open-loop methods unsuitable for applications requiring precision during startup, positioning, or low-speed operation. In such cases, noise, parameter variations, and system uncertainties further exacerbate inaccuracies in rotor position and speed estimation. While computationally simple and cost-effective, open-loop estimators often require additional mechanisms, such as signal injection or parameter compensation, to extend their usability to low-speed conditions.

Closed-loop observers, such as SMO, are more resilient to noise and parameter variations, making them more suitable for sensorless control in a broader speed range. SMOs leverage discontinuous control laws to achieve robustness against system nonlinearities and disturbances, ensuring better estimation performance under varying load and environmental conditions. However, traditional SMO-based methods still face challenges at very low or zero speeds, where the lack of sufficient saliency information and parameter mismatches can lead to estimation errors. Additionally, SMOs can

introduce chattering effects, which, if not mitigated, may lead to instability or reduced system efficiency.

To address these challenges, hybrid approaches are being explored that integrate the advantages of multiple techniques. For instance, SMOs integrated with high-frequency (HF) voltage injection techniques can improve low-speed estimation by leveraging rotor saliency. Similarly, adaptive SMOs that incorporate parameter estimation algorithms can dynamically compensate for variations in resistance, inductance, or magnetic flux due to temperature changes or magnetic saturation. These hybrid strategies aim to maintain the robustness of closed-loop observers while extending their effectiveness to the full-speed range, including zero-speed and transient conditions.

Furthermore, advancements in digital signal processing (DSP) and machine learning are opening new avenues for sensorless control. By embedding data-driven models into traditional estimation frameworks, it becomes possible to enhance accuracy and adaptability, even in the face of significant system uncertainties. These developments highlight the ongoing evolution of sensorless control techniques for IPMSMs, underscoring the importance of achieving reliable, efficient, and cost-effective solutions for modern industrial and automotive applications.

2.2.1 Flux or EMF Estimation

As stated in the paragraph above, EMF and flux can be estimated using open-loop and closed-loop techniques in model-based sensorless control. Closed-loop estimation is generally preferred because of its superior robustness and precision. The closed-loop approach enhances the system's ability to adapt dynamically to changes in motor parameters, such as resistance fluctuations caused by temperature variations or magnetic saturation effects, ensuring consistent and reliable performance across a wide range of operational circumstances. Accurate prediction of EMF or flux is critical for reliable position and velocity estimation, as any inaccuracies in these estimates can propagate through the control system, resulting in degraded performance or instability. Factors influencing estimation accuracy include the reference frame implementation, which determines how system dynamics are represented in either stationary, rotor, or synchronous reference frames, and the underlying mathematical model of the motor, which must account for the machine's non-linearities, saliency, and parameter variations.

Additionally, the choice of error convergence method plays a significant role in determining how quickly and precisely the estimation process aligns with the true motor states. Techniques such as PI controllers, SMC, or adaptive algorithms are commonly employed to achieve fast and stable error convergence [20]–[23].

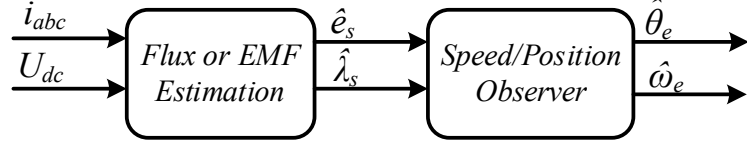


Figure 2.1 Overview speed/position estimation block diagram for model-based sensorless control

Sensorless control using a model-based approach can operate either in the synchronous (dq) axis or the stationary ($\alpha\beta$) axis. In the dq frame, it is essential to align an estimated frame with the actual one. Accurately estimating EMF or flux involves minimizing the alignment error between these frames, which is crucial for achieving precise control and ensuring the desired torque production and efficiency. Proper alignment facilitates seamless motor operation across various speed ranges, enabling the system to adapt to dynamic operating conditions while maintaining stability.

In salient PMSMs, position data is derived from both the BEMF and the variations in the stator inductance resulting from rotor saliency. This unique characteristic restricts the direct application of traditional model-based sensorless methods, as it introduces complexities in the mathematical representation of the motor. To address this limitation, techniques like the extended EMF approach [24, 25] and the active flux model [12, 26, 27] have been introduced. These methods exploit the innate saliency of the machine to enhance the precision of the estimation of the rotor position, particularly in scenarios where traditional back EMF-based methods fall short, such as low-speed operation.

In closed-loop estimation methods, a control law is typically designed to drive the error between measured and estimated outputs to zero, ensuring the observer's convergence to the true motor states. Various techniques, such as linear state observers [28–31] and sliding mode observers [25, 27, 32], can be employed to enhance the accuracy of the estimation process. Linear state observers provide a straightforward implementation with moderate robustness, while sliding mode observers excel in handling noise, parameter variations, and system non-linearities, albeit at the cost of potential chattering effects that must be mitigated through advanced control strategies.

EMF and flux estimation methods are now fairly advanced, with significant improvements in computational efficiency and accuracy. However, recent research has shifted its focus toward addressing specific challenges, such as improving dynamic performance, enhancing robustness to parameter variations, and ensuring reliable operation at low frequencies. Factors like inverter nonlinearities, magnetic saturation, and flux spatial harmonics continue to pose challenges, particularly at low frequency ratios where the back EMF signal is negligible. Compensating for these factors is essential to maintain stability and reliability in sensorless control systems.

The estimated position can be determined from the estimated flux, which is obtained by integrating the flux model in relation to stator current and voltage [33, 34]. However, practical implementation requires careful consideration of integration drift and errors caused by noise and parameter mismatches. Advanced signal processing techniques, such as digital filtering and adaptive integration, are often employed to mitigate these issues. Furthermore, hybrid estimation approaches that combine flux-based methods with saliency tracking or high-frequency signal injection are gaining traction, as they offer improved performance across a vast range of speeds, including a standstill and very low speeds.

As the demand for high-performance, sensorless IPMSM control grows in applications like as renewable energy systems, EVs, and industrial automation, ongoing advancements in flux and EMF estimation methods will remain critical. By addressing challenges such as low-speed operation, parameter sensitivity, and environmental disturbances, these innovations will contribute to the development of robust, efficient, and scalable sensorless control systems that meet the demands of modern industrial and automotive applications.

2.2.2 Position/Speed Observer

When the observer error reaches zero, it enables accurate estimation of the EMF or flux, which facilitates the calculation of position and velocity. The position of the rotor can be directly detected utilizing the arctan function. To enhance estimation accuracy, a speed/position observer is typically preferred. This observer uses the error signal of the position, which can be derived from vector cross-products or basic arithmetic, to iteratively refine its estimates, leveraging the discrepancy between the measured and estimated positions to adjust and align the internal model with the actual system dynamics for improved accuracy. Subsequently, a PI-type PLL or a PID Luenberger estimator can

be employed to decrease the error extracted from the position and obtain the required position and speed information [17, 35–37]. While the PLL is effective, the PID-based Luenberger estimator offers better performance, through it requires knowledge of the moment of inertia.

2.2.2.1 Kalman Filter and Extended Kalman Filter

It is noteworthy that the Kalman Filter and its nonlinear counterpart, the Extended Kalman Filter(EKF) [38–40], are effectively utilized as high-gain observers due to their ability to handle dynamic state prediction in the presence of noise and uncertainty. These filters leverage a recursive estimation framework, combining predictions from system models with real-time observations to refine estimates, making them invaluable in applications requiring robust and accurate state tracking. Additionally, many researchers have relied on a “dirty” derivative approach for speed estimation, where the position is estimated position, which involves approximating the derivative of the position. This approximation is essential due to the non-casual nature of the time derivative operator, which limits direct computation in real-time systems. While this method provides a pragmatic solution for scenarios where computational simplicity and rapid response are critical, it may introduce noise amplification, requiring careful tuning. Such estimators can also be categorized as high-gain observers, with the prediction term being zero.

2.2.2.2 Nonlinear and Linear Luenberger-Like Estimators

By the late 1980s, numerous high-gain estimators had been designed leveraging Taylor linearization, enabling the effective approximation of nonlinear systems and advancing state estimation techniques for a wide range of control applications [19, 41, 42]. However, this method is commonly acknowledged to have shortcomings in delivering effective performance throughout a wide range of state space. Consequently, the 1990s brought about the emergence of nonlinear techniques, nonlinear Luenberger-like observers and SMOs . A detailed exploration of various methods can be found in the references provided in [43, 44].

Nonlinear Luenberger-based estimators are proposed in [45–47]. In this approach, the estimation term is formulated based on the dynamic equations of the EMF, along with a correction term that features a nonlinear gain. Assuming the model is accurate, meaning the parameters used in the prediction process correspond to those of the machine and there is no ambiguity or noise affecting the measured variables, the prediction error will asymptotically converge to zero. This occurs when the estimator gain is carefully

designed following the methodologies outlined in these papers. A significant benefit of this approach is its ability to ensure convergence across a large area of the state space, with the rate of convergence being adjustable as required.

2.2.2.3 Sliding Mode Observer

Sliding Mode Observers (SMOs) are constructed on a well-recognized variable structure design, employing a discontinuous gain for the correction term, typically represented by a sign function. This approach enables the estimation error to converge to a specific surface, after which it slides along that surface to reach zero within a finite timeframe. SMOs have proven effective in predicting the angular position of the rotor and speed of PMSMs, as referenced in [48–52].

Nevertheless, the use of discontinuous functions like the sign function can introduce undesirable chattering effects. To mitigate this issue, several modifications have been suggested, such as substituting the sign function with a sigmoid function [13, 53] or using a LPF to reduce HF noise [48]. In certain methods, EMF estimates rely on an approximated estimation term that considers constant acceleration, which adequately describes scenarios at a steady speed but fails to account for periods of acceleration or deceleration.

Reference [50] introduces an extended SMO intended to directly predict lumped uncertainties, compensating for severe disruptions and achieving high servo precision; however, it is important to note that this technique estimates speed while position is quantified using a transducer. In another sensor-based method, parameters are predicted with an SMO. Reference [52] suggests improving observer performance by replacing the LPF with a mechanism designed to extract the essential components of EMFs, such as a PLL, to effectively eliminate harmonic components from EMF estimates.

2.3 Saliency Tracking-Based Estimation Methods

These techniques rely on the magnetic anisotropy of the rotor to predict the position of the rotor. These methods can track the rotor's magnetic saliency by inserting a HF signal into the stator windings and analyzing the response. Techniques such as rotating or pulsating HF signal injection and PWM excitation have proven effective at very low speeds and standstill [54–57]. These methods offer robust performance in low-speed ranges but require more complex control algorithms and specialized hardware, adding to system complexity and computational burden. Furthermore, to achieve full-speed range

sensorless control, saliency tracking methods must be used in combination with model-based observers.

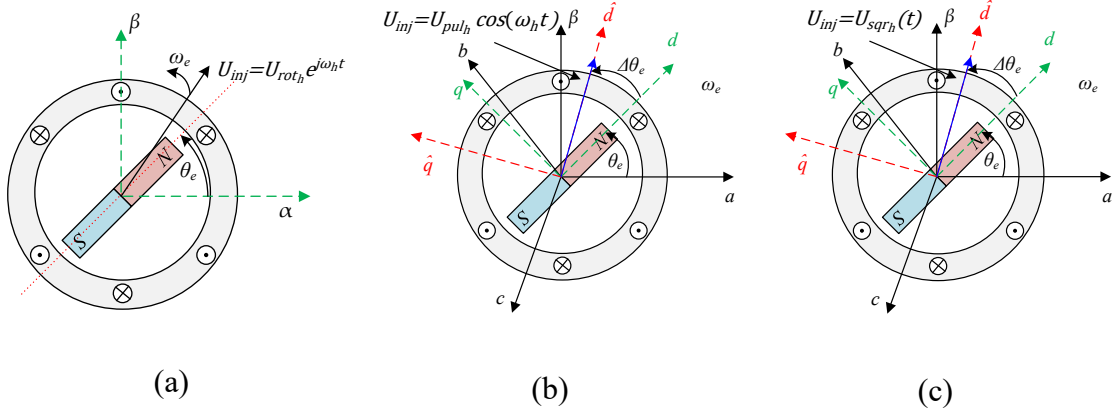


Figure 2.2 Analogy of Different HF Injection Schemes (a) Rotating signal injection, (b) Pulsating sinusoidal signal injection, (c) pulsating square-wave signal injection

As mentioned in the above paragraph, HF signal injection methods, based on the selected injection reference frame, can be generally grouped into pulsating and rotating signal injection techniques. Pulsating injections can be further categorized into sinusoidal and square-wave injection schemes. Figure 2.2 provides a comparative summary of the different HF signal injection methods.

2.3.1 Rotating Signal Injection

Figure 2.2(a) illustrates the block diagram of a sensorless control scheme based on rotating injection [55–57]. An HF rotating voltage space vector is inserted into the $\alpha\beta$ axes. To maintain the quality of the inserted sinusoidal voltage signal, the injection frequency should not be set too high, typically kept below 10% of the PWM frequency. Using the $\alpha\beta$ frame voltage equations, two common rotating signal injection methods, the zero sequence voltage and the negative sequence current technique can be used.

2.3.2 Pulsating Signal Injection

Pulsating injection is another effective approach for sensorless control based on additional signal injection. Since high-frequency signals are injected in the rotor frame, the initial position of the rotor is necessary for this method. Pulsating signal injection can

be categorized into two main types based on the nature of the injected signals: pulsating sinusoidal injection [57–61] and pulsating square-wave injection [62–67].

In **pulsating sinusoidal injection**, a HF sinusoidal voltage is inserted into the motor's rotor reference frame along a specific axis, typically aligned with d-q axis. This HF voltage signal induces the corresponding HF currents in the stator windings, which interact with the motor's magnetic properties. Because the rotor is an anisotropic, the current response to this injected HF voltage depends on the rotor position. This position dependency means that the response current carries information about the rotor's angular position. According to the HF current response, the position is encrypted in the envelope of the HF current within the $\alpha\beta$ axis. By selecting a demodulation signal [58], the HF response currents can then be demodulated using an LPF to extract the position-dependent information effectively.

The **pulsating square-wave injection**-based diagram shares a similar structure with the pulsating sinusoidal approach; it uses a square-wave signal for the injection voltage. this adjustment enables a higher injection frequency compared to both the pulsating and rotating sinusoidal injection schemes, enhancing the extraction and separation of HF signals. PWM frequency injections are synchronized with current sampling, with PWM updates occurring twice per switching period. When the PWM switching frequency is near the audible range, acoustic noise from the injected signal can be significantly minimized, effectively reducing noise impact [68].

2.4 Conclusion

This thesis introduces a novel approach to the sensorless control of interior permanent magnet synchronous motors (IPMSMs), addressing the key challenges of the prediction of the rotor position at a standstill and extremely low-speed conditions under which traditional methods often face significant limitations. Through the development and rigorous validation of a HF injection-based SMO combined with a PLL estimator, this research offers a robust, accurate, and computationally efficient solution capable of delivering reliable full-speed range sensorless control. The integration of HF signal injection enhances rotor position identification at low speeds, where BEMF methods struggle, while the SMO and PLL combination ensures noise resilience and stability across the speed range. The proposed method, by focusing on computational efficiency and ease of implementation, offers a promising alternative to conventional back-EMF-

based and saliency tracking-based approaches, which are often complex and computationally intensive. With its adaptability for practical industrial applications, this technique opens new avenues for IPMSM sensorless control systems, potentially reducing costs and enhancing reliability in applications such as robotics, and renewable energy systems.



Chapter 3

Basic Structure and Operating Principles of Interior Permanent Magnet Synchronous Machines (IPMSMs)

3.1 Introduction

Electric motors come in various types, each tailored to specific applications and operating conditions, ensuring their adaptability across a broad spectrum of industries. Based on their power supply, they are primarily categorized into DC motors and AC motors. DC motors, known for their simplicity and ease of speed control, are often employed in applications requiring precise adjustments, such as robotics and electric vehicles. AC motors, on the other hand, are widely used due to their efficiency and robustness and are further divided into synchronous motors and induction motors (IM). Synchronous motors operate at a constant speed determined by the supply frequency, making them ideal for applications like pumps and compressors. Induction motors, distinguished by their self-starting nature and rugged design, dominate industries requiring heavy-duty machinery such as conveyors and cranes.

By structure and operating principles, electric motors can also be classified into induction or asynchronous motors, electrically excited synchronous motors, permanent magnet motors, and specialized motor types, each offering distinct advantages. For instance, permanent magnet motors are valued for their high efficacy and compact size, making them a preferred choice for applications such as electric vehicles and drones. Additionally, specialized motors, including stepper motors and servo motors, are designed for specific tasks, such as precise positioning and motion control in CNC machines and automation systems. Further classifications arise based on parameters like volume (compact or large-scale), phase (single-phase or three-phase), speed range (fixed-

speed or variable-speed), control type (open-loop or closed-loop), and excitation method (self-excited or separately excited). These diverse classifications underscore the remarkable versatility of electric motors, enabling them to meet the unique demands of various industries and technological advancements. As innovation continues to drive the development of smart, energy-efficient motor technologies, their role in defining the future of transportation, manufacturing, and sustainable energy systems becomes increasingly evident.

3.2 Permanent Magnet Synchronous Motor

PMSMs encompass a diverse array of designs to meet varying application needs. Broadly, they are categorized into two primary types based on the placement of the PMs on the rotor: Surface mounted permanent magnet synchronous motors (SMPMSM) and interior permanent magnet synchronous motors (IPMSM). This classification plays a critical role in determining the motor's performance characteristics, including efficiency, torque capability, and operational stability.

In SMPMSMs, the PMs are securely affixed to the rotor surface, as depicted in Figure 3.1. This configuration allows for simple construction and facilitates a uniform magnetic flux distribution in the air gap, leading to high torque accuracy and low cogging torque. Due to their straightforward design and ease of manufacturing, SMPMSMs are commonly employed in applications requiring precision and consistent speed, such as robotics, medical devices, and low-power industrial equipment. However, the exposed position of the magnets can limit their thermal performance and mechanical robustness in high-stress environments.

Conversely, IPMSMs embed the permanent magnets within the rotor structure, providing enhanced mechanical protection and higher field-weakening capability. This design makes IPMSMs better suited for applications that demand high efficacy over a vast range of speeds, such as automotive traction systems and advanced industrial drives.

The choice between SMPMSM and IPMSMs is influenced by the specific requirements of the application, such as operating conditions, torque demands, and efficiency needs. As technology advances, improvements in magnet materials and rotor designs continue to expand the capabilities of both SMPMSM and IPMSMs, reinforcing their significance in modern engineering and energy-efficient systems.

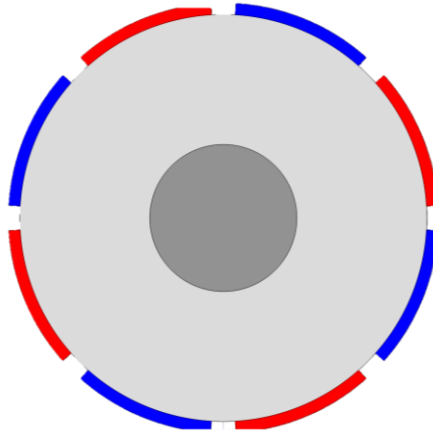


Figure 3. 1 Surface-Mounted PMSM (SMPM) rotor

In IPMSMs, the PMs are strategically embedded within the rotor core, offering significant advantages in terms of mechanical robustness, thermal stability, and operational efficiency. This design ensures that the magnets are well-protected from external forces, such as centrifugal stress and thermal demagnetization, which are common in high-speed and high-torque environments. The structure of an IPMSM rotor, as depicted in Figure 3.2, reflects a thoughtful balance between durability and performance, with the magnets positioned to maximize torque production and enable efficient energy conversion.

IPMSMs are also known by various names, including "embedded," "internal," "buried," or "built-in" rotor PMSMs. These terms highlight the key design feature of incorporating the magnets inside the rotor, as opposed to being mounted on its surface. This embedded configuration enhances the motor's ability to handle higher speeds and provides improved field-weakening capabilities, which is critical for extending the operational range in applications such as industrial drives, wind turbines, and electric vehicles.

Additionally, the embedded magnet design improves the saliency ratio, enabling efficient production of both torque components: magnet torque and reluctance torque. This dual-torque generation capability makes IPMSMs particularly effective in applications requiring high dynamic performance and energy efficiency. The advancements in magnetic materials and rotor engineering continue to optimize the design of IPM motors, further cementing their role as a preferred choice for demanding and energy-sensitive applications.

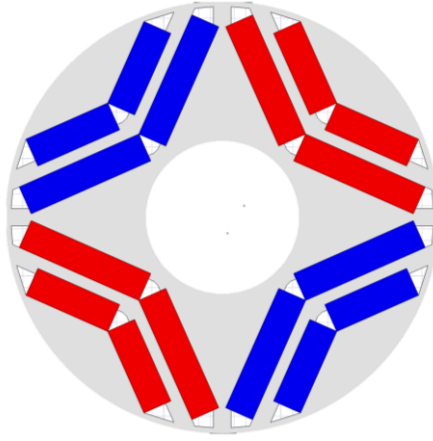


Figure 3.2 Rotor design featuring embedded, radially polarized magnets in an IPMSMs.

The rotor axis aligned with the field flux generated by the permanent magnets is referred to as the flux or direct axis (d-axis). Perpendicular to this, the axis responsible for producing torque is known as the torque or quadrature axis (q-axis). These axes are fundamental to understanding the electromagnetic interaction within the rotor and play a crucial role in defining the motor's performance characteristics.

In wound rotor synchronous machines with salient pole, the rotor is not uniformly cylindrical, resulting in a significant air gap along the q-axis. This design causes the q-axis to have higher magnetic reluctance compared to the d-axis. Consequently, the d-axis reactance typically measures around 1 per unit (p.u), whereas the reactance of the q-axis is noticeably lower.

In contrast, the situation is reversed in many IPMSMs. Here, the q-axis reluctance is lower than the d-axis due to the presence of PM material along the d-axis flux path. Permanent magnets, characterized by their low magnetic permeability, create a higher reluctance path for the d-axis flux. Meanwhile, the q-axis flux path is primarily composed of magnetic material with high permeability, resulting in lower reluctance. This disparity leads to a larger q-axis inductance (L_q) compared to the d-axis inductance (L_d), a characteristic that influences the performance and control strategies of IPMSMs ($L_q > L_d$).

In IPMSMs, the air gap is comparatively small, comparable to that of an asynchronous machine, resulting in both the d-axis and q-axis reactances being significant relative to the leakage inductance of stator winding. The notable saliency and substantial reactances - often reaching up to 0.9 p.u. - categorize the IPMSM as a salient pole machine. This saliency characteristic is key to the motor's operation, as it contributes to the generation of a reluctance torque along with the primary PM torque. The combined

effect of these torque components enhances the motor's efficiency and performance, particularly in applications requiring high dynamic response and energy optimization.

SMPMSMs feature PM material covering the rotor's whole surface, resulting in symmetrical magnetic reluctance along both the d-axis and q-axis. This symmetry classifies SMPM motors as non-salient pole synchronous motors. Because of the low permeability of the PM material, the d-axis and q-axis reactances in SMPMSMs are typically small and equal ($L_q = L_d$), significantly lower than those observed in IPMSMs.

In contrast, IPM motors exhibit higher reactance values, a characteristic that enables them to support field-weakening operations. This capability is critical for applications requiring extended speed ranges and enhanced control flexibility, making IPM motors particularly suitable for high-performance applications like electric vehicles and industrial drives. The distinctions between SMPM and IPM motor reactance and field-weakening capability will be explored further in subsequent sections.

3.2.1 IPMSM Characteristics

The IPMSM has been designed to address the stringent performance demands and efficiency of industrial systems and advanced motion control applications. Its key attributes include a compact design that delivers high torque and power density, making it ideal for space-constrained environments. The motor also boasts exceptional energy efficiency, a robust and durable structure capable of withstanding high mechanical and thermal stresses, and a unique field-weakening capability that enables extended operational speed ranges. These features make the IPM motor a preferred choice for cutting-edge applications such as robotics, electric vehicles, and high-performance industrial automation.

In controllable-speed drive applications such as fans, air conditioners, pumps, and spindles, asynchronous machines have traditionally been the dominant choice due to their cheap cost, robust construction, and ability to operate directly from main power. However, in most modern drive systems, induction motors are paired with AC machine drives to achieve high speed and torque performance along with improved efficiency. With the widespread use of AC motor drives, there has been growing interest in replacing induction motors with IPM motors to further enhance the efficiency and performance of drive systems.

Although IPM motors are more expensive than induction motors, they are increasingly favored in adjustable-speed drive applications because of their superior efficacy, higher performance, and minimal maintenance requirements. Moreover, for high-power applications where Surface-Mounted Permanent Magnet Synchronous motors (SMPMSM) fall short in meeting power demands, IPM motors offer a compelling solution. Their high-power density, robust performance, and ability to support field-weakening operations make them ideal for demanding applications, ensuring that they remain a preferred choice in industries requiring precision, reliability, and energy efficiency.

The advantageous characteristics of IPM motors can be summarized as follows:

-Enhanced Flux Density and Torque Generation: The embedded PM structure enhances flux density and distribution, improving both torque and power density. Additionally, this design introduces saliency, which generates a reluctance torque component. The combined effects result in superior efficiency and performance compared to other motor types.

-Field-Weakening Capability: The embedded magnet design allows for a shorter effective air gap owing to the presence of core material on the rotor surface. This short air gap increases synchronous reactance, enabling the armature MMF to generate sufficient flux for field weakening. Unlike SMPM motors, which have a longer air gap and negligible synchronous reactance, IPM motors can perform effective field weakening. This capability allows IPM motors to maintain constant power across a vast range of speeds, extending the operational speed range critical for many industrial applications, similar to the functionality of induction motors.

-Robustness at High Speeds: The embedded magnet configuration in IPM motors protects the magnets from centrifugal forces at high speeds, unlike the surface-mounted magnets in SMPM motors, which are prone to mechanical stress. This results in a robust and balanced rotor, enabling higher-speed operation and improved durability.

-Sensorless Operation: The pronounced saliency characteristic of IPMSMs makes the sensorless control simpler, as utilized in this study. This feature makes IPM motors particularly suitable for applications requiring reduced complexity and cost associated with sensorless systems.

-Reduced Cogging Torque: Optimized flux distribution minimizes cogging torque, which is crucial for precision applications requiring smooth and accurate motion control.

-Compact Design and Low Inertia: The interior PM configuration allows for a rotor with lower inertia, making IPM motors ideal for high-dynamic-response applications. This design also contributes to reducing the overall size of the motor while maintaining performance.

-Hybrid Functionality in Some Designs: In certain IPMSMs, cage segments or conducting bars (typically made of aluminum or copper) are integrated into the rotor alongside the PMs. These conducting elements act as a cage, similar to that of an asynchronous motor, enabling asynchronous activity for direct starting. This design allows the motor to function as an induction-start, synchronous-run PMSM, combining the benefits of both motor types for added versatility.

-In summary, IPMSMs are the preferred choice for adjustable drive applications, offering a high-performance alternative to asynchronous motors and serving as the leading solution for high-power drive systems. Renowned for their efficiency, precision, and robust design, IPMSMs are extensively used across a diverse range of applications, including air conditioning systems, pumps, fans, cranes, elevators, escalators, ship propulsion systems, traction drives, and hybrid and electric vehicles. Their adaptability is underscored by their wide power range, from a few watts for precision control applications to hundreds of kilowatts for heavy-duty industrial and transportation needs. This broad applicability, combined with their ability to meet stringent demands for energy efficiency, reliability, and advanced performance, cements IPM motors as a cornerstone technology in modern electrical and mechanical systems.

3.3 Steady-State Analysis of PMSMs

To build a foundational understanding of the steady-state behavior of PMSMs, we begin with the SMPMSM. This type of motor, characterized by its non-salient pole construction, is simpler to analyze compared to the salient-pole IPMSM. The steady-state behavior of the SMPMSM can be symbolized using the single-phase equivalent circuit given in Figure 3.3(a). For a clearer insight into the motor's operation, the steady-state phasor scheme is provided in Figure 3.3(b). In this diagram, the stator resistance is neglected for simplicity, focusing on the primary voltage, current, and back EMF relationships under steady-state conditions. This analysis serves as a stepping stone to understanding more complex PMSM configurations.

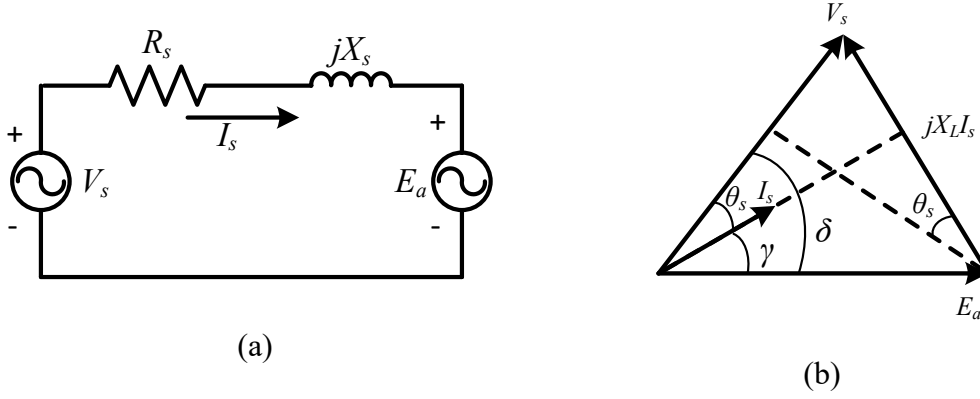


Figure 3.3 (a) Equivalent circuit and (b) phasor diagram of the surface-mounted PMSM.

Where V_s represents the machine's terminal voltage, and E_a corresponds to the BEMF, generated by the excitation of the rotor's PMs. The value of E_a is directly proportional to both the rotor velocity and the flux produced by the PMs, mirroring the behavior of back EMF in a DC motor. This proportional relationship underscores the critical role of rotor speed and magnetic flux in determining the back EMF magnitude.

Wound-field synchronous motors are well-established for their application as synchronous generators. As a result, the torque/power equations and steady-state analysis for synchronous motors are often presented in terms of voltage equations in textbooks and technical literature. This approach reflects their operation, which typically involves maintaining or generating a constant voltage under steady-state conditions. The torque/power output equation, derived from the machine's input power, is expressed below, with the stator resistance (R_s) omitted for simplicity.

$$P_s = 3V_s I_s \cos \theta_s \quad (3.1)$$

$$I_s X_s \cos \theta_s = E_a \sin \delta \quad (3.2)$$

According to these two equations, the output power can be obtained as below:

$$P_s = 3V_s I_s \cos \theta_s = 3 \frac{V_s E_a \sin \delta}{X_s} \quad (3.3)$$

The angle δ , commonly referred to as the load angle or torque angle, represents the phase difference between the back EMF and stator voltage. Maximum power is achieved when $\delta = 90^\circ$. The torque created by the motor can be determined using the mechanical power equation:

$$P_{mech} = T_e \times \omega_r \quad (3.4)$$

When the rotational losses and stator resistance are neglected, the mechanical power output of the machine becomes equal to its input electrical power. Therefore:

$$T_e = \frac{P_s}{\omega_r} \quad (3.5)$$

Where ω_r is the mechanical rotor angular velocity. In synchronous motors, the rotational speed is intrinsically related to the electrical angular speed (excitation frequency) through the number of pole pairs:

$$\omega_r = \frac{\omega_e}{(P/2)} \quad (3.6)$$

From these equations, the electromechanical torque can be computed as:

$$T_e = 3 \frac{P V_s E_a \sin \delta}{2 \omega_e X_s} \quad (3.7)$$

The power and torque equations above are expressed in terms of the load angle between the BEMF voltage and terminal voltage. This formulation is particularly suited for applications where synchronous motors or generators operate with a constant voltage supply. Wound-field synchronous motors are typically utilized in high-power operations. In contrast, PMSMs are commonly employed in modern motion control systems. These machines are powered by AC machine drives, as discussed in this study. Modern AC drives regulate the motor current through a voltage source inverter (VSI), that provides variable magnitude voltage and variable frequency to the machine. Consequently, analyzing the steady-state behavior of a surface-mounted PMSM (SMPM) is more effectively done using torque and current based power equations, as shown below. The power transferred to the machine's shaft can then be determined from the internal power.

$$P_s = 3 E_a I_s \cos \gamma \quad (3.8)$$

Likewise, the generated electromechanical torque can be obtained using the expression below.

$$T_e = 3 \frac{P E_a I_s \cos \gamma}{2 \omega_e} \quad (3.9)$$

The angle γ between the BEMF E_a and the current is referred to as the power factor (PF) angle. Maximum torque or power output is achieved when $\gamma = 0$. The BEMF voltage E_a is directly proportional to the rotor velocity and the field flux is mathematically represented as:

$$E_a = \omega_e \lambda_{pm} \quad (3.10)$$

Incorporating E_a from the preceding equation in the torque expression, the resulting torque for the surface-mounted PMSM is obtained as:

$$T_e = \frac{3 P}{2} \lambda_{pm} I_s \cos \gamma \quad (3.11)$$

The torque in (3.11) clearly shows that the torque of the surface-mounted PMSM can be controlled by varying the magnitude of the stator current and the angle γ between

the stator current and BEMF voltage. As derived from Faraday's law, the BEMF voltage, induced by the rotor's PM flux, is phase-shifted by 90 degrees ahead of the PM flux.

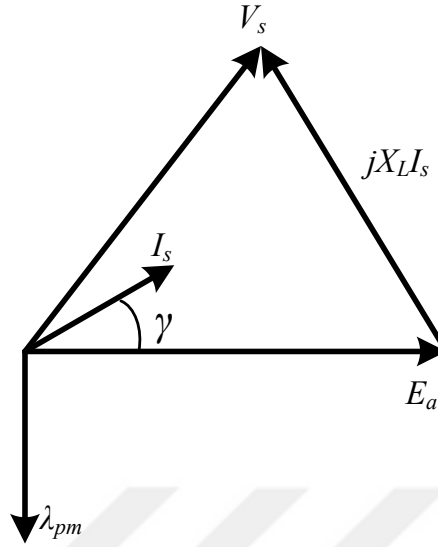


Figure 3.4 Phasor diagram of a SMPMSM and the rotor PM flux.

As depicted in Figure 3.4, the rotor flux is lagging the BEMF voltage by 90 degrees, enabling the control of the angle γ through rotor position feedback for precise current commutation. This feedback, necessary for regulating the angle γ between the stator current and BEMF can be obtained either with a position transducer mounted on the motor shaft or via sensorless control techniques that eliminate the need for a mechanical transducer. To achieve MTPA in the surface-mounted PMSM, the angle γ must be maintained at 0 degrees. When the current is controlled to ensure $\gamma = 0^\circ$, the motor delivers maximum torque, making torque directly in proportion with the stator current, akin to the operation of a DC machine. This approach allows the surface-mounted PMSM to function as a DC machine through current control facilitated by vector coordinate transforms.

For the interior PMSM, it is impossible to model using a steady-state single-phase equivalent circuit, because of its salient pole design and lack of spatial symmetry. Its behavior and performance are more accurately captured using phasor diagrams.

The magnetic flux direction is known as the direct (d -) axis, while the direction perpendicular to and leading to the direct axis is termed the quadrature (q -) axis. The BEMF, which leads to the PM flux by 90 degrees, is in alignment with the q -axis. The stator current can be disintegrated into components along the d - and q axes, following the d - q axis framework. In an IPMSM, the reactances related to the d - and q -axis current

Figure 3.5 Phasor diagram of the Interior PMSM.

The torque of the IPMSM in steady-state can be calculated, as in the case of the surface-mounted PMSM, with the division of the input power by the rotor mechanical velocity, stator resistance losses are negligible as shown in (3.1).

The torque of the IPMSM in steady-state can be calculated, as in the case of the surface-mounted PMSM, with the division of the input power by the rotor mechanical velocity, stator resistance losses are negligible as shown in (3.1).

$$P_s = 3V_s(I_q \cos \delta - I_d \sin \delta) \quad (3.11)$$

$$V_s \cos \delta = E_a + X_d I_d \quad (2.12)$$

$$V_s \sin \delta = X_q I_q \quad (5.12)$$

$$P_s = 3 \frac{V_s E_a}{X_{ds}} \sin \delta + \frac{3}{2} V_s^2 \frac{(X_d - X_q)}{X_d X_q} \sin 2\delta \quad (3.13)$$

$$T_e = \frac{P_s}{\omega_r} \quad (3.14)$$

$$T_e = \frac{3}{\omega_r} \frac{V_s E_a}{X_{ds}} \sin \delta + \frac{3}{\omega_r} \frac{1}{2} V_s^2 \frac{(X_d - X_q)}{X_d X_q} \sin 2\delta = T_{reaction} + T_{reluctance} \quad (3.15)$$

26

Expressing the torque or power of a synchronous AC motor in terms of current amplitude and angle is particularly convenient, as modern AC drives, including those used in this thesis, operate with current control. The torque of the IPMSM, driven by the magnetic field, can be represented with (3.9), as shown for surface-mounted PMSM. In (3.9), the angle γ represents the angle between the stator current and the BEMF and becomes positive for I_s lagging E_a . In the instance of the IPMSM, the additional reluctance torque is represented by (3.16) and the total torque is described in (3.17).

$$T_{reluctance} = 3 \frac{P}{2} (L_d - L_q) I_s^2 \cos \gamma \sin \gamma \quad (3.16)$$

$$T_e = T_{reaction} + T_{reluctance} \quad (3.17)$$

The internal power factor angle γ is regulated through the vector control method, which transforms the AC motor's behavior into synchronously rotating reference frame coordinates a concept explored further in the next chapter. This method allows precise control of the motor's current vector to achieve specific operational objectives. For example, the angle γ can be set to $\gamma = 0^\circ$ or adjusted to optimize maximum torque per ampere (MTPA) performance, or to enable field weakening for extended speed range operation.

Field weakening, a critical feature of industrial drives, is employed to enhance the performance of voltage source inverters (VSI) during high-speed motor operation. At high speeds, the motor's back-emf increases, leading to higher voltage demands from the current controller to the VSI, which are constrained by the DC-bus voltage limit. To extend motor speed or maintain effective current regulation without saturating the VSI, field weakening reduces the BEMF, which is in proportion with the flux and motor velocity. This is achieved by attenuating the field flux through the generation of an MMF component opposing the flux, controlled by setting $\gamma < 0^\circ$.

For non-salient AC machines, like as induction motors and SMPMSMs where flux weakening is challenging or impractical, this approach reduces the torque-per-ampere ratio due to variation from the maximum torque condition ($\gamma = 0^\circ$). In wound-field synchronous machines, $\gamma < 0^\circ$ reduces reaction torque and introduces negative reluctance torque. Conversely, in IPM motors, field weakening ($\gamma < 0^\circ$) not only lowers back-emf to support high-speed operation but can also positively influence torque output, as MTPA is achieved at specific negative γ angles.

As a result, IPM motors are typically operated with $\gamma < 0^\circ$ across the entire speed range, including low and zero speeds, to maximize efficiency and accommodate field weakening at higher speeds. However, this operational approach necessitates external controllers to determine the optimal γ angle for MTPA and field weakening.

3.4 Conclusion

This chapter has provided a comprehensive overview of the basic structure and operating principles of PMSMs, with a particular focus on IPMSMs. By examining their electromagnetic principles, construction, and operational characteristics, this discussion has highlighted the distinct advantages of IPMSMs over other motor types. Key attributes such as enhanced flux density, robust field-weakening capability, and high torque and power density make IPMSMs particularly suitable for demanding industrial applications.

Additionally, the distinctions between Surface-Mounted Permanent Magnet (SMPM) motors and IPMSMs were explored, emphasizing the latter's mechanical and thermal resilience and its capability to support extended operational speed ranges. The salient pole design of IPMSMs introduces a reluctance torque component, further enhancing their efficiency and performance in modern drive systems.

Chapter 4

Field Oriented Control Structure of IPMSM

4.1 Introduction

This chapter explores the field-oriented control (FOC) structure of Interior Permanent Magnet Synchronous Motors (IPMSMs), a critical methodology for achieving high-performance motor control. The focus of this chapter is to explore the mathematical modelling of IPMSMs, including coordinate transformations, which simplify the control problem by transitioning to a two-axis framework from the three-phase system. These transformations, specifically Clarke and Park transformations, enable the decoupling of torque and flux control, making the IPMSM operate similarly to a DC motor in terms of control simplicity.

The chapter also introduces the essential role of inverters in implementing the field-oriented control strategy, discussing their function in converting control signals into three-phase voltages for the motor. PWM techniques, such as sinusoidal PWM and space vector PWM, are detailed and integral to generating precise control signals for inverters.

By establishing the mathematical foundation and practical implementation details, Chapter 4 provides the necessary insights for understanding how FOC enables efficient, precise, and responsive control of IPMSMs. This knowledge is instrumental in addressing the challenges of sensorless control systems, which rely heavily on accurate mathematical models and robust control strategies. Through this exploration, the chapter lays the groundwork for developing advanced control mechanisms discussed in later chapters.

4.2 Mathematical Model of IPMSM

An IPMSM features a rotor embedded with permanent magnets, which generate a magnetic field essential for motor operation. These magnets are strategically placed within the rotor to enhance performance characteristics such as torque generation,

efficiency, and thermal stability. In the stator, three-phase windings are distributed symmetrically, with currents flowing through them offset by a 120-degree phase difference to produce a rotating magnetic field. This interaction between the rotor's magnetic field and the stator's rotating field is the fundamental principle behind torque production in an IPMSM. The structure of an IPMSM, as illustrated in Figure 4.1, showcases the integration of the rotor, stator windings, and permanent magnets. According to Faraday's Law, a changing magnetic field induces a voltage. As the rotor's permanent magnets rotate, they create a back electromotive force (BEMF) in the stator windings. Each stator winding possesses its own inherent resistance and inductance due to the coil structure. Based on these properties, an equivalent circuit for a single phase can be modeled as shown below.

- The stator windings generate a sinusoidal magnetomotive force (MMF) distribution. Any space harmonics in the air gap are assumed to be negligible.
- The air-gap reluctance consists of a constant component along with a sinusoidally varying component.
- The three-phase sinusoidal voltage is balanced across all phases.
- Hysteresis and eddy current effects are disregarded.

Using Faraday's Law on the equivalent circuit shown in Figure 2.2 results in the following three-phase equations:

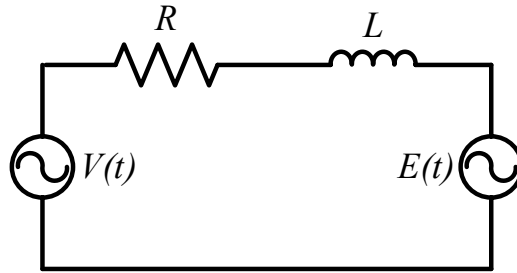


Figure 4.1 Single Phase Equivalent Circuit

$$\mathbf{v}_{abc}(t) = R_s \mathbf{i}_{abc}(t) + p\mathbf{\Lambda}_{abc} \quad (4.1)$$

Here “ p ” represents the derivative operator. For simplicity, time dependencies will not be explicitly shown in the equations unless specified otherwise. According to Faraday's Law, the time derivative of the stator flux linkage corresponds to the BEMF. The flux linkage for each phase is given as follows:

$$\mathbf{\Lambda}_{abc} = \mathbf{L}_{abc} \mathbf{i}_{abc} + \mathbf{\Lambda}_{abc}^{\text{pm}} \quad (4.2)$$

Equation (3.1) and (3.2) provides the most comprehensive form applicable to any alternating current (AC) machine, serving as the basis from which all AC machine models can be derived.

4.2.1 Clarke and Park Transformation

To control three-phase motors effectively, coordinate transformations are necessary. To control a three-phase motor similar to a DC motor, the equations in the abc frame are first expressed in a stationary system using the Clarke transformation. Next, the stationary system is transformed into a rotating dq -axis system through the Park transformation, allowing the abc system to be divided into d - and q -axes for control purposes. In this setup, the d -axis current can be thought of as analogous to the field current, while the q -axis current resembles the armature current, as in a DC motor.

The Clarke transformation is used to convert the abc system to the stationary $\alpha\beta 0$ -system. This process is also known as the $\alpha\beta 0$ -transformation. The matrix used for converting from the abc system to stationary $\alpha\beta 0$ -system is provided in equation 4.3.

$$\begin{bmatrix} u_\alpha \\ u_\beta \\ u_0 \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & -\frac{1}{3} & -\frac{1}{3} \\ 0 & \frac{1}{\sqrt{3}} & \frac{1}{\sqrt{3}} \\ \frac{1}{3} & \frac{1}{3} & \frac{1}{3} \end{bmatrix} \begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} \quad (4.3)$$

Applying the Clarke transformation to equation (4.2), results in the model in $\alpha\beta$ reference frame which can be represented as follows:

$$\begin{aligned} \begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} &= R_s \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \begin{bmatrix} \Sigma L + \Delta L \cos 2\theta_e & \Delta L \sin 2\theta_e \\ \Delta L \sin 2\theta_e & \Sigma L - \Delta L \cos 2\theta_e \end{bmatrix} p \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \\ &+ 2\Delta L \omega_e \begin{bmatrix} -\sin 2\theta_e & \cos 2\theta_e \\ \cos 2\theta_e & \sin 2\theta_e \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \omega_e \lambda_{pm} \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \end{aligned} \quad (4.4)$$

where $u_{\alpha\beta}$ and $i_{\alpha\beta}$ are the stator voltages and stator currents in the $\alpha\beta$ -frame, respectively; R_s is the stator resistance; L_d and L_q are the dq -frame inductances; $\Sigma L = (L_d + L_q)/2$; and $\Delta L = (L_d - L_q)/2$ are the average and differential inductances respectively and p is the differential operator. Rotor position information is explicitly included in both the state matrix and the disturbance matrix. When designing the observer, only the back EMF or flux component carries this rotor position information. This requirement becomes increasingly complex when the L_d and L_q inductances differ, as is often the case with IPMSMs, leading to more intricate mathematical expressions. Moreover, from a control perspective, the mathematical model in $\alpha\beta$ -frame does not

provide significant advantages, as it does not clearly define torque and flux-related terms. Nonetheless, this reference frame is valuable for implementing space vector modulation and DC bus ripple reduction algorithms. Additionally, with some modifications, demonstrated in references [12], [69], [70], the $\alpha\beta$ -frame can also be effectively employed in the observer model.

The Park transformation is utilized to convert the alpha-beta stationary system into the dq rotating system. This transformation creates a rotating system in the d- and q-axes, which is offset by an angle of θ_e from the original axes. Figure 4.2 illustrates the dq and alpha-beta axes. Equation (4.5) details the Park transformation matrix, while equation (4.6) outlines the inverse Park transformation matrix. The equation (4.5) is used when the d-axis aligns with the α -axis.

$$\begin{bmatrix} u_d \\ u_q \end{bmatrix} = \begin{bmatrix} \cos \theta_e & \sin \theta_e \\ -\sin \theta_e & \cos \theta_e \end{bmatrix} \begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} \quad (4.5)$$

$$\begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} = \begin{bmatrix} \cos \theta_e & -\sin \theta_e \\ \sin \theta_e & \cos \theta_e \end{bmatrix} \begin{bmatrix} u_d \\ u_q \end{bmatrix} \quad (4.6)$$

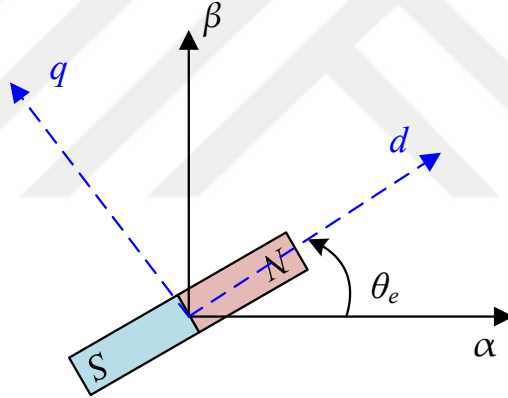


Figure 4.2 dq- and $\alpha\beta$ axis

By utilizing these transformations, the IPMSM's model based on the dq -axis is expressed in the equation (4.7).

$$\begin{aligned} u_d &= R_s i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q \\ u_q &= R_s i_q + L_q \frac{di_q}{dt} + \omega_e L_d i_d + \omega_e \lambda_{pm} \end{aligned} \quad (4.7)$$

Here, i_d and i_q represent the currents in the dq -axis, while u_d and u_q denote the stator voltages in the same axis. A key advantage of representing the system in the dq -frame is its ability to directly associate the current components with torque and flux linkage. The torque equation according to the dq -axis currents is given below:

$$T_e = \left(\frac{3}{2}\right) \left(\frac{P}{2}\right) (\lambda_{pm} i_q + i_d i_q (L_d - L_q)) \quad (4.8)$$

$$T_e - T_L = \frac{J d\omega_m}{dt} + B\omega_m \quad (4.9)$$

Here, ω_e represents the electrical angular velocity (rad/s), and λ_{pm} denotes the constant flux value in the permanent magnets (Wb). In the mechanical equations, T_L represents the load torque on the motor (N.m), J is the inertia ($kg.m^2$), and B indicates the friction coefficient. The term ω_m represents the mechanical angular velocity, which is proportional to the electrical angular velocity by the number of poles P . This proportional relationship between mechanical and electrical speeds also applies to mechanical and electrical positions. Additionally, as shown in equation (4.10) and (4.11), position values can be derived from speed values.

$$\omega_m = \left(\frac{2}{P}\right) \omega_e \quad (4.10)$$

$$\theta_e = \int \omega_e dt \quad (4.11)$$

4.3 Inverters

Through coordinate transformations, the motor can be controlled similarly to a DC motor, producing reference voltage values in the d- and q-axes once control is achieved. Although it is possible to convert from the dq axes back to the abc axes in a simulation environment, directly powering a three-phase motor with these dq-axis voltage values is not feasible in practice. To convert the generated dq-axis voltages to three-phase AC power, inverters are used. While inverters come in various topologies, the three-phase inverter is typically used for three-phase motor drives.

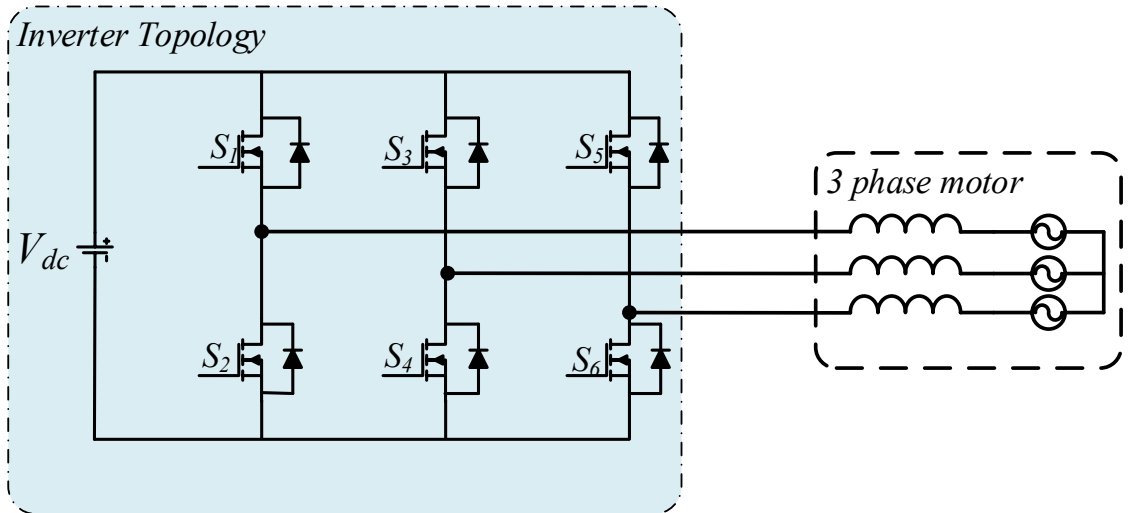


Figure 4.3 The Inverter Scheme

A three-phase inverter has six switching elements, which generate the output waveform based on the PWM signals provided. Figure 4.3 shows the schematic structure

of a three-phase inverter circuit. The six switches in the inverter circuit are triggered by information from the controller to produce the three-phase waves. The inverter output signal is adjusted based on whether each switch is set to 1 or 0. On the same leg, one switch must be set to 1 while the other is set to 0, and vice versa; otherwise, short-circuit or open-circuit issues may occur. Based on the switching functions, the phase voltages for an ideal, lossless inverter can be formulated, resulting in equation (4.12) [71]. This way, the three-phase voltage signals supplied to the motor are created through switching functions, leveraging the symmetry of the three-phase components in deriving equation (4.12) from the switching functions.

$$\begin{aligned} V_a &= \frac{V_{dc}}{3}(2S_a - S_b - S_c) \\ V_b &= \frac{V_{dc}}{3}(2S_b - S_a - S_c) \\ V_c &= \frac{V_{dc}}{3}(2S_c - S_a - S_b) \end{aligned} \quad (4.12)$$

S_a, S_b and S_c represent the upper or lower switches on the inverter legs. If S_a, S_b and S_c are considered the upper switches, then the lower switches will operate in the exact opposite states of S_a, S_b and S_c . For example, in a switching state of 0,0,1 for the upper switches, the lower switches will be set to 1,1,0 respectively. The inverter should be switched using PWM techniques to ensure the control system operates accurately and as desired by adjusting the output voltage and the switching frequency.

4.4 PWM Methods

4.4.1 Sinusoidal PWM

SPWM is a widely used and well-established technique for generating control signals in power electronics, particularly in inverter circuits. This method operates by comparing a sinusoidal reference waveform, which represents the desired output voltage or current, with a high-frequency triangular carrier waveform. The sinusoidal reference wave determines the amplitude and fundamental frequency of the output, enabling precise control of the inverter's output characteristics, such as frequency, magnitude, and phase. Meanwhile, the triangular carrier wave governs the converter's switching frequency, which directly impacts the quality of the output waveform and the efficiency of the power conversion process.

As illustrated in Figure 4.4, the activation signal for the high-side switch is derived by identifying the points where the sinusoidal reference wave intersects the triangular

carrier wave. The sinusoidal waveform represents the desired output voltage or current, while the triangular waveform acts as the modulating signal that defines the switching frequency. When the sinusoidal waveform is greater than the triangular waveform, the high-side switch is activated, allowing the corresponding phase to conduct current. Conversely, when the triangular waveform exceeds the sinusoidal waveform, the switch is turned off, and the current path is diverted through the complementary low-side switch or freewheeling diode. This process creates a sequence of pulses with widths that vary sinusoidally over a cycle, ensuring a close approximation of the desired sinusoidal output after filtering.

The SPWM technique offers several advantages, including reduced harmonic distortion in the output waveform and improved efficiency in power conversion. By adjusting the frequency and amplitude of the sinusoidal reference wave, the modulation index and output characteristics can be precisely controlled, making SPWM an effective and flexible strategy for a variety of applications.

$$S_a = \begin{cases} 1, & \text{if } V_{sine} \geq V_{tri} \\ 0, & \text{if } V_{sine} < V_{tri} \end{cases} \quad (4.13)$$

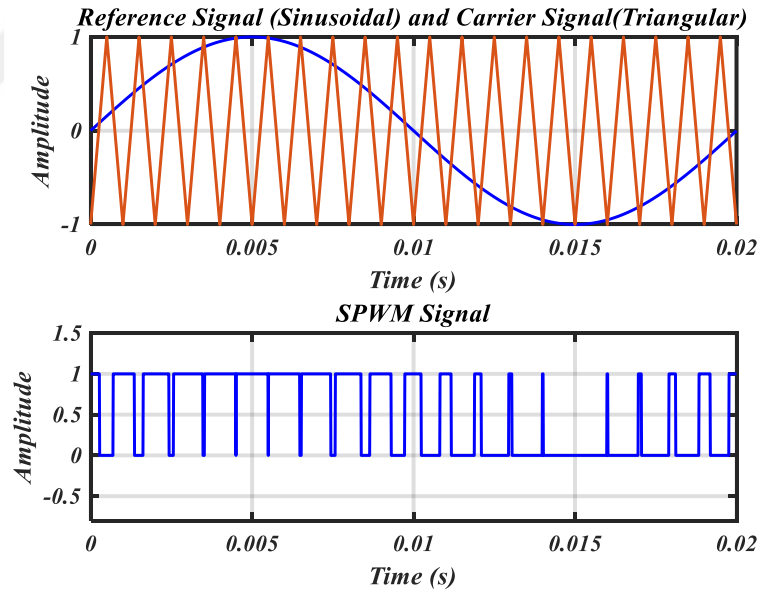


Figure 4.4 Sinusoidal PWM (SPWM) method by comparing with the triangular wave

Here, V_{sine} and V_{tri} represent the sinusoidal reference wave and the triangular carrier wave, respectively. The operation of SPWM, along with the resulting output voltage, is exhibited in Figure 4.4. The maximum peak output voltage achievable is $\frac{V_{dc}}{2}$.

To evaluate the usage of the DC-bus voltage in a PWM method, its fundamental output component is compared to the theoretical maximum achievable value, $\frac{2}{\pi}$, which

corresponds to the primary component in a six-step operation. The ratio of these values defines the modulation index (M). In the instance of SPWM, this index governs the amplitude of the fundamental output voltage relative to the DC-bus voltage, highlighting the efficiency of voltage utilization in this modulation technique.

4.4.2 Space Vector Pulse Width Modulation (SVPWM)

The SVPWM technique, commonly used in motor drives, is superior to other modulation techniques in the literature due to its higher battery utilization rate, lower switching losses, and reduced harmonic distortion [72]. Although the Sinusoidal PWM (SPWM) technique is also widely used, SVPWM provides approximately 15% higher voltage output and produces a smoother output compared to SPWM [73]. The disadvantage of SVPWM, however, is that it is more complex to implement than other modulation techniques.

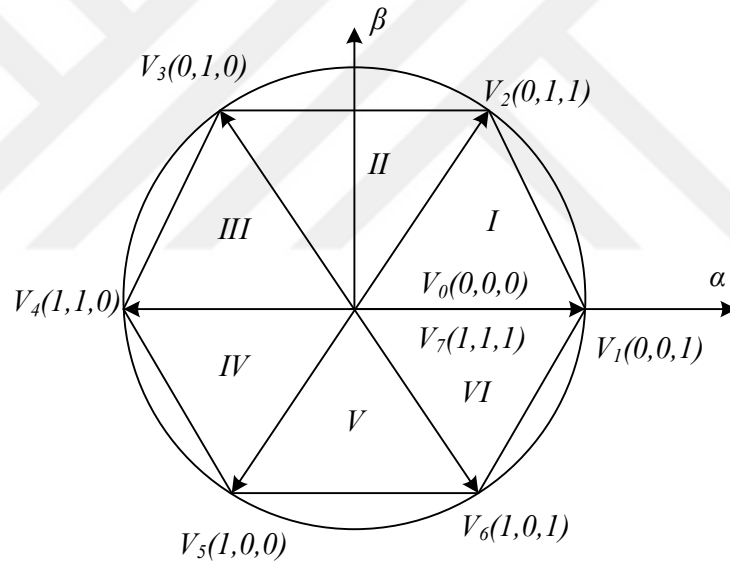


Figure 4.5 Voltage Hexagon and Sectors

In SVPWM, the aim is to generate a voltage reference vector V_{ref} that approximates a rotating magnetic field for the motor, based on a stationary $\alpha\beta$ reference frame. Here's the process and mathematical formulation for SVPWM:

- Convert the desired three-phase AC voltages V_a , V_b , V_c to the $\alpha\beta$ reference frame utilizing Clarke transformation:

$$V_\alpha = V_a \quad (4.14)$$

$$V_\beta = \frac{1}{\sqrt{3}}(V_\alpha + 2V_b) \quad (4.15)$$

With V_α and V_β we now have the voltage vector $V = V_\alpha + jV_\beta$.

- The desired output voltage vector is defined as:

$$V_{ref} = V_{ref} e^{j\theta} \quad (4.16)$$

Where V_{ref} is the amplitude of the reference vector, and θ is its angular position.

- The reference vector V_{ref} is synthesized using the nearest three switching states (V_1, V_2, V_0) in a hexagonal space defined by the inverter's switching states.

These vectors are :

$$\begin{aligned} V_1 &= \frac{2}{3}V_{dc}e^{j0}, & V_2 &= \frac{2}{3}V_{dc}e^{j\frac{\pi}{3}}, & V_3 &= \frac{2}{3}V_{dc}e^{j\frac{2\pi}{3}}, & V_4 &= \frac{2}{3}V_{dc}e^{j\pi}, \\ V_5 &= \frac{2}{3}V_{dc}e^{j\frac{4\pi}{3}}, & V_6 &= \frac{2}{3}V_{dc}e^{j\frac{5\pi}{3}} \end{aligned} \quad (4.17)$$

The zero vectors V_0 and V_7 are zero.

- With calculating the time intervals, it is aimed to approximate V_{ref} using the selected vectors within a switching period T_s . If the reference vector is in sector k (1 through 6), the time intervals T_1 and T_2 for active vectors V_0 and V_7 , and T_0 for the zero vectors, can be computed as follows:

$$\begin{aligned} T_1 &= \frac{\sqrt{3}T_s}{V_{dc}} |V_{ref}| \sin(n\frac{\pi}{3} - \theta) \\ T_2 &= \frac{\sqrt{3}T_s}{V_{dc}} |V_{ref}| \sin(\theta - \frac{(n-1)\pi}{3}) \\ T_0 &= T_s - T_1 - T_2 \end{aligned} \quad (4.18)$$

- The PWM switching signals for each phase are derived by combining T_1, T_2 and T_0 with appropriate switching sequences for each sector.

$$V_{ref} = \frac{T_1}{T_s} V_1 + \frac{T_2}{T_s} V_2 + \frac{T_0}{T_s} V_0 \quad (4.19)$$

In an inverter leg, the switches are designed such that they are never simultaneously open or closed. If both switches are open at the same time, no output voltage will be generated; if both are closed simultaneously, it would cause a short circuit. Therefore, the on and off durations of the upper and lower switches are carefully determined based on the calculated switching times. Figure 4.6 illustrates the required on and off durations for the switches in each sector. As shown in Figure 4.6, when switches S_1, S_3 and S_5 are in the "off" (0) position, switches S_2, S_4 and S_6 are in the "on" (1) position, and vice versa.

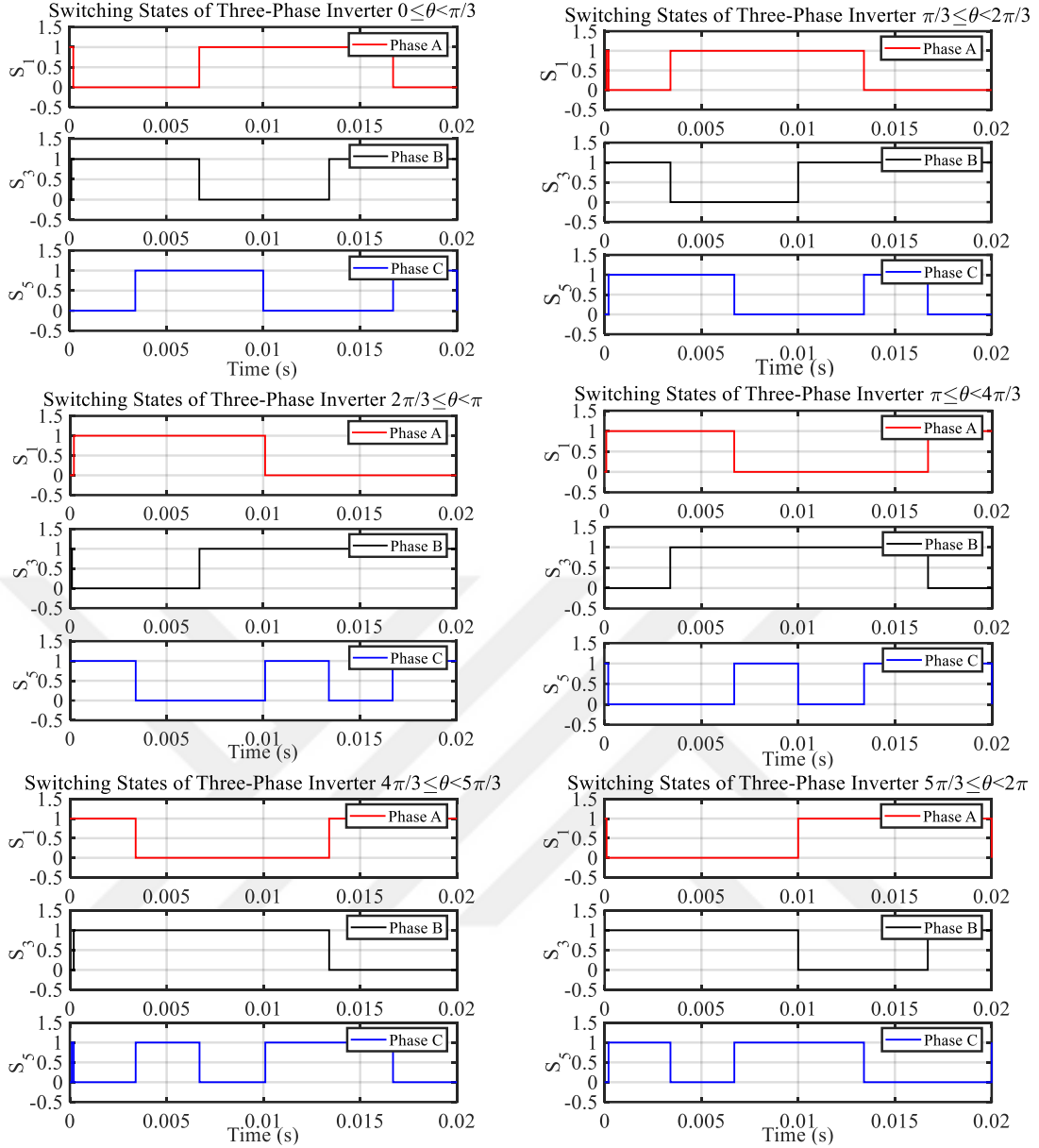


Figure 4.6 Switching times

4.5 IPMSM Control Methods

4.5.1 Vector Control

Vector control, also named as field-oriented control (FOC), is based on aligning the rotor flux with the stator current vector. This approach enables precise torque control in PMSMs by independently regulating the components of the stator current, typically into direct (d-axis) and quadrature (q-axis) components. The d-axis current aligns with the rotor flux to control magnetization, while the q-axis current governs torque production. By decoupling these components, vector control allows for highly efficient

and dynamic motor operation, mimicking the simplicity and precision of DC motor control while leveraging the superior performance characteristics of PMSMs.

This method not only enhances torque accuracy but also minimizes energy losses by optimizing the current usage according to load requirements. Additionally, vector control enables smooth and responsive performance across a wide speed range, including low-speed operation where traditional control methods might falter. The ability to maintain a constant torque-to-current ratio and achieve rapid acceleration or deceleration makes it ideal for applications demanding high performance, such as electric vehicles, robotics, and industrial automation. Furthermore, modern implementations of vector control utilize advanced digital signal processors (DSPs) or microcontrollers, ensuring real-time computation of control algorithms and seamless integration with sensor feedback systems, such as encoders or resolvers, for enhanced precision.

This method has become the standard for high-performance motor drives across various applications. For precise electromagnetic torque control, the stator current is categorized into two orthogonal components in the rotor reference frame as shown in (4.7):

- The i_d component controls the flux, often referred to as the excitation current.
- The i_q component generates torque, known as the torque-producing current.

To maximize performance, different control strategies are applied to manage i_d and i_q . Among these, two of the most widely used techniques are $i_d = 0$ control and MTPA control.

4.5.1.1 $i_d = 0$ Control Method

In vector control, setting the excitation current i_d to zero allows precise control of the motor. This ensures that the electromagnetic torque depends solely on the torque current i_q . As illustrated in Figure 4.7, the permanent flux linkage λ_{pm} aligns with the d -axis and rotates in sync with the rotor's electrical position θ_e , achieving rotor flux inducement. With the excitation current nullified, the torque current directly corresponds to the magnitude of the stator current. In the scheme, the stator voltage is represented with $\mathbf{v}$, while the stator current is represented as $\mathbf{i}$.

In this control strategy, torque is produced by managing i_q , making the system simple to execute while achieving effective torque control. It is particularly advantageous for surface-mounted PMSMs, as it minimizes the stator current necessary for a given torque output. However, for IPMSMs, this approach reduces motor efficiency by

neglecting the contribution of reluctance torque. The stator current load rises, as the load increases, leading to greater cross-axis flux. As a result, the phase difference between the stator current and voltage vectors becomes larger.

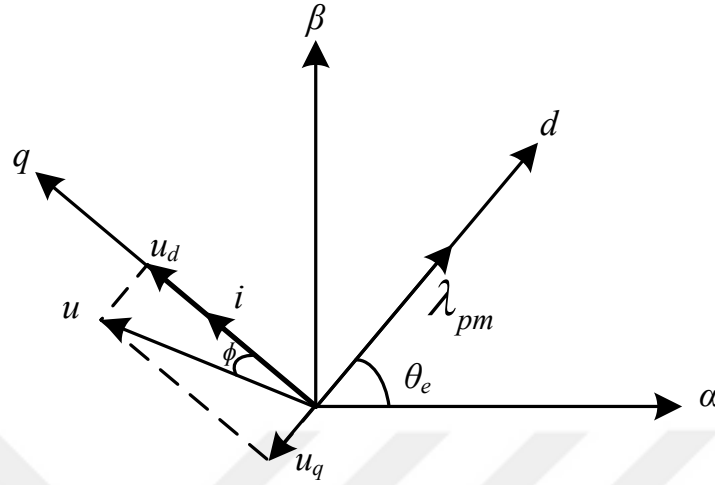


Figure 4.7 $i_d = 0$ Control Diagram

4.5.1.2 Maximum Torque Per Ampere (MTPA) Control Method

The MTPA method is a commonly used vector control approach for optimizing the operation of IPMSMs [74], [75]. Its key goal is to reduce the stator current required to achieve a specific output torque, thereby enhancing efficiency and minimizing energy losses. In the current plane, the MTPA point represents the point on the constant torque curve that is nearest to the origin. This location corresponds to the operating condition with the lowest current magnitude, enabling maximum torque generation with minimal electrical input. Reaching the MTPA point requires a careful adjustment of the d and q-axis currents to exploit the motor's magnetic saliency. As illustrated in Figure 4.8, this approach not only reduces copper losses but also helps to improve the motor's thermal performance and efficiency. This makes MTPA control particularly advantageous in high-demand applications such as electric vehicles and industrial systems that prioritize energy efficiency and dynamic response.

In MTPA strategy, if i_d and i_q are considered as the vector components of the stator current.

$$\begin{aligned} i_d &= -i_s \sin \beta \\ i_q &= i_s \cos \beta \end{aligned} \tag{4.20}$$

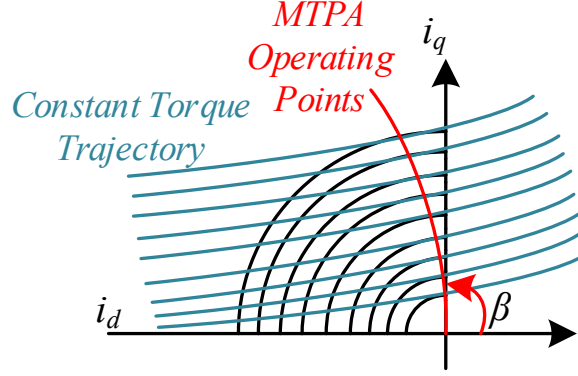


Figure 4.8 The constant torque trajectories and MTPA operating points in the current region

Where $i_s = \sqrt{i_d^2 + i_q^2}$ and β are stator current vector and the angle of the optimal stator current respectively. Substituting Equation (4.20) into Equation (4.8), the optimal operating point is described by finding the angle where the derivative of the torque according to the current angle equals zero. The derivative expression of the torque equation is provided in Equation (3.20).

$$\frac{dT_e}{d\beta} = \left(\frac{3}{2}\right) \left(\frac{P}{2}\right) [-i_s \lambda_{pm} \sin \beta + (L_d - L_q) i_s^2 \cos 2\beta] = 0 \quad (4.21)$$

The value of the β angle is determined by setting the derivative expression in (4.21) equal to zero. Considering that the machine parameters are constant, the obtained β angle is given in (4.22).

$$\beta = \sin^{-1} \left(\frac{-\lambda_{pm} + \sqrt{\lambda_{pm}^2 + 8(L_d - L_q)^2 i_s^2}}{4(L_d - L_q) i_s} \right) \quad (4.22)$$

By substituting (4.20) into (4.21), and solving for i_d , (4.23) is derived.

$$i_d = \frac{-\lambda_{pm}}{4(L_d - L_q)} - \sqrt{\left(\frac{\lambda_{pm}^2}{16(L_d - L_q)^2} + \frac{i_s^2}{2} \right)} \quad (4.23)$$

$$i_q = \sqrt{i_s^2 - i_d^2}$$

4.5.2 Flux Weakening Control

In PMSM drive systems, the stator current amplitude and voltage magnitude are constrained by the rated capacity of the IGBT and DC-bus voltage, respectively. These limitations are commonly expressed as the maximum current $I_{s,max}$ and maximum voltage $U_{s,max}$, which define the operational boundaries of the system. The stator current amplitude, i_s , represents the electrical loading of the motor and is directly linked to the

torque-generating capability of the drive. Exceeding $I_{s,max}$ risks thermal overload and potential damage to the IGBT modules. Similarly, the voltage vector magnitude, $|\mathbf{u}|$, determines the motor's ability to achieve the required speed under load. If $U_{s,max}$ is exceeded, it can lead to overvoltage conditions that compromise the DC-link components or prevent stable operation of the motor control system. These limitations are critical design parameters that must be accounted for in both hardware selection and control strategy development to ensure system reliability, efficiency, and performance under all operating conditions.[76]. For optimal and safe operation, the magnitudes of the stator voltage and current vectors must adhere to the following limitations:

$$\begin{aligned} |i_s| &\leq I_{s,max} \\ |\mathbf{u}| &\leq U_{s,max} \end{aligned} \quad (4.24)$$

Where $|\mathbf{u}| = \sqrt{u_d^2 + u_q^2}$ and $|i_s| = \sqrt{i_d^2 + i_q^2}$ denote the magnitudes of the stator voltage and current vectors, respectively.

From Equations (4.7), it is evident that under steady-state operation of the PMSM, the differential terms in the voltage equations are reduced to zero. Furthermore, the stator resistance voltage drop becomes negligible above the base speed. As a result, the motor voltage equations can be given as:

$$\begin{aligned} u_d &= -\omega_e L_q i_q \\ u_q &= \omega_e L_d i_d + \omega_e \lambda_{pm} \end{aligned} \quad (4.25)$$

By substituting the steady state voltages in (4.25) into (4.24), the voltage constraint can be derived as :

$$(L_d i_d + \lambda_{pm})^2 + (L_q i_q)^2 \leq \left(\frac{U_{s,max}}{\omega_e} \right)^2 \quad (4.26)$$

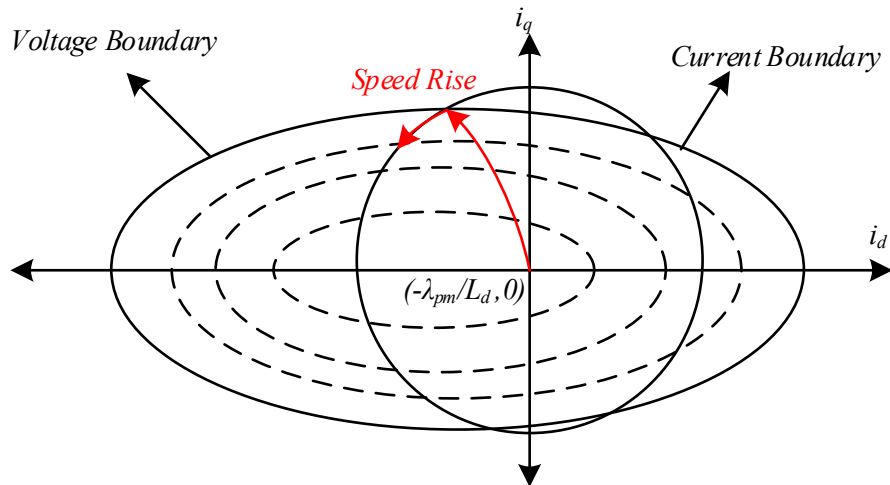


Figure 4.9 Schematic diagram of current and voltage boundary

From (4.26), it is evident that the voltage border for the IPMSM forms an elliptical shape, centered at the point $(\lambda_{pm}/L_d, 0)$. The radius of this ellipse decreases proportionally as the operating speed increases. The current boundary, described by Equation (4.27), is represented as a circle centered at the origin, with a radius equal to $I_{s,max}$. These boundaries are illustrated in Figure 4.9.

$$i_d^2 + i_q^2 \leq I_{s,max}^2 \quad (4.27)$$

The voltage limit ellipse progressively decreases toward its center as velocity increases. When the current path follows the indicated arrow toward the point where the current and voltage boundaries meet, the d -axis current is constrained to shift leftward. This movement characterizes the current trajectory during weak-field operations.

4.6 Conclusion

This chapter has provided a detailed exploration of the field-oriented control (FOC) structure for IPMSMs, establishing its importance in achieving precise and efficient motor control. By presenting the mathematical modeling of IPMSMs, including Clarke and Park transformations, the chapter demonstrated how these coordinate transformations simplify the control problem by decoupling torque and flux control.

The role of inverters in translating control signals into three-phase voltages was emphasized, along with an examination of Pulse Width Modulation (PWM) techniques such as sinusoidal PWM and Space Vector PWM. These techniques were shown to be vital in ensuring accurate and smooth control signal generation, enabling stable and efficient motor operation.

The insights provided in this chapter form a strong foundation for understanding the practical implementation of FOC in sensorless control systems. This understanding is critical as the thesis progresses to more advanced topics, such as the integration of observer techniques and adaptive control strategies. By bridging theoretical models and practical applications, this chapter sets the stage for addressing the complex challenges associated with real-time control of IPMSMs in dynamic operating conditions.

Chapter 5

The Proposed Sensorless Control and Parameter Estimation

5.1 Introduction

This chapter focuses on the proposed sensorless control strategy for IPMSMs, addressing the challenges of rotor position and speed estimation across a wide range of operating conditions, including low-speed and standstill scenarios. This chapter builds upon the foundational principles established in previous chapters, leveraging the mathematical models and control structures of IPMSMs to develop advanced sensorless techniques.

The discussion begins with an overview of the dynamic model of IPMSMs as applied in sensorless control systems, emphasizing the importance of accurate estimation of the position and speed for effective motor control. The limitations of conventional sensorless methods are then highlighted, providing context for the introduction of a novel approach based on high-frequency voltage injection and sliding mode observers (SMOs) integrated with a PLL.

This chapter also explores the design and implementation of the proposed SMO-PLL structure, focusing on its ability to overcome the challenges of low-speed operation and parameter variations. The method's robustness, efficiency, and computational simplicity make it a promising solution for real-time applications. Through this exploration, this chapter lays the groundwork for experimental validation and performance analysis in subsequent chapters, aiming to contribute to significant advancement in the field of sensorless motor control.

5.2 Position Sensorless Control of IPMSM

The position sensorless control method utilizing the dynamic models of IPMSMs is preferable for electric drives due to its cost-effectiveness, mechanical robustness, reduced

hardware complexity, and enhanced reliability of the overall drive system. By removing the need for physical position transducers, this method reduces component costs and increases durability, as it removes potential points of failure that sensors may introduce. However, the performance of sensorless control is strongly reliant on the accuracy with which the rotor speed and/or position is estimated. This accuracy is influenced by both the proposed sensorless estimation algorithm and the precision of the dynamic model used for the IPMSM. Variations in model parameters owing to temperature changes, magnetic saturation, and load disturbances can affect the excellence of the position and speed estimates, potentially leading to degraded control performance. Therefore, a well-designed sensorless control system must consider these factors to ensure optimal drive performance across various operating conditions.

5.2.1 Dynamic Model of IPMSM Used in Sensorless Control

A diagram illustrating the basic configuration of a magnet and winding orientations for an IPMSM is provided in Figure 5.1. This diagram serves as a foundational tool for visualizing the motor's operational principles. In this diagram, the stator stationary reference frame, represented by the $\alpha\beta$ axes, is fixed relative to the stator windings and provides a basis for understanding the motor's magnetic interactions from a static perspective. These axes are crucial for analyzing the interaction of alternating currents within the stator windings and their contribution to generating a rotating magnetic field. In contrast, the dq reference frame revolves at a synchronous velocity, aligning with the rotor's magnetic field. This frame is particularly useful for analyzing and simplifying the dynamic behavior of the motor during operation. Here, the d -axis, is aligned with the rotor's magnetic flux, while the q -axis is orthogonal to it. This orthogonal relationship between the axes aids in decoupling the torque and flux components of the motor's operation, a key aspect of advanced control strategies. The electrical angle, θ_e , represents the angular position of the rotor's magnetic field relative to the stationary $\alpha\beta$ axes, providing a dynamic measure of the rotor's orientation in the stator frame, while ω_e denotes the rotor's electrical angular velocity, corresponding to the rate of rotation in electrical rad/s. These parameters are pivotal in synchronizing control signals and ensuring efficient motor operation under varying load and speed conditions. The use of both the $\alpha\beta$ axes and the rotating dq -axes is essential for transforming the control equations into a form that simplifies the analysis of magnetic interactions within the

IPMSM. This transformation enables effective deployment of advanced control strategies, such as sensorless estimation techniques, where rotor position and speed are estimated rather than directly measured, and field-oriented control (FOC), which achieves precise control of the motor's torque and flux. Additionally, these reference frames facilitate real-time computations in digital controllers, enhancing the motor's responsiveness and efficiency in practical applications across diverse industries.

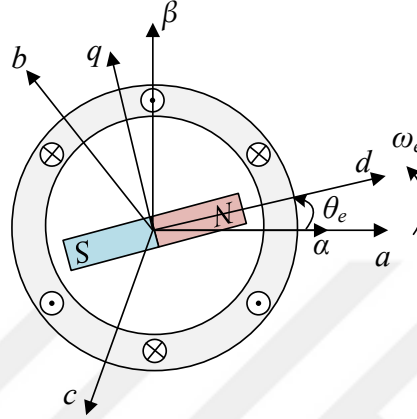


Figure 5.1 The reference frame definitions of the IPMSM

In the suggested sensorless control technique, in order to directly estimate θ_e , the EEMF-based model of IPMSM can be obtained by rewriting (4.4) including both $2\theta_e$ and θ_e as below:

$$\begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} = \begin{bmatrix} R_s + pL_d & \omega_e(L_d - L_q) \\ -\omega_e(L_d - L_q) & R_s + pL_d \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} \quad (5.1)$$

In (4.1), the EEMF components (e_α and e_β) consisting of only θ_e are as follows:

$$\begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = \Gamma \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \quad (5.2)$$

$$\Gamma = (L_d - L_q)(\omega_e i_d - di_q/dt) + \omega_e \lambda_{pm}$$

where λ_{pm} is the PM flux linkage; Γ is the speed -dependent EEMF gain; and i_d and i_q are stator currents in the dq -frame, respectively. These parameters are fundamental in describing the electromagnetic behavior of the Interior Permanent Magnet Synchronous Motor (IPMSM) within the rotating reference frame. λ_{pm} captures the contribution of the rotor's permanent magnets to the magnetic circuit, while Γ encapsulates the relationship between rotor speed and the induced EMF, which plays a pivotal role in dynamic performance and control. Please note that the derivative term in the EEMF (Eq. (5.2)) remains effective even at very low and zero speed ranges, where traditional back-EMF-based sensing methods often struggle. This characteristic makes the derivative term particularly advantageous for scenarios requiring high-precision control or estimation at

5.2.3 Conventional SMO Design

The conventional SMO structure by utilizing the state model in (5.3) can be given as:

$$\begin{aligned}\dot{\hat{i}}_{\alpha} &= \left(-\frac{R_s}{L_d}\right)\hat{i}_{\alpha} - \frac{\widehat{\omega}_e(L_d - L_q)}{L_d}\hat{i}_{\beta} + \frac{u_{\alpha}}{L_d} - \frac{\mu_1 F(\tilde{i}_{\alpha})}{L_d} \\ \dot{\hat{i}}_{\beta} &= \left(-\frac{R_s}{L_d}\right)\hat{i}_{\beta} + \frac{\widehat{\omega}_e(L_d - L_q)}{L_d}\hat{i}_{\alpha} + \frac{u_{\beta}}{L_d} - \frac{\mu_2 F(\tilde{i}_{\beta})}{L_d}\end{aligned}\quad (5.4)$$

where $\hat{i}_{\alpha}$ and $\hat{i}_{\beta}$ are the estimated current value; μ_1 and μ_2 are the observer gains; $F(s_x)$ is the signum function but in order to overcome the chattering phenomena, the signum function is changed with the sigmoid function in the SMO, which is defined as

$$F(s_x) = (1 - e^{-\sigma s_x}) / (1 + e^{-\sigma s_x}) \quad (5.5)$$

where σ is a positive parameter; and s_x represents the sliding surface selected on the estimation error of the stator current and is defined as

$$\begin{aligned}s_{\alpha} &= \tilde{i}_{\alpha} = \hat{i}_{\alpha} - i_{\alpha} = 0 \\ s_{\beta} &= \tilde{i}_{\beta} = \hat{i}_{\beta} - i_{\beta} = 0\end{aligned}\quad (5.6)$$

The estimation error dynamics can be written by taking the difference between (5.4) and (5.3).

$$\begin{aligned}\dot{\tilde{i}}_{\alpha} &= \left(-\frac{R_s}{L_d}\right)\tilde{i}_{\alpha} - \frac{(L_d - L_q)}{L_d}(\widehat{\omega}_e\hat{i}_{\beta} - \omega_e i_{\beta}) + \frac{1}{L_d}(e_{\alpha} - \mu_1 F(\tilde{i}_{\alpha})) \\ \dot{\tilde{i}}_{\beta} &= \left(-\frac{R_s}{L_d}\right)\tilde{i}_{\beta} + \frac{(L_d - L_q)}{L_d}(\widehat{\omega}_e\hat{i}_{\alpha} - \omega_e i_{\alpha}) + \frac{1}{L_d}(e_{\beta} - \mu_2 F(\tilde{i}_{\beta}))\end{aligned}\quad (5.7)$$

For analyzing the stability of the error dynamics in (5.7), a Lyapunov function is defined by

$$V_{\alpha\beta} = \frac{s_{\alpha\beta}^T s_{\alpha\beta}}{2} = \frac{\tilde{i}_{\alpha}^2 + \tilde{i}_{\beta}^2}{2} \quad (5.8)$$

where $s_{\alpha\beta} = \begin{bmatrix} \tilde{i}_{\alpha} \\ \tilde{i}_{\beta} \end{bmatrix}$. The stability condition for the SMO can be obtained from the time derivative of (5.8) as

$$\dot{V}_1 = \tilde{i}_{\alpha}\dot{\tilde{i}}_{\alpha} + \tilde{i}_{\beta}\dot{\tilde{i}}_{\beta} \leq 0 \quad (5.9)$$

Substituting (5.7) into (5.9) and taking $\sigma \gg 1$, (5.9) is rewritten as below:

$$\dot{V}_1 = \frac{-R_s}{L_d}(\tilde{i}_{\alpha}^2 + \tilde{i}_{\beta}^2) + \frac{\tilde{i}_{\alpha}}{L_d}(e_{\alpha} - \mu_1 F(\tilde{i}_{\alpha})) + \frac{\tilde{i}_{\beta}}{L_d}(e_{\beta} - \mu_2 F(\tilde{i}_{\beta})) < 0 \quad (5.10)$$

To ensure the stability of the conventional SMO, the observer gains μ_1 and μ_2 should be selected large enough to satisfy (5.11):

$$\mu_1 F(\tilde{i}_{\alpha}) > \max(|e_{\alpha}|), \mu_2 F(\tilde{i}_{\beta}) > \max(|e_{\beta}|) \quad (5.11)$$

When (5.11) is satisfied by selecting appropriate μ_1 and μ_2 , the system will reach the sliding surface ($s_\alpha = 0$ and $s_\beta = 0$) and then the following criteria are derived by using (5.3) and (5.4):

$$\hat{e}_\alpha = \mu_1 F(s_\alpha), \quad \hat{e}_\beta = \mu_2 F(s_\beta) \quad (5.12)$$

5.2.4 The Proposed Voltage Injection-Based SMO And PLL For EEMF, Speed, and Position Estimation

Conventionally, the SMO estimating the EEMF terms and then PLL uses these estimated EEMF terms to perform the simultaneous estimations of rotor speed and position; Thus, the conventional SMO with PLL needs reasonably accurate stator voltages u_α and u_β for a desirable estimation performance. However, u_α and u_β are extremely small at extremely low speeds, and thus, the strong effect of inverter nonlinearities on the system behavior becomes inevitable, which can cause unstable behavior in the observer and drive system. To eliminate such an issue in this thesis, it is the first time a HF voltage injection-based SMO is developed and then combined with PLL for the online estimations of EEMF, rotor speed, and rotor position. In the proposed SMO, the HF voltage is inserted only on d -axis in order to prevent torque ripple due to the HF current in q -axis; The purpose of the voltage injection is to strengthen the attenuated EEMF signal by providing phase alignment at extremely low and zero speed and to enhance the very low/zero speed estimation performance of the conventional SMO with PLL. For this aim, firstly the cosinusoidal HF voltage in (5.13) is inserted into dq -frame, as shown in Figure

$$\begin{bmatrix} u_{dh} \\ u_{qh} \end{bmatrix} = \begin{bmatrix} U_h \cos \omega_h t \\ 0 \end{bmatrix} \quad (5.13)$$

where u_{dh} and u_{qh} are the injected signals to the reference voltages; U_h and ω_h are the magnitude and the frequency of the inserted voltage, respectively.

Next, (5.13) are transformed into the $\alpha\beta$ frame as

$$\begin{bmatrix} u_{\alpha h} \\ u_{\beta h} \end{bmatrix} = \begin{bmatrix} u_{dh} \cos \theta_e \\ u_{dh} \sin \theta_e \end{bmatrix} \quad (5.14)$$

Then, differently from [71] using the dq -model of IPMSM, signal processing algorithms, and not depending on EEMF, this paper proposes to utilize the $\alpha\beta$ -model with the EEMF term of IPMSM in (5.1). As seen in Figure 5.2, since (5.14) is added to the main driving voltages (u_α and u_β), the $\alpha\beta$ -model in (5.1) becomes as below:

$$\begin{bmatrix} u_\alpha + u_{\alpha h} \\ u_\beta + u_{\beta h} \end{bmatrix} = \begin{bmatrix} R_s + pL_d & \omega_e(L_d - L_q) \\ -\omega_e(L_d - L_q) & R_s + pL_d \end{bmatrix} \times \begin{bmatrix} i_\alpha + i_{\alpha h} \\ i_\beta + i_{\beta h} \end{bmatrix} + \begin{bmatrix} e_\alpha + e_{\alpha h} \\ e_\beta + e_{\beta h} \end{bmatrix} \quad (5.15)$$

where $u_{\alpha h}$, $u_{\beta h}$, $i_{\alpha h}$, $i_{\beta h}$, $e_{\alpha h}$, $e_{\beta h}$ are the HF terms of voltages, currents, and the EEMF

values in the $\alpha\text{-}\beta$ frame, respectively.

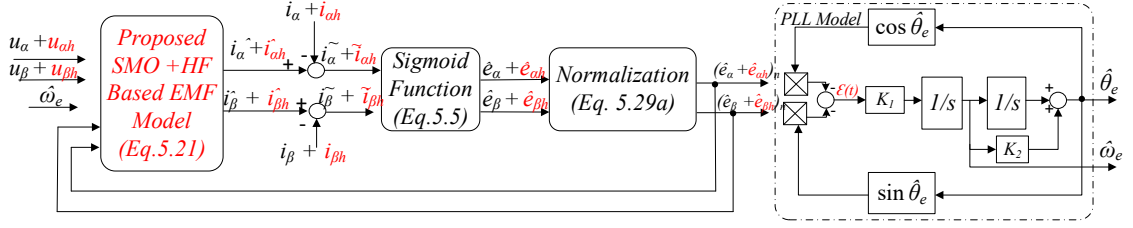


Figure 5.3 Sliding mode observer with PLL structure.

Later, the counterpart of (5.2), which is the components of EEMF with the HF terms in (5.15), can be derived by considering stationary $\alpha\beta$ frame with the HF terms as below:

$$\begin{aligned} \begin{bmatrix} e_\alpha + e_{\alpha h} \\ e_\beta + e_{\beta h} \end{bmatrix} &= 2\Delta L \begin{bmatrix} -\sin^2 \theta_e & \sin \theta_e \cos \theta_e \\ \sin \theta_e \cos \theta_e & -\cos^2 \theta_e \end{bmatrix} p \begin{bmatrix} i_\alpha + i_{\alpha h} \\ i_\beta + i_{\beta h} \end{bmatrix} \\ &+ 4\omega_e \Delta L \begin{bmatrix} -\sin \theta_e \cos \theta_e & -\sin^2 \theta_e \\ \cos^2 \theta_e & \sin \theta_e \cos \theta_e \end{bmatrix} \begin{bmatrix} i_\alpha + i_{\alpha h} \\ i_\beta + i_{\beta h} \end{bmatrix} \\ &+ \omega_e \lambda_{pm} \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \end{aligned} \quad (5.16)$$

Thus, considering the injected voltage vector in the $\alpha\text{-}\beta$ frame separately and neglecting the ω_e -dependent terms in (5.16) since the HF voltage injection will be used at low and zero speeds, the HF voltage part of (5.15) becomes.

$$\begin{bmatrix} u_{\alpha h} \\ u_{\beta h} \end{bmatrix} = \begin{bmatrix} R_s + pL_d & 0 \\ 0 & R_s + pL_d \end{bmatrix} \begin{bmatrix} i_{\alpha h} \\ i_{\beta h} \end{bmatrix} + \begin{bmatrix} e_{\alpha h} \\ e_{\beta h} \end{bmatrix} \quad (5.17)$$

Where,

$$\begin{bmatrix} e_{\alpha h} \\ e_{\beta h} \end{bmatrix} = 2\Delta L \begin{bmatrix} -\sin^2 \theta_e & \sin \theta_e \cos \theta_e \\ \sin \theta_e \cos \theta_e & -\cos^2 \theta_e \end{bmatrix} p \begin{bmatrix} i_{\alpha h} \\ i_{\beta h} \end{bmatrix} \approx -2\Delta L \left(\frac{di_{qh}}{dt} \right) \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} = (\Gamma_h) \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \quad (5.18)$$

The main driving voltage part (without the HF terms) of (5.15) is exactly as in (5.1) and (5.2). Finally, the EEMF components in (5.16) including the effects of both HF voltage injection and main driving voltage can be written as Finally, the EEMF components in (5.16) including the effect of the HF voltage injection can be written as

$$\begin{bmatrix} e_\alpha + e_{\alpha h} \\ e_\beta + e_{\beta h} \end{bmatrix} = (\Gamma + \Gamma_h) \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \quad (5.19)$$

where, Γ_h is defined as in (5.18) and represents the HF term of EEMF because of the injected HF voltage.

Now, it is time to design a suitable SMO estimator in order to estimate the EEMF components in (5.19) and a PLL for the estimations of the rotor velocity and position. For this aim, (5.15) can be given in state space form as in (5.20):

$$\begin{aligned} \dot{i}_\alpha + \dot{i}_{\alpha h} &= \left(-\frac{R_s}{L_d}\right)(i_\alpha + i_{\alpha h}) - \frac{\omega_e(L_d - L_q)}{L_d}(i_\beta + i_{\beta h}) + \frac{u_\alpha + u_{\alpha h}}{L_d} - \frac{e_\alpha + e_{\alpha h}}{L_d} \\ \dot{i}_\beta + \dot{i}_{\beta h} &= \left(-\frac{R_s}{L_d}\right)(i_\beta + i_{\beta h}) + \frac{\omega_e(L_d - L_q)}{L_d}(i_\alpha + i_{\alpha h}) + \frac{u_\beta + u_{\beta h}}{L_d} - \frac{e_\beta + e_{\beta h}}{L_d} \end{aligned} \quad (5.20)$$

The proposed SMO structure by using the state model in (5.20) can be given as

$$\begin{aligned} (\dot{\hat{i}}_\alpha + \dot{\hat{i}}_{\alpha h}) &= \left(-\frac{R_s}{L_d}\right)(\hat{i}_\alpha + \hat{i}_{\alpha h}) - \frac{\omega_e(L_d - L_q)}{L_d}(\hat{i}_\beta + \hat{i}_{\beta h}) + \frac{(u_\alpha + u_{\alpha h})}{L_d} - \frac{\mu_1 F(\tilde{i}_\alpha + \tilde{i}_{\alpha h})}{L_d} \\ (\dot{\hat{i}}_\beta + \dot{\hat{i}}_{\beta h}) &= \left(-\frac{R_s}{L_d}\right)(\hat{i}_\beta + \hat{i}_{\beta h}) + \frac{\omega_e(L_d - L_q)}{L_d}(\hat{i}_\alpha + \hat{i}_{\alpha h}) + \frac{(u_\beta + u_{\beta h})}{L_d} - \frac{\mu_2 F(\tilde{i}_\beta + \tilde{i}_{\beta h})}{L_d} \end{aligned} \quad (5.21)$$

Again, considering the HF part of (5.21) separately and neglecting the ω_e -dependent terms, (5.21) becomes

$$\begin{aligned} (\dot{\hat{i}}_{\alpha h}) &= \left(-\frac{R_s}{L_d}\right)(\hat{i}_{\alpha h}) + \frac{(u_{\alpha h})}{L_d} - \frac{\mu_1 F(\tilde{i}_{\alpha h})}{L_d} \\ (\dot{\hat{i}}_{\beta h}) &= \left(-\frac{R_s}{L_d}\right)(\hat{i}_{\beta h}) + \frac{(u_{\beta h})}{L_d} - \frac{\mu_2 F(\tilde{i}_{\beta h})}{L_d} \end{aligned} \quad (5.22)$$

Then, the error dynamics can be expressed as:

$$\begin{aligned} \dot{\tilde{i}}_{\alpha h} &= (-R_s/L_d)\tilde{i}_{\alpha h} + \frac{e_{\alpha h}}{L_d} - \frac{\mu_1 F(\tilde{i}_{\alpha h})}{L_d} \\ \dot{\tilde{i}}_{\beta h} &= (-R_s/L_d)\tilde{i}_{\beta h} + \frac{e_{\beta h}}{L_d} - \frac{\mu_2 F(\tilde{i}_{\beta h})}{L_d} \end{aligned} \quad (5.23)$$

For analyzing the stability of the error dynamics in (5.23), a Lyapunov function is defined by

$$V_{\alpha\beta h} = \frac{s_{\alpha\beta h}^T s_{\alpha\beta h}}{2} = \frac{\tilde{i}_{\alpha h}^2 + \tilde{i}_{\beta h}^2}{2} \quad (5.24)$$

where $s_{\alpha\beta h} = \begin{bmatrix} \tilde{i}_{\alpha h} \\ \tilde{i}_{\beta h} \end{bmatrix}$. The stability condition for the SMO can be obtained from the time-derivative of (5.24) as

$$\dot{V}_2 = \tilde{i}_{\alpha h} \dot{\tilde{i}}_{\alpha h} + \tilde{i}_{\beta h} \dot{\tilde{i}}_{\beta h} \leq 0 \quad (5.25)$$

(5.23) is substituted into (5.25), and (5.25) becomes

$$\dot{V}_2 = \frac{-R_s}{L_d}(\tilde{i}_{\alpha h}^2 + \tilde{i}_{\beta h}^2) + \frac{\tilde{i}_{\alpha h}}{L_d}(e_{\alpha h} - \mu_1 F(\tilde{i}_{\alpha h})) + \frac{\tilde{i}_{\beta h}}{L_d}(e_{\beta h} - \mu_2 F(\tilde{i}_{\beta h})) < 0 \quad (5.26)$$

To establish the stability of the proposed SMO, the observer gains μ_1 and μ_2 should be selected large enough to satisfy the following conditions at the same time:

$$\begin{aligned} \mu_1 F(\tilde{i}_{\alpha h}) &> \max(|e_{\alpha h}|), \mu_2 F(\tilde{i}_{\beta h}) > \max(|e_{\beta h}|) \text{ for the HF part} \\ \mu_1 F(\tilde{i}_\alpha) &> \max(|e_\alpha|), \mu_2 F(\tilde{i}_\beta) > \max(|e_\beta|) \text{ for the main part} \end{aligned} \quad (5.27)$$

When (5.27) is satisfied by selecting appropriate μ_1 and μ_2 , the system will reach the sliding surface. Then the following criteria providing the estimated EEMF terms are valid by using (5.20) and (5.21):

$$\begin{aligned}\hat{e}_\alpha + \hat{e}_{\alpha h} &= \mu_1 F(\tilde{i}_\alpha + \tilde{i}_{\alpha h}) \\ \hat{e}_\beta + \hat{e}_{\beta h} &= \mu_2 F(\tilde{i}_\beta + \tilde{i}_{\beta h})\end{aligned}\quad (5.28)$$

Here, F is the sigmoid function as defined in (5.5). Finally, the estimations of ω_e and θ_e are performed by substituting $\hat{e}_\alpha + \hat{e}_{\alpha h}$ and $\hat{e}_\beta + \hat{e}_{\beta h}$ in (5.28), which can be also considered as (5.19), into the angle tracking (PLL) algorithm as shown in Figure 5.3, where $(\hat{\theta}_e, \hat{\omega}_e)$ are estimations of (θ_e, ω_e) , respectively; K_1 and K_2 are positive constant gains; $\hat{\omega}_e = d\hat{\theta}_e/dt$, and $\varepsilon(t)$ denotes estimation error for the rotor electrical position and is defined as

$$\begin{aligned}(\hat{e}_\alpha + \hat{e}_{\alpha h})_n &= \frac{1}{\sqrt{(\hat{e}_\alpha + \hat{e}_{\alpha h})^2 + (\hat{e}_\beta + \hat{e}_{\beta h})^2}} (\hat{e}_\alpha + \hat{e}_{\alpha h}) \\ (\hat{e}_\beta + \hat{e}_{\beta h})_n &= \frac{1}{\sqrt{(\hat{e}_\alpha + \hat{e}_{\alpha h})^2 + (\hat{e}_\beta + \hat{e}_{\beta h})^2}} (\hat{e}_\beta + \hat{e}_{\beta h})\end{aligned}\quad (5.29a)$$

$$\varepsilon(t) = -\left[(\hat{e}_\alpha + \hat{e}_{\alpha h})_n \cos \hat{\theta}_e - (\hat{e}_\beta + \hat{e}_{\beta h})_n \sin \hat{\theta}_e\right] \quad (5.29b)$$

The estimated EEMF in (4.28) can be also given in terms of (4.19) as below:

$$\begin{aligned}\hat{e}_\alpha + \hat{e}_{\alpha h} &= -(\Gamma + \Gamma_h) \sin \theta_e \\ \hat{e}_\beta + \hat{e}_{\beta h} &= (\Gamma + \Gamma_h) \cos \theta_e\end{aligned}\quad (5.30)$$

Substituting (5.30) into (5.29) results in

$$\varepsilon(t) = (\sin \theta_e \cos \hat{\theta}_e - \cos \theta_e \sin \hat{\theta}_e) = \sin(\theta_e - \hat{\theta}_e) \cong \theta_e - \hat{\theta}_e \quad (5.31)$$

When $\varepsilon(t)$ goes to zero with the help of the implementation including the integrator and the PI controller, which are connected in series and enclosed within a unit feedback loop as seen in Figure 5.3, $\hat{\theta}_e$ converges to θ_e .

Now, the PLL's closed-loop transfer function can be given as below:

$$\frac{\hat{\theta}_e}{\theta_e} = \frac{(K_1 + K_1 K_2 s)}{s^2 + K_1 K_2 s + K_1} \quad (5.32)$$

where, K_1 and K_2 are the coefficients that affect the dynamic performance of the PLL estimator. For a stable operation with an acceptable estimation error, these coefficients should be selected according to the typical second-order system as

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (5.33)$$

where ω_n is the natural frequency and ζ is the damping factor. After determining the desired response characteristics for the general second-order system, the PLL coefficients K_1 and K_2 can be computed as $K_1 = \omega_n^2$ and $K_2 = 2\zeta/\omega_n$.

Thus, the complete diagram of the proposed voltage signal injection-based SMO and PLL to effectively predict the EEMF, rotor speed, and rotor position is given in Figure 5.3 for a reliable and accurate position sensorless control of IPMSM.

5.3 VSC System in MRAS Adaptation Mechanism for Parameter Estimation

In this section, it is aimed to evaluate the efficiency of sensorless control techniques that incorporate online estimation methods for IPMSMs in the presence of parameter variations. By analyzing various control strategies and their effectiveness in mitigating the adverse effects of parameter changes, this study seeks to provide insights into improving the reliability and performance of sensorless IPMSM drives.

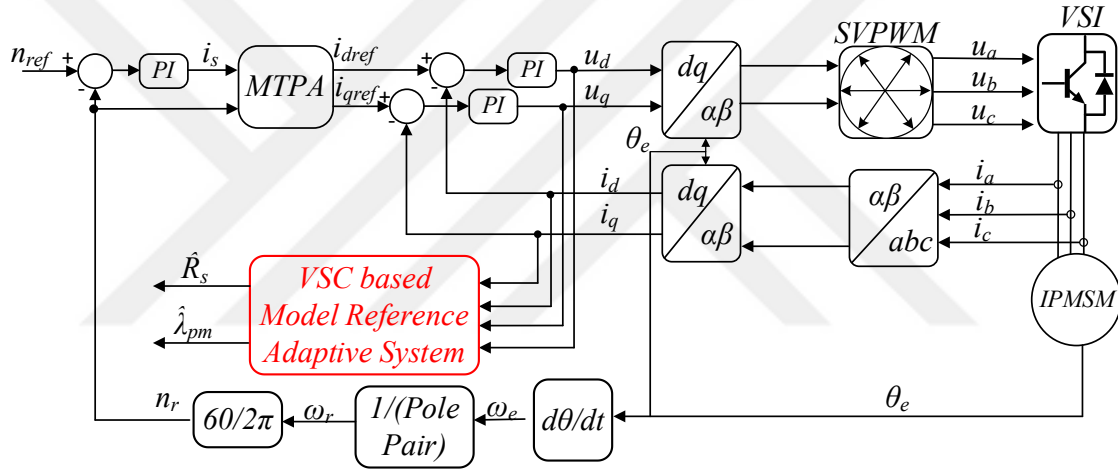


Figure 5.4 The IPMSM vector control block diagram

In this method, the rotating dq -axis is utilized and the stator voltage equation is used as the reference model.

5.3.1 The Reference Model and the Adjustable Model

An algorithm that depends on the model is the MRAS. In addition to the adaptive mechanism, it primarily consists of the adjustable model and the reference model. The adaptive model is utilized to determine the estimated states, whereas the reference model is utilized to identify the intended states. The reference and adaptive models yield actual and estimated values as their output. The adaptation mechanism receives an error signal that is generated by the difference between the output of the reference and adjustable models.

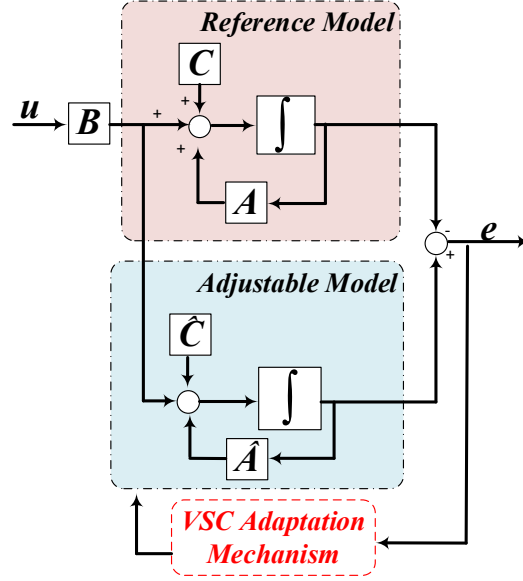


Figure 5.5 Workflow diagram of VSC-based MRAS.

To obtain the adaptive model, the rotating dq -frame of the IPMSM in (4.7) is used as the reference model and dq -axis stator currents are computed. If the reference model is written as the state space, the model can be given as:

$$\mathbf{i} = \mathbf{A}\mathbf{i} + \mathbf{B}\mathbf{u} + \mathbf{C} \quad (5.34)$$

$$\mathbf{A} = \begin{bmatrix} -\frac{R_s}{L_d} & \frac{L_q}{L_d}\omega_e \\ -\frac{L_d}{L_q}\omega_e & -\frac{R_s}{L_q} \end{bmatrix}, \mathbf{B} = \begin{bmatrix} \frac{1}{L_d} & 0 \\ 0 & \frac{1}{L_q} \end{bmatrix}, \mathbf{C} = \begin{bmatrix} 0 & -\frac{\lambda_{pm}}{L_q}\omega_e \end{bmatrix}^T$$

Where $\mathbf{i} = [i_d \ i_q]^T$ and $\mathbf{u} = [u_d \ u_q]$ represent the current state vector and voltage state vector, respectively, $\mathbf{A}$ and $\mathbf{B}$ are state coefficient matrix and input coefficient matrix respectively and $\mathbf{C}$ is the constant term matrix.

The dq -axis stator currents in the current state model of the IPMSM, designed as an adjustable model, are given by:

$$\dot{\mathbf{i}} = \hat{\mathbf{A}}\mathbf{i} + \mathbf{B}\mathbf{u} + \hat{\mathbf{C}} \quad (5.35)$$

$$\hat{\mathbf{A}} = \begin{bmatrix} -\frac{\hat{R}_s}{L_d} & \frac{L_q}{L_d}\omega_e \\ -\frac{L_d}{L_q}\omega_e & -\frac{\hat{R}_s}{L_q} \end{bmatrix}, \mathbf{B} = \begin{bmatrix} \frac{1}{L_d} & 0 \\ 0 & \frac{1}{L_q} \end{bmatrix}, \hat{\mathbf{C}} = \begin{bmatrix} 0 & -\frac{\hat{\lambda}_{pm}}{L_q}\omega_e \end{bmatrix}^T$$

Where $\hat{\mathbf{i}} = [\hat{i}_d \ \hat{i}_q]^T$ are the observed currents. The equation for the dynamic current error is obtained by subtracting (5.35) from (5.34)

$$\dot{\mathbf{e}} = \mathbf{A}\mathbf{e} + \boldsymbol{\zeta} \quad (5.36)$$

$$\mathbf{e} = \begin{bmatrix} e_d \\ e_q \end{bmatrix}, \quad \boldsymbol{\zeta} = \begin{bmatrix} \frac{\hat{R}_s - R_s}{L_d} & 0 \\ 0 & \frac{\hat{R}_s - R_s}{L_q} \end{bmatrix} \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} + \begin{bmatrix} 0 \\ \omega_e \frac{(\hat{\lambda}_{pm} - \lambda_{pm})}{L_q} \end{bmatrix}$$

where the variable $\mathbf{e}$ denotes the difference between the measured currents and the estimated currents.

The error dynamics are established through feedforward propagation combined with a nonlinear feedback mechanism. The system achieves hyperstability if the system matrix is strictly positive and the input-output mechanism satisfies Popov's integral inequality, as outlined in (5.37).

$$\int_0^{t_1} \mathbf{e}^T \boldsymbol{\zeta} dt \geq -\gamma_0^2$$

$$\mathbf{e}^T \boldsymbol{\zeta} = [e_d \quad e_q] \begin{bmatrix} \frac{(\hat{R}_s - R_s)\hat{i}_d}{L_d} \\ \frac{(\hat{R}_s - R_s)\hat{i}_q}{L_q} + \omega_e \frac{(\hat{\lambda}_{pm} - \lambda_{pm})}{L_q} \end{bmatrix} \quad (5.37)$$

Where γ_0^2 represents a positive finite number for all $t_1 \geq 0$. If $\mathbf{e}$ and $\boldsymbol{\zeta}$ substituted into the Popov's integral inequality.

$$\int_0^{t_1} \left[\frac{e_d \hat{i}_d}{L_d} + \frac{e_q \hat{i}_q}{L_q} \right] (\hat{R}_s - R_s) dt + \int_0^{t_1} \frac{e_q \omega_e}{L_q} (\hat{\lambda}_{pm} - \lambda_{pm}) dt \geq -\gamma_0^2 \quad (5.38)$$

As can be seen in (5.38), two distinct sources of parameter error are included. One arising from stator resistance and the other from flux linkage. These errors can be expressed as two separate input-output inner products[78], each corresponding to its specific parameter error:

$$\int_0^{t_1} \mathbf{e}^T \hat{\mathbf{i}} \mathbf{L}_{dq} (\hat{R}_s - R_s) dt \geq -\gamma_1^2$$

$$\int_0^{t_1} \mathbf{e}^T \mathbf{L}_{dq} \begin{bmatrix} \omega_e \\ 0 \end{bmatrix} (\hat{\lambda}_{pm} - \lambda_{pm}) dt \geq -\gamma_2^2 \quad (5.39)$$

Where γ_1^2 and γ_2^2 are finite positive constants. $\mathbf{L}_{dq} = \begin{bmatrix} \frac{1}{L_d} & 0 \\ 0 & \frac{1}{L_q} \end{bmatrix}$ is the inductance matrix.

5.3.2 Proposed Variable Structured Control Method for MRAS Adaptation Mechanism

Standard MRAS schemes often employ PI regulators in their adaptation mechanisms to accurately predict the target variable by reducing the error between the adjustable and reference models. However, PI regulators demand an exact mathematical model and continuous tuning of the gains. Moreover, they are highly susceptible to changes in operating conditions. For these reasons, this study aims to develop a variable structured control method that integrates the robustness and quick response of sliding mode control with the effectiveness and ease of use of PI controllers [79]. In this method,

the error, defined as the difference between the outputs of the adjustable model and the reference model, can serve as a switching surface for estimating the parameters. The basic equations of the variable structured control method are given as:

$$x_1 = K_1 |\tilde{x}_1|^{\frac{1}{2}} \text{sgn}(\tilde{x}_1) + \int_0^t K_1 K_2 |\tilde{x}_1|^{\frac{1}{2}} \text{sgn}(\tilde{x}_1) d\tau \quad (5.40)$$

where $x_1, \tilde{x}_1$, represent the state variables and the error between the actual and the estimated state variables respectively, K_1, K_2 are controller coefficients, and sgn is the signum function. Accordingly, the switching surfaces for stator resistance R_s and PM flux linkage are updated and expressed as follows:

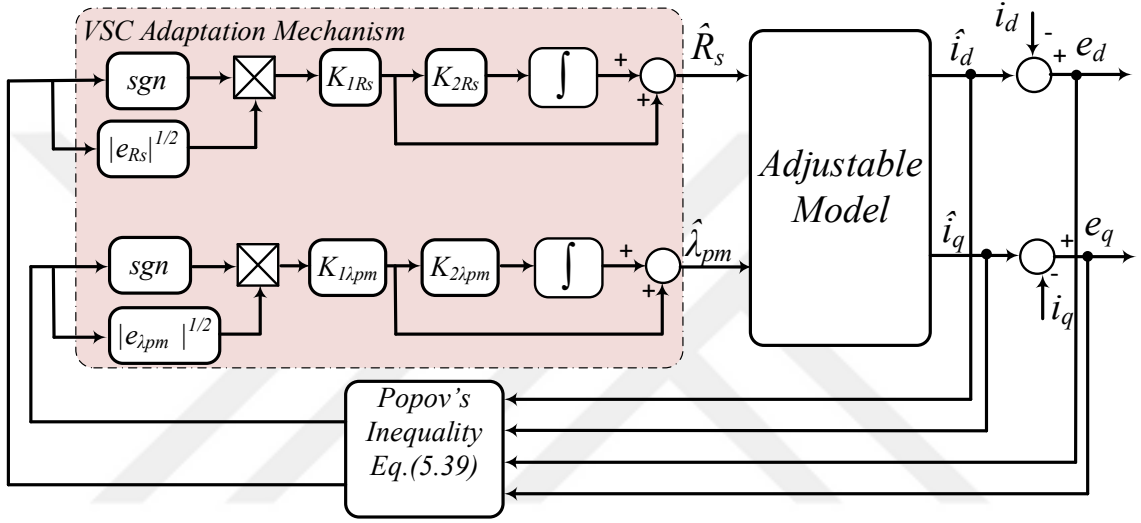


Figure 5.6 Proposed VSC Adaptation Mechanism Scheme

$$\hat{R}_s = K_{1R_s} |e^T \hat{\mathbf{L}}_{dq}|^{\frac{1}{2}} \text{sgn}(e^T \hat{\mathbf{L}}_{dq}) + \int_0^t K_{1R_s} K_{2R_s} |e^T \hat{\mathbf{L}}_{dq}|^{\frac{1}{2}} \text{sgn}(e^T \hat{\mathbf{L}}_{dq}) d\tau + \hat{R}_s(0) \quad (5.41)$$

$$\hat{\lambda}_{pm} = K_{1\lambda_{pm}} |e^T \mathbf{L}_{dq} \begin{bmatrix} \omega_e \\ 0 \end{bmatrix}|^{\frac{1}{2}} \text{sgn}(e^T \mathbf{L}_{dq} \begin{bmatrix} \omega_e \\ 0 \end{bmatrix}) + \int_0^t K_{1\lambda_{pm}} K_{2\lambda_{pm}} |e^T \mathbf{L}_{dq} \begin{bmatrix} \omega_e \\ 0 \end{bmatrix}|^{\frac{1}{2}} \text{sgn}(e^T \mathbf{L}_{dq} \begin{bmatrix} \omega_e \\ 0 \end{bmatrix}) d\tau + \hat{\lambda}_{pm}(0) \quad (5.42)$$

Proof: For the proposed controller, the nonlinear part is $|\tilde{x}_1|^{\frac{1}{2}} \text{sgn}(\tilde{x}_1)$. If the nonlinear part of the equation in (5.40) is defined as $\phi(t)$, the controller can be rearranged as:

$$x_1 = K_1 \phi(t) + \int_0^t K_1 K_2 \phi(\tau) d\tau \quad (5.43)$$

By identifying $\phi(t)$ as a nonlinear part, it is possible to arrange the system as a linear model, which is a key step in utilizing Popov's criterion.

If the Popov's integral inequality in (5.39) is rearranged for stator resistance and PM flux estimation according to (5.41) and (5.42):

$$\int_0^{t_1} \phi_{R_s}(t) \left(K_{1R_s} \phi_{R_s}(t) + \int_0^t K_{1R_s} K_{2R_s} \phi_{R_s}(\tau) d\tau + \hat{R}_s(0) - R_s \right) dt$$

$$\int_0^{t_1} \phi_{\lambda_{pm}}(t) \left(K_{1\lambda_{pm}} \phi_{\lambda_{pm}}(t) + \int_0^t K_{1\lambda_{pm}} K_{2\lambda_{pm}} \phi_{\lambda_{pm}}(\tau) d\tau + \hat{\lambda}_{pm}(0) - \lambda_{pm} \right) dt \quad (5.44)$$

$$\phi_{R_s}(t) = |\mathbf{e}^T \hat{\mathbf{L}}_{dq}|^{\frac{1}{2}} \text{sgn}(\mathbf{e}^T \hat{\mathbf{L}}_{dq})$$

$$\phi_{\lambda_{pm}}(t) = |\mathbf{e}^T \mathbf{L}_{dq} [\omega_e]_0|^{\frac{1}{2}} \text{sgn}(\mathbf{e}^T \mathbf{L}_{dq} [\omega_e]_0) \quad (5.45)$$

The expressions $\phi_{R_s}(t)$ and $\phi_{\lambda_{pm}}(t)$ represent the nonlinear parts of (5.41) and (5.42) respectively. Firstly, if the inequality model of stator resistance estimation is considered:

$$\psi_1(0, t_1) = \int_0^{t_1} \phi_{R_s}(t) \left(\int_0^t K_{1R_s} K_{2R_s} \phi_{R_s}(\tau) d\tau + \hat{R}_s(0) - R_s \right) dt$$

$$\psi_2(0, t_1) = \int_0^{t_1} \phi_{R_s}(t) (K_{1R_s} \phi_{R_s}(t)) dt \quad (5.46)$$

For $\psi_1(0, t_1)$, it can be defined as

$$K_{1R_s} K_{2R_s} f(t) = \int_0^t K_{1R_s} K_{2R_s} \phi_{R_s}(\tau) d\tau + \hat{R}_s(0) - R_s \quad (5.47)$$

where $f(t)$ is a time-dependent function. If taking the derivation of (5.47), it can be obtained as:

$$\frac{df(t)}{dt} = \phi_{R_s}(t) \quad (5.48)$$

After that, the inequality can be rearranged as:

$$\int_0^{t_1} \frac{df(t)}{dt} K_{1R_s} K_{2R_s} f(t) dt = \frac{K_{1R_s} K_{2R_s}}{2} [f^2(t_1) - f^2(0)] \geq -\frac{K_{1R_s} K_{2R_s}}{2} f^2(0) \quad (5.49)$$

It can be seen that $\psi_1(0, t_1) \geq -\gamma_0^2$.

For $\psi_2(0, t_1)$, the equation is

$$\psi_2(0, t_1) = K_{1R_s} \int_0^{t_1} (\phi_{R_s}(t))^2 dt \geq 0 \quad (5.50)$$

that is $\psi_2(0, t_1) \geq -\gamma_0^2$. In a similar manner, by using the second part of the (5.39), the PM flux linkage can be proved and estimated based on (5.42).

Thus, by using the adaptive law for MRAS as expressed in (5.41) and (5.42), it can be determined that the proposed adaptation mechanism satisfies the conditions of Popov's integral inequality stated in (5.37).

5.4 Conclusion

This chapter presented a novel sensorless control strategy for Interior Permanent Magnet Synchronous Motors (IPMSMs), addressing the critical challenges of the position of the rotor and speed estimation, particularly at low speeds and standstill. By leveraging high-frequency voltage injection and a SMO integrated with a PLL, the proposed method demonstrated robustness and accuracy in real-time estimation of motor parameters.

The chapter outlined the limitations of conventional sensorless control techniques and detailed how the integration of SMO and PLL overcomes these challenges by enhancing noise resilience and parameter adaptability. The ability of the suggested method to maintain precise rotor position and speed estimation, even under varying operating conditions and parameter changes, highlights its effectiveness for practical applications.

This innovative approach forms the basis for a more reliable, efficient, and computationally feasible sensorless control system, paving the way for its implementation in diverse industrial and automotive contexts. The insights and developments from this chapter set the stage for the experimental validation and performance analysis discussed in the following sections, providing a strong foundation for advancing the state of sensorless motor control technology.

Chapter 6

Simulation and Experimental Results

6.1 Introduction

This chapter explores the simulation and experimental validation of the proposed sensorless control strategy for IPMSMs. Building on the theoretical framework and control methods developed in earlier chapters, this chapter focuses on evaluating the real-world performance of the HF voltage injection-based SMO combined with a PLL.

The chapter begins by describing the simulation setup, detailing the motor parameters, control algorithms, and operating conditions utilized to evaluate the effectiveness of the proposed method. The motor parameters include specifications such as inductance, resistance, and flux linkage, while the control algorithms cover sensorless techniques and parameter estimation strategies tailored to various operating conditions. Key performance metrics, such as accuracy in rotor position and speed estimation, robustness to noise and parameter variations, and computational efficiency, are analyzed under diverse scenarios. These scenarios encompass a wide range of operating conditions, including standstill, low-speed, and high-speed regimes, as well as transient states like acceleration and deceleration. The analysis also includes tests under varying load conditions, ensuring the method's reliability and adaptability in real-world applications. Additionally, the influence of external disturbances, such as sensor noise and sudden parameter changes, is examined to assess the robustness and stability of the control algorithms. Detailed comparisons with existing methods are presented, highlighting the proposed method's advantages in terms of precision, adaptability, and computational cost.

Following the simulation results, the chapter transitions to experimental validation using an IPMSM test bench. The experimental setup, including hardware components, instrumentation, and testing protocols, is thoroughly documented to ensure reproducibility. The comparison between simulation and experimental outcomes not only

validates the proposed method but also highlights its practical feasibility and advantages over conventional techniques.

Through a combination of theoretical rigor and practical application, Chapter 6 demonstrates the capability of the suggested sensorless control system to meet the demands of industrial and automotive applications. This validation process provides critical insights into the robustness, efficiency, and versatility of the approach, paving the way for its deployment in real-world scenarios.

6.2 Simulation Results

The simulation results of the proposed sensorless control strategy for IPMSM are presented and analyzed. These simulations serve as a critical step in validating the theoretical framework and control algorithms developed in earlier chapters. By modeling the dynamic behavior of the motor and implementing the HF voltage injection-based SMO integrated with PLL, the simulations aim to evaluate the effectiveness of the proposed method under diverse operating conditions.

The VSC adaptation mechanism for parameter estimation is also incorporated into the simulation framework. This mechanism is designed to adapt dynamically to changes in motor parameters such as resistance, and PM flux, which can vary due to temperature fluctuations, magnetic saturation, or aging. By integrating this method into the sensorless control strategy, the simulations assess the system's ability to maintain accuracy and stability even in the presence of significant parameter variations.

The MTPA performance of the IPMSM during simulation is given in Figure 6.1. For the simulation part, the real-time parameters of IPMSM given in Table 6.1 are utilized. The results provide key insights into the motor's operational efficiency and dynamic performance under varying speed conditions. For speed tracking performance, the blue plot represents the commanded speed and the red plot represents the measured speed. According to this plot, the measured speed closely follows the commanded speed across all scenarios, including acceleration, deceleration, and reversals. The other plots display the d-axis and q-axis currents respectively. These currents are regulated in a manner that achieves the MTPA operating point, minimizing current consumption while maximizing torque production. The currents remain within expected limits, demonstrating efficient control of torque and flux. The currents exhibit well-defined transitions during speed changes, showing minimal oscillations or instability. This

behavior reflects the robustness of the control algorithm in dynamically adjusting to new operating points while maintaining stability.

The effectiveness of the proposed sensorless control and parameter estimation methods is evaluated under various challenging scenarios across a broad speed range, including standstill and extremely low speeds, as detailed below:

Scenario 1: proposed SMO operation in Simulation

Firstly, the speed estimation performance of the SMO for the IPMSM, with the actual rotor speed compared to the estimated rotor speed under variable loaded conditions is shown in Figure 6.2. The estimated speed shown with red dashed line closely tracks the measured speed across the entire range of operation, including acceleration, steady-state, deceleration, and reversals. This indicates that the SMO provides accurate estimation, a crucial factor in achieving robust sensorless control.

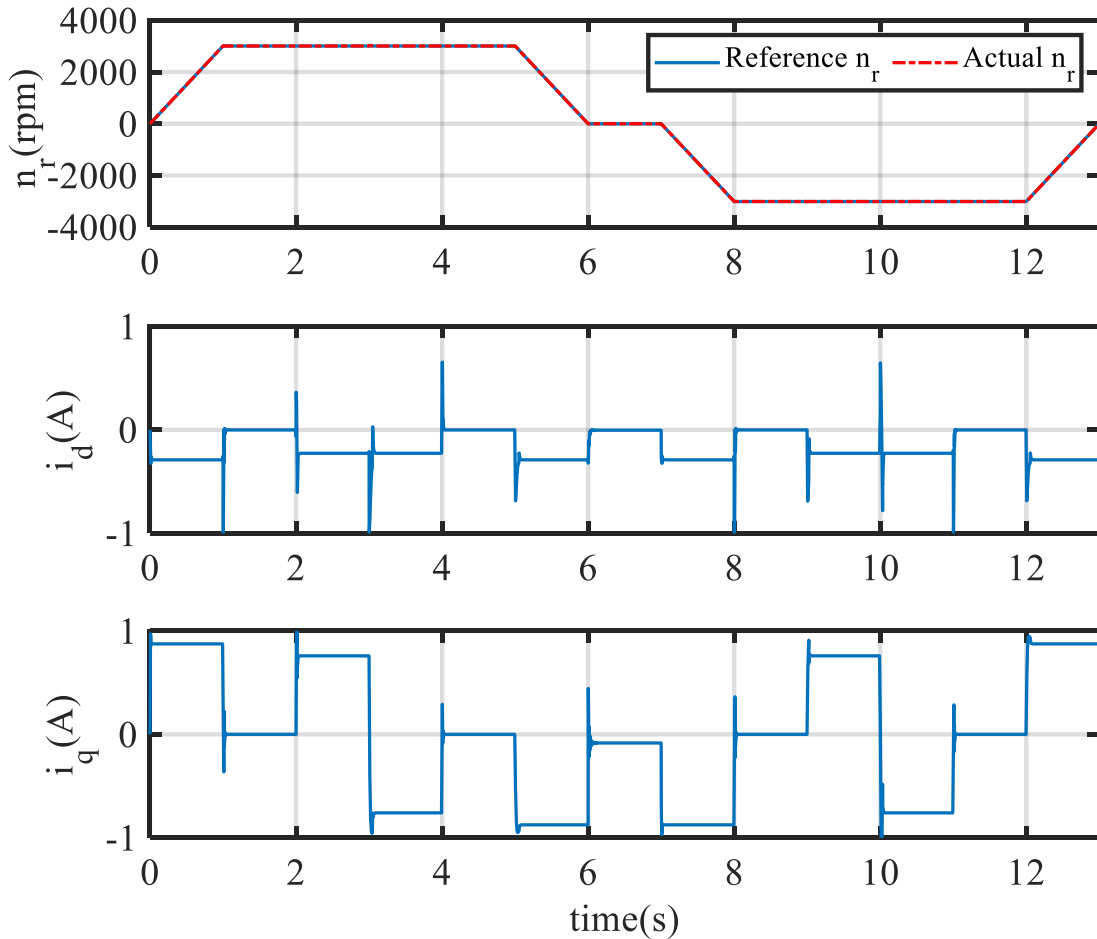


Figure 6.1 MTPA Performance of the IPMSM drive

During speed transitions, the SMO shows minimal lag or overshoot, demonstrating its ability to handle rapid changes effectively. This highlights the observer's capability to maintain reliable estimation even under dynamic conditions.

The results confirm the effectiveness of the SMO in providing precise and reliable rotor speed estimation for IPMSM control. Its robustness to dynamic changes, low-speed operation, and load disturbances makes it an excellent choice.

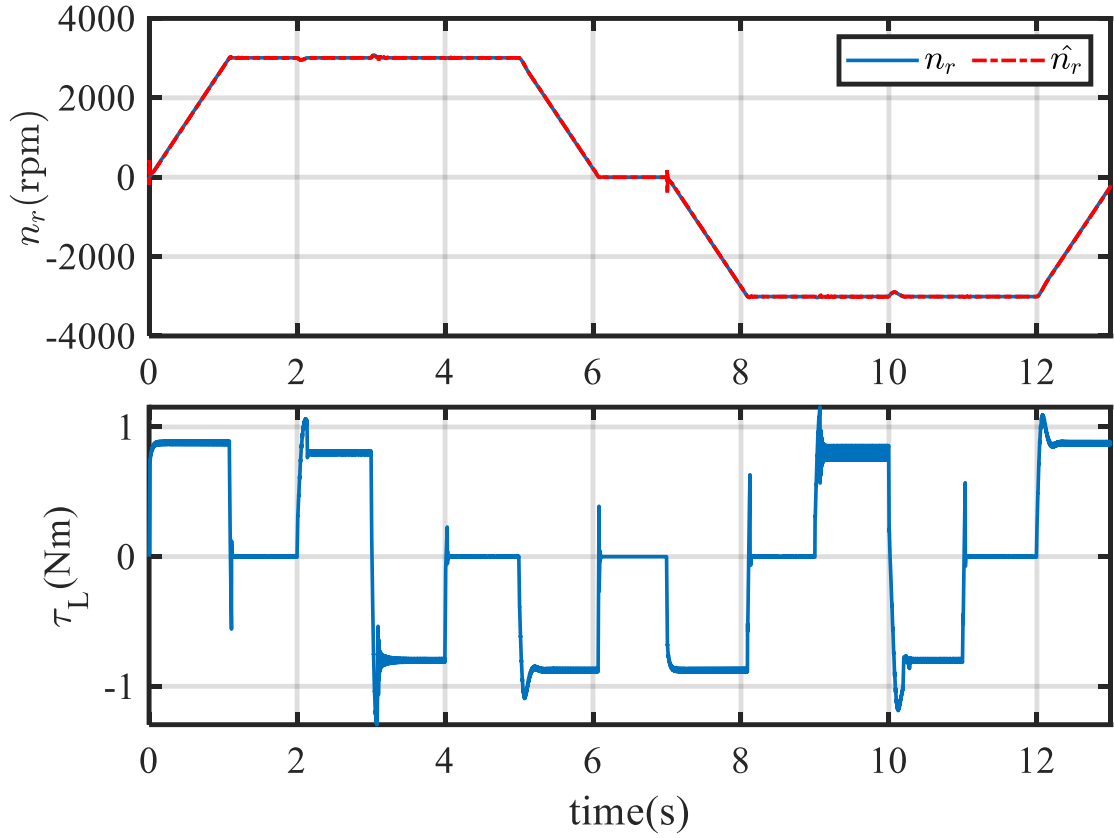
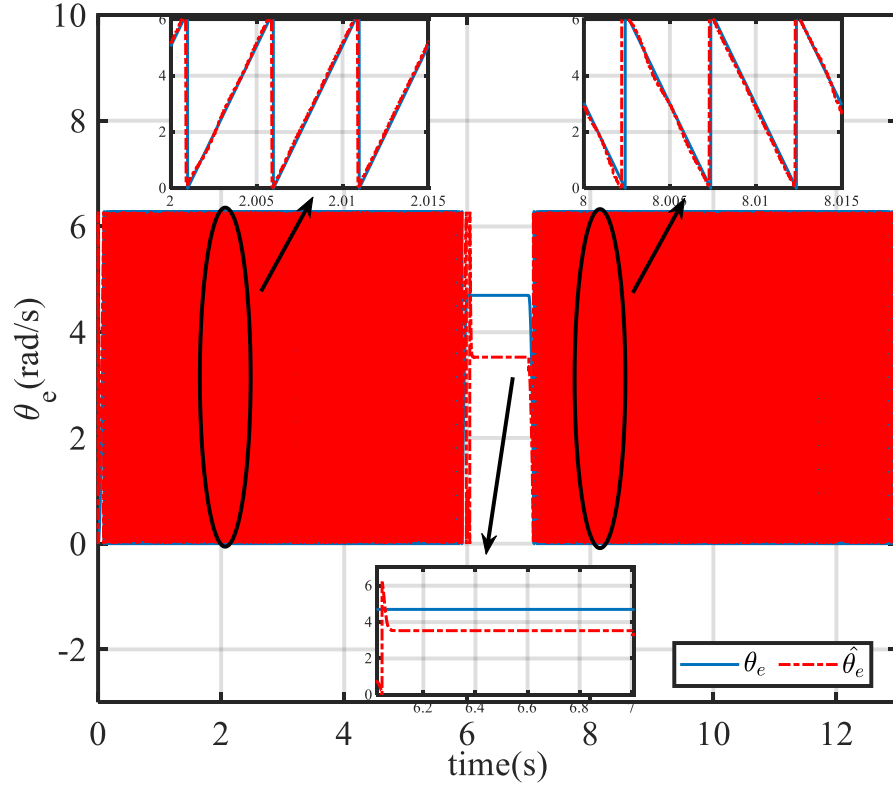
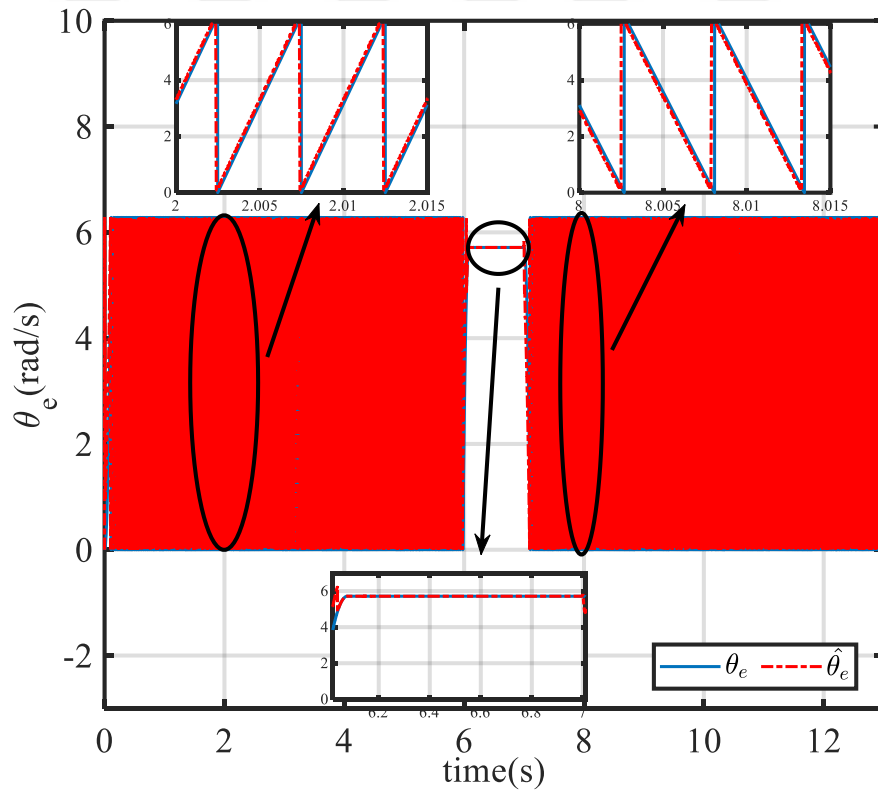


Figure 6.2 SMO Speed Estimation Performance in Variable Load Condition

The position estimation performances of the conventional SMO and the proposed SMO are given in Figure 6.3. The results show that the proposed SMO significantly improves the position estimation accuracy, with the estimated position closely matching the actual rotor position across all operating conditions. Especially at low speeds and standstill, the proposed SMO maintains accurate position estimation, unlike the conventional SMO. The integration of HF injection enables precise estimation even in scenarios where traditional methods fail. The enhancements incorporated into the proposed SMO using HF injection and improved noise handling contribute to its superior performance. The proposed SMO's ability to estimate rotor position reliably under diverse operating conditions makes it a robust solution for sensorless control applications, addressing the limitations of the conventional approach.



(a)



(b)

Figure 6.3 Position estimation performance of (a) the conventional SMO and (b) the proposed SMO

Scenario 2: Proposed MRAS operation and compared with conventional MRAS in Simulation

The comparison of the parameter estimation performance of the PI adaptation mechanism and the proposed VSC adaptation mechanism for the MRAS is introduced in this scenario. The estimation performance is evaluated for R_s and λ_{pm} . The real R_s and λ_{pm} are shown in blue, the estimated values with the PI adaptation mechanism are shown in red, and the VSC adaptation mechanism in green. First, the estimation performances of these adaptation mechanisms are compared under rated load and constant speed conditions as shown in Figure 6.4. In this condition, both the PI and VSC mechanisms accurately estimate the R_s and λ_{pm} during rated load and constant speed conditions. However, the VSC mechanism demonstrates slightly better tracking accuracy, with smaller deviations from the actual value than the PI mechanism. As shown in the figure, the estimation errors for these parameters are consistently lower with the VSC mechanism, reflecting its robustness and higher precision.

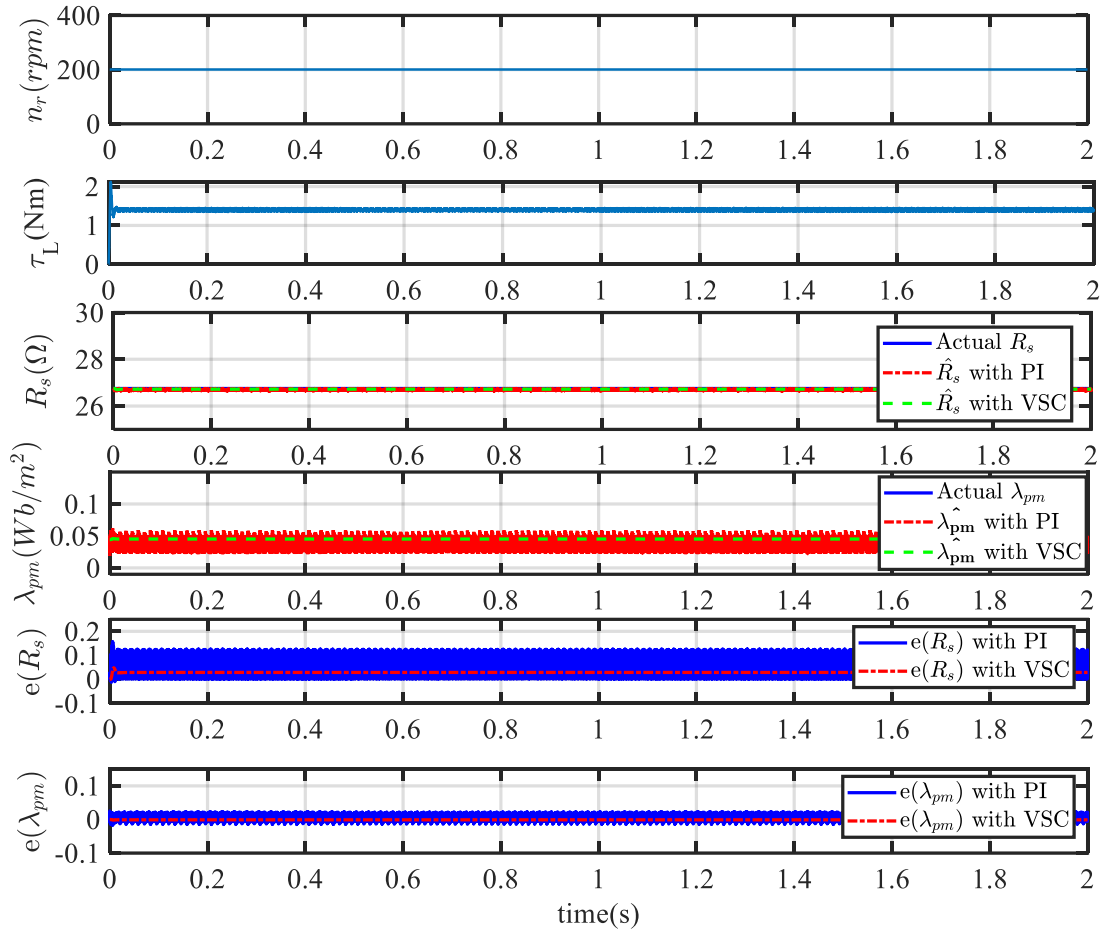


Figure 6.4 R_s and PM Flux Estimation Performance in rated Load and Constant Speed Condition

The other results compared the PI adaptation mechanism and the proposed VSC adaptation mechanism under variable load and constant speed conditions are shown in Figure 6.5. As shown in the figure, both PI and VSC mechanisms track the R_s and λ_{pm} during load variations, but VSC mechanism shows significantly better accuracy and stability. Moreover, the VSC mechanism demonstrates smoother tracking of R_s and λ_{pm} quickly adapting to load changes with minimal oscillations. If the response to load changes is observed, it is clearly seen that the VSC mechanism effectively handles the load variations, providing stable and accurate estimations without significant oscillations or delays. This indicates the VSC mechanism's robustness to variable load conditions. Finally, the estimation errors for these two parameters are substantially lower with the VSC mechanism compared to the PI mechanism. This further validates the superiority of the VSC approach under dynamic load conditions.

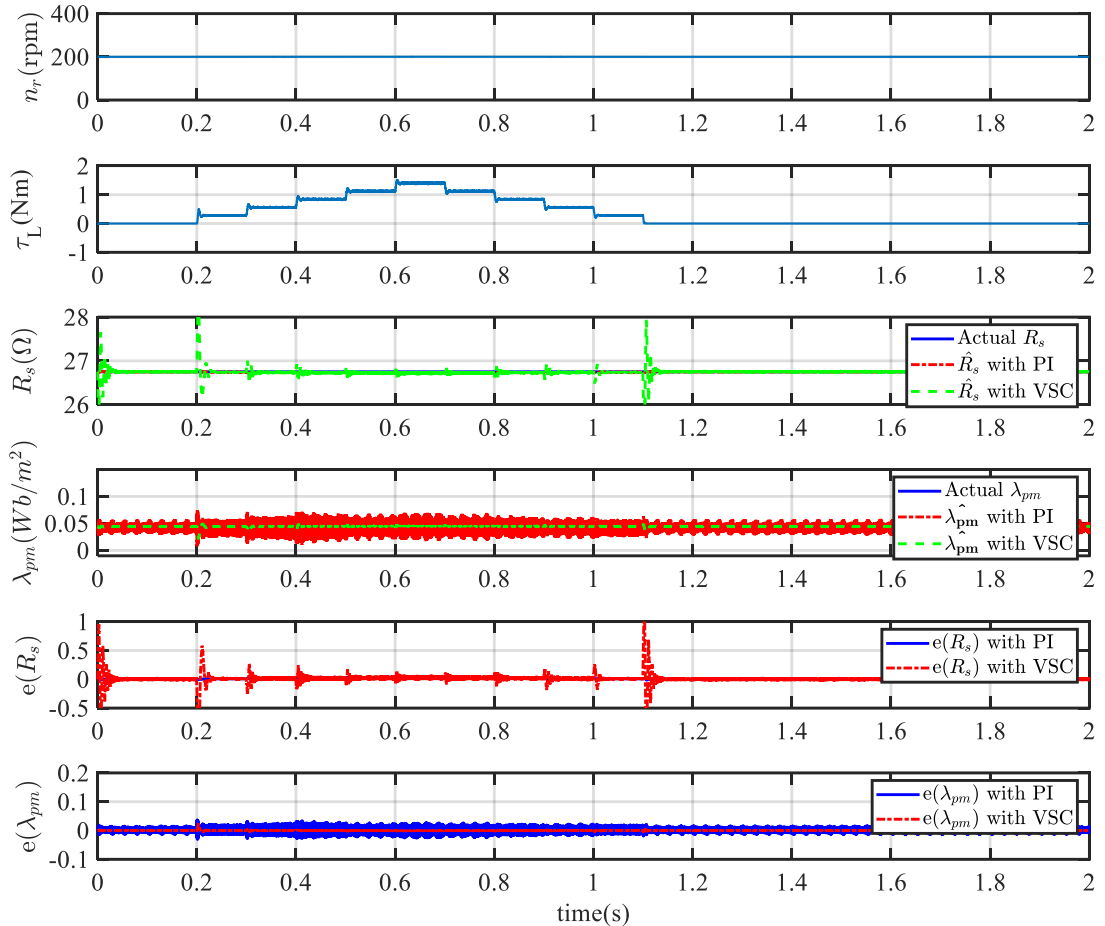


Figure 6. 5 R_s and PM flux estimation performance in variable load and constant speed conditions

The other condition is shown in Figure 6.6. In this figure, the performance of the PI and VSC adaptation mechanisms for the MRAS is compared under variable R_s , variable

load and constant speed conditions. The parameters including the estimated R_s and estimated λ_{pm} are analyzed, along with their respective errors. A step change in R_s occurs at $t=1.25$ s. the proposed VSC mechanism quickly adapts to this change, showing minimal delay and oscillation in tracking the new R_s , but the PI adaptation mechanism exhibits significant oscillations and slower convergence to this updated R_s indicating lower adaptability under dynamic R_s conditions. For λ_{pm} estimation performance, the proposed VSC mechanism provides smoother and more stable estimations. Moreover, the estimation errors show that the proposed VSC mechanism consistently achieves smaller error magnitudes and smoother behavior.

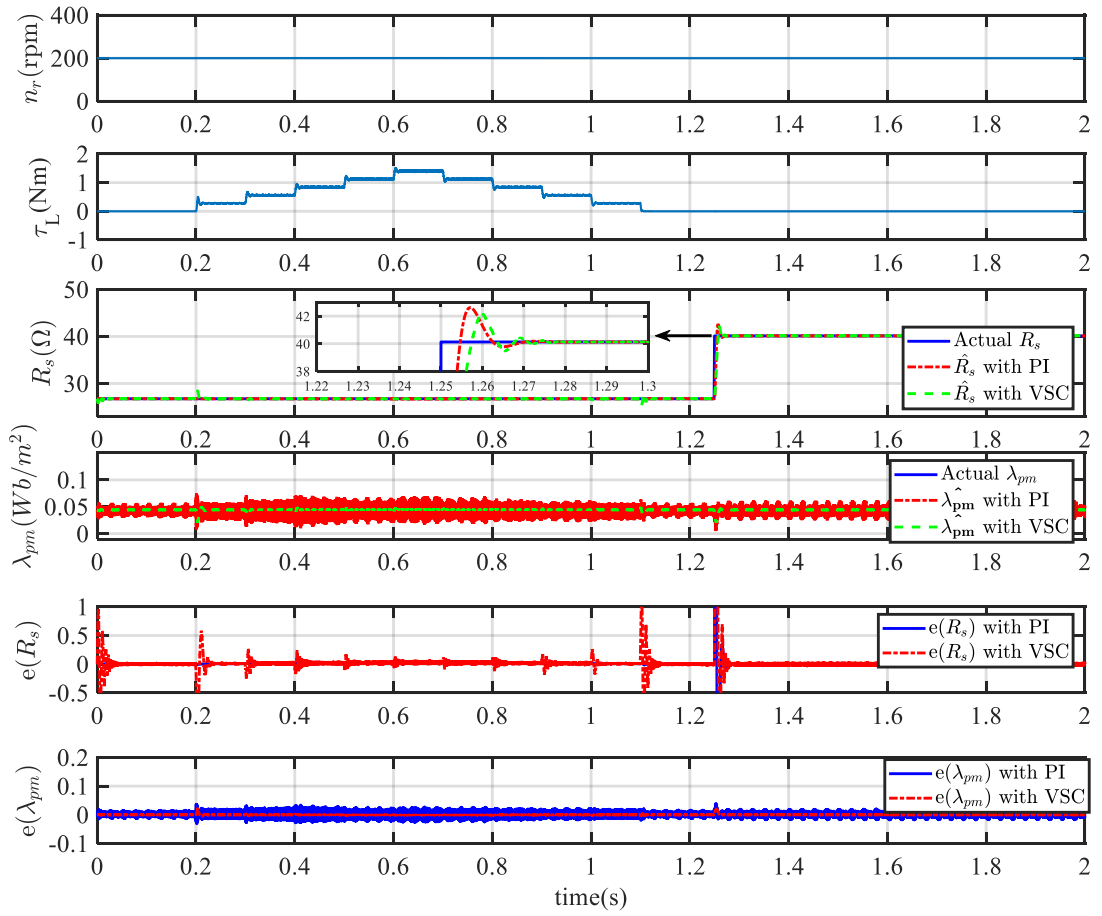


Figure 6.6 Variable R_s and PM flux estimation performance in variable load and constant speed condition

The other condition is given in Figure 6.7. In this figure, the estimation performances of these two adaptation mechanisms for the MRAS under highly dynamic operating conditions, where R_s and λ_{pm} are changing along with variable load and speed profiles. In this condition, a step change in R_s occurs in $t=1.25$ s. Moreover, a step decreasing in λ_{pm} occurs in $t=0.7$ s. In addition, the speed profile is changing from 200 to

-200 rpm in $t=0-2$ s. As shown in the figure, the proposed VSC mechanism rapidly and smoothly adapts to the changing of R_s and λ_{pm} showing minimal overshoot or oscillation. The proposed method maintains consistently higher accuracy throughout the variable speed and load conditions, closely following the real R_s and λ_{pm} . The estimation errors confirm the superior performance of the proposed VSC mechanism, with significantly lower errors compared to the conventional PI mechanism.

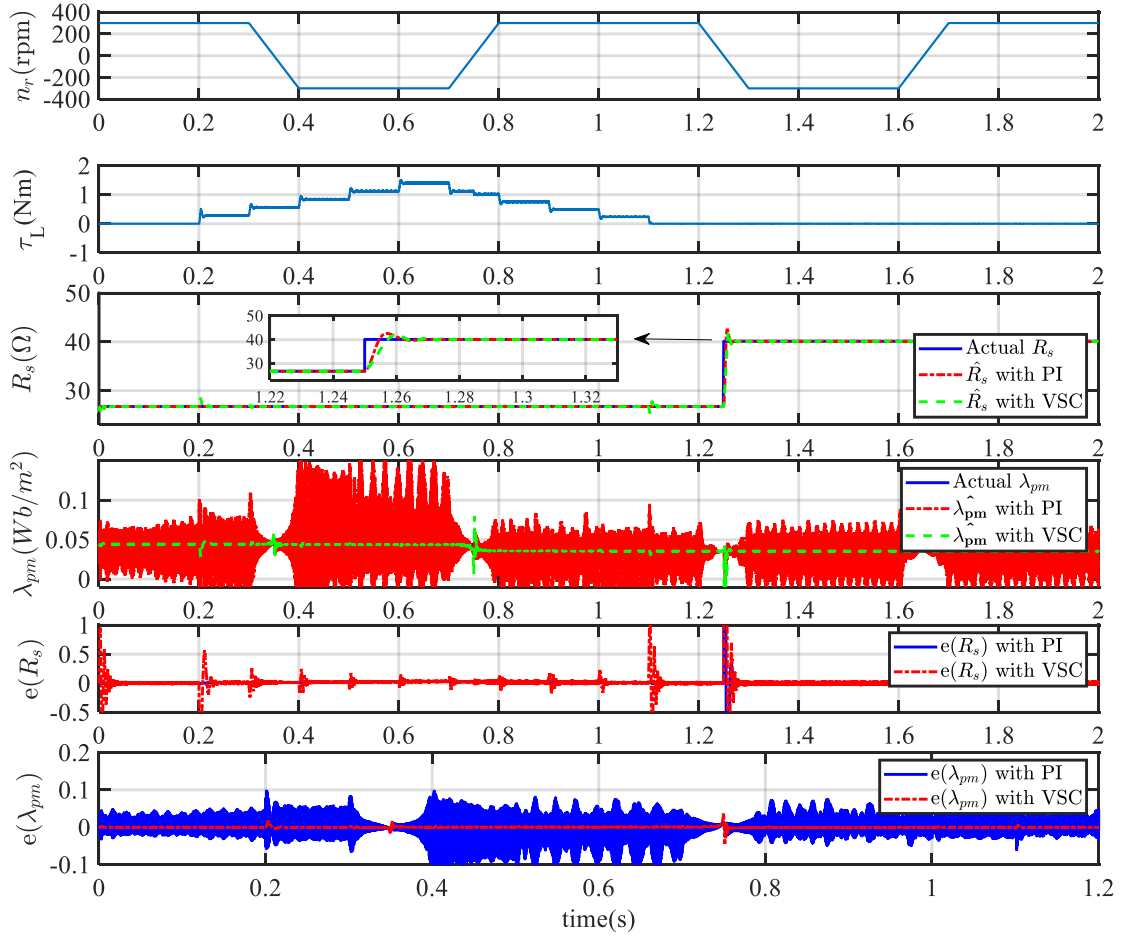


Figure 6.7 Variable R_s and PM flux estimation performance in variable load and variable speed condition

Scenario 3: SMO and MTPA performance with/without estimated parameters (estimated with proposed MRAS) in Simulation

The performance of the SMO and MTPA operations under two different conditions is given in Figure 6.8 and 6.9. In Figure 6.8, SMO and MTPA operations are applied with constant parameters, without considering parameter variations. In Figure 6.9, SMO and MTPA operations are applied with parameters adaptively estimated using the proposed VSC mechanism. When parameter variations occur on the motor, the SMO and the MTPA operations are applied with constant parameters:

- The estimated rotor position deviates significantly from the actual position, particularly as R_s and λ_{pm} change. This mismatch introduces errors, degrading the accuracy and reliability of the SMO.
- The MTPA operation is adversely affected, as the control does not adapt to the changing motor parameters. This results in suboptimal torque production and potential instability in motor performance.

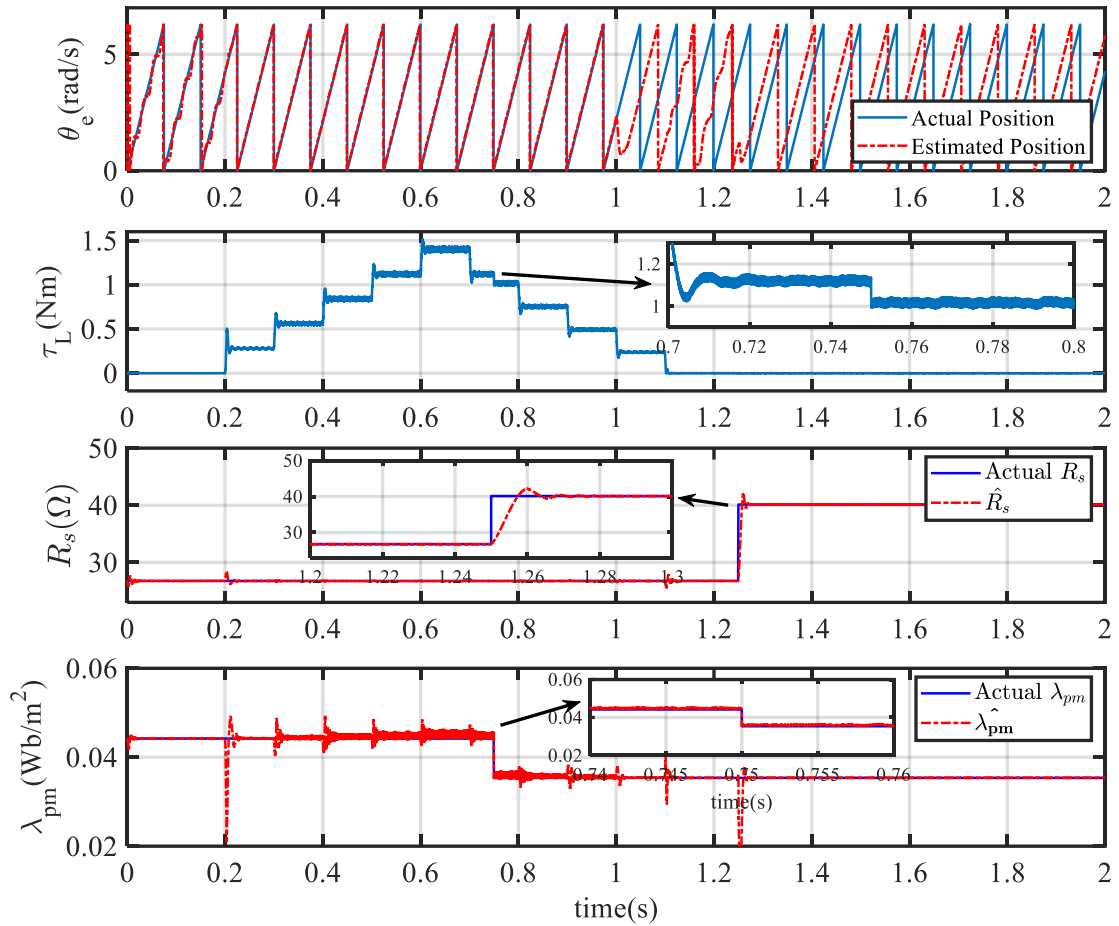


Figure 6.8 SMO and MTPA performance with constant R_s and constant λ_{pm}

If the SMO and MTPA operations are applied with adaptively estimated parameters as shown in Figure 6.9:

- The estimated rotor position closely tracks the actual position throughout the operation. This demonstrates that the VSC mechanism effectively compensates for parameter changes, enhancing the SMO's accuracy.
- The MTPA operation is significantly improved, with torque production being stable and efficient across varying conditions. The use of adaptive parameter

estimation ensures that the MTPA strategy remains optimal even as motor parameters change.

- Figure 6.9 shows a clear improvement in the accuracy of rotor position estimation and torque performance compared to Figure 6.8. this is directly attributable to the use of adaptive parameter estimation with the VSC mechanism.

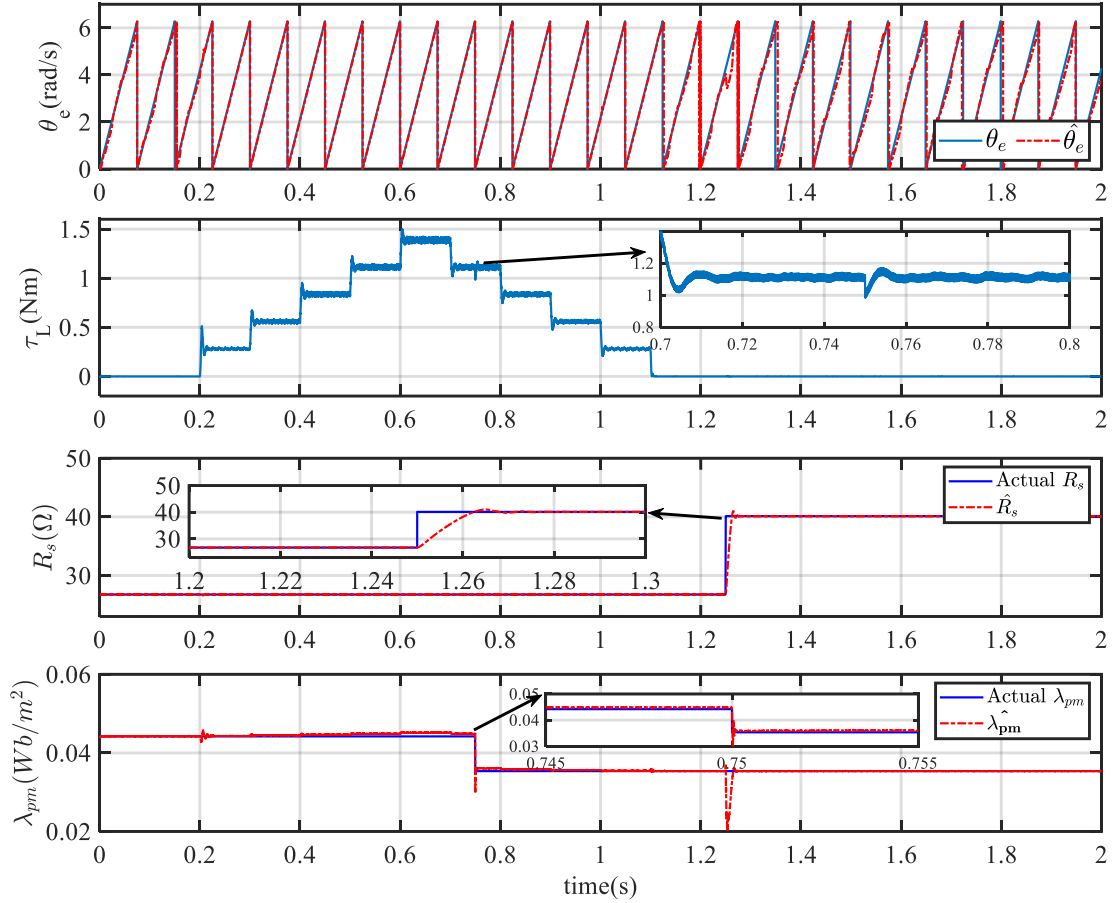


Figure 6.9 SMO and MTPA performance with adaptively estimated R_s and λ_{pm}

6.3 Design Specifications

A comprehensive evaluation of the position sensorless IPMSM drive is conducted by implementing both the proposed high-frequency (HF) voltage signal injection-based SMO with PLL and the conventional SMO in real-time, respectively. This evaluation aims to exhibit the tracking performance, stability, and robustness of the proposed SMO&PLL approach in comparison to the conventional SMO&PLL under various operating conditions, including low-speed and high-speed regions, load disturbances, and transient responses, by systematically analyzing and quantifying both steady-state and

dynamic behavior, assessing sensitivity to parameter variations and external perturbations, and ultimately providing a comprehensive assessment of their relative effectiveness in achieving precise position estimation in sensorless IPMSM drives, thereby offering valuable insights into the potential improvements and trade-offs associated with the adoption of the novel SMO&PLL scheme for advanced motor control applications. In the experimental setup,

- the parameters of the utilized IPMSM are given in Table 6.1.
- the proposed or conventional estimation method along with the vector control algorithm is embedded in a Texas Instruments of TMS320F28379D DSP that can process floating and fixed-point operations at a rate of 200 MHz.
- Texas Instruments TMDXIDDK379D Design Drive development kit is used as the inverter.
- the switching frequency of the inverter is $f_s = 10$ kHz and the control period is 100 μ sec.
- a torque transducer of 20 Nm is interposed between the IPMSM and the dynamometer for load torque measurements.
- the actual speed and rotor position are measured with a resolver integrated into the motor for validating the predicting rotor speed and position.
- the actual and estimated values of the speed and position are sent to the oscilloscope with digital analog converter outputs of the DSP for observing and data collecting.

Table 6.1 IPMSM Parameters

Parameters	Values
R_s	26.5 Ω
L_d	82 mH
L_q	118 mH
λ_{pm}	0.0042Wb
P	8
J	0.03kg.m ²

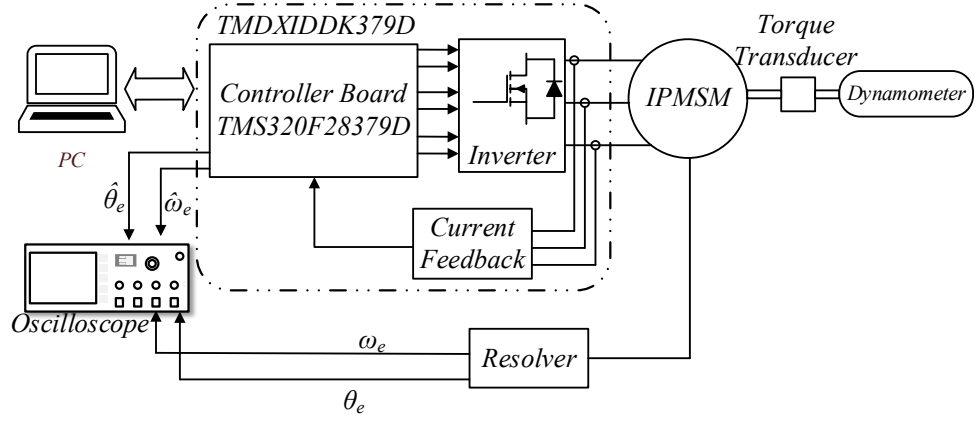


Figure 6.10 The schematic representation of the experimental setup.

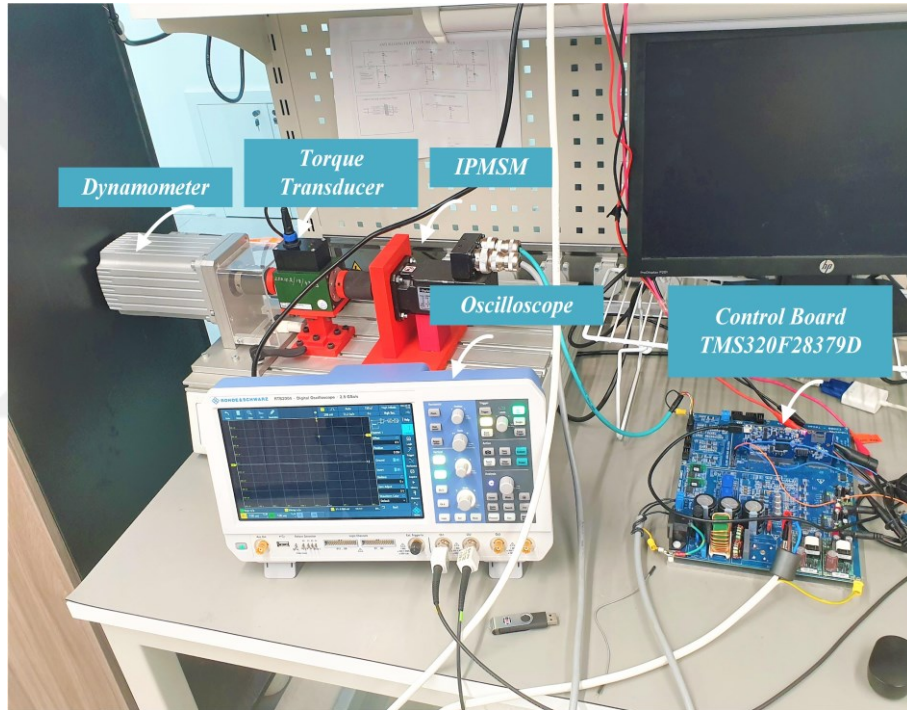


Figure 6.11 The view of the experimental setup.

6.4 Experimental Results

For the initialization of the proposed and conventional estimation algorithms,

- it is observed that the increased frequency of the inserted HF voltage decreases the effect on the accurateness of the estimations because of the reduced stator currents while the boosted amplitude of the inserted HF voltage causes the increased fluctuations in the stator current, which results in the amplified noise of the operation. Consequently, the

amplitude and the angular frequency of the inserted voltage are selected as $U_h = 25V$ and $\omega_h = 1000\pi$ rad/s, respectively, considering these effects.

- the SMO parameters are selected so that the estimation output is accurate and experiences as low chattering noise as possible; Thus, the observer gains are determined as $\mu_1 = \mu_2 = 400$ and the sigmoid coefficient is taken as $\sigma = 0.001$.

- the PLL gains are calculated using (4.26) for a non-overshoot response and around 200 Hz bandwidth, and then $\zeta = 1$ and $\omega_n = 500$ are determined; Therefore, the PLL coefficients are calculated as $K_1 = 250000$ and $K_2 = 0.0016$.

- the bus voltage is applied as $V_{DC} = 250 V$ in order to operate the experimental setup.

The current response when the suggested HF voltage injection-based SMO & PLL is applied is shown in Figure 6.12.

The effectiveness of the proposed voltage injection-based SMO&PLL are tested by using different challenging scenarios in a vast range of speeds including very low and zero speed as follows:

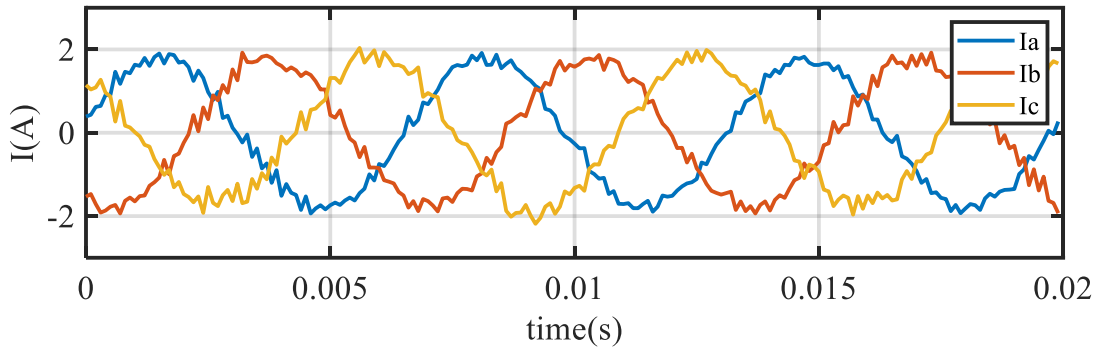


Figure 6.12 HF Current Response under loaded condition.

Scenario 1 (Figure 6.13): Performances for high-speed reversal under step-like variations in the load torque.

In this scenario, the estimation performance of the proposed voltage injection-based SMO&PLL is checked under the step-like variable load torque in a wide speed range including zero crossover in both clockwise (CW) and counter clockwise (CCW) rotations, as illustrated in Figure 6.13 where the HF voltage of $25 \cos(1000\pi t)$ is injected on d -axis at 150 rpm in order to improve the estimation performance at very low speeds and

zero crossover. Inspecting the estimation error given in Figure 6.13 (c), it is clear that the estimated velocity closely follows the actual motor speed in all operating ranges of the IPMSM for this scenario even when the speed approaches and crosses zero.

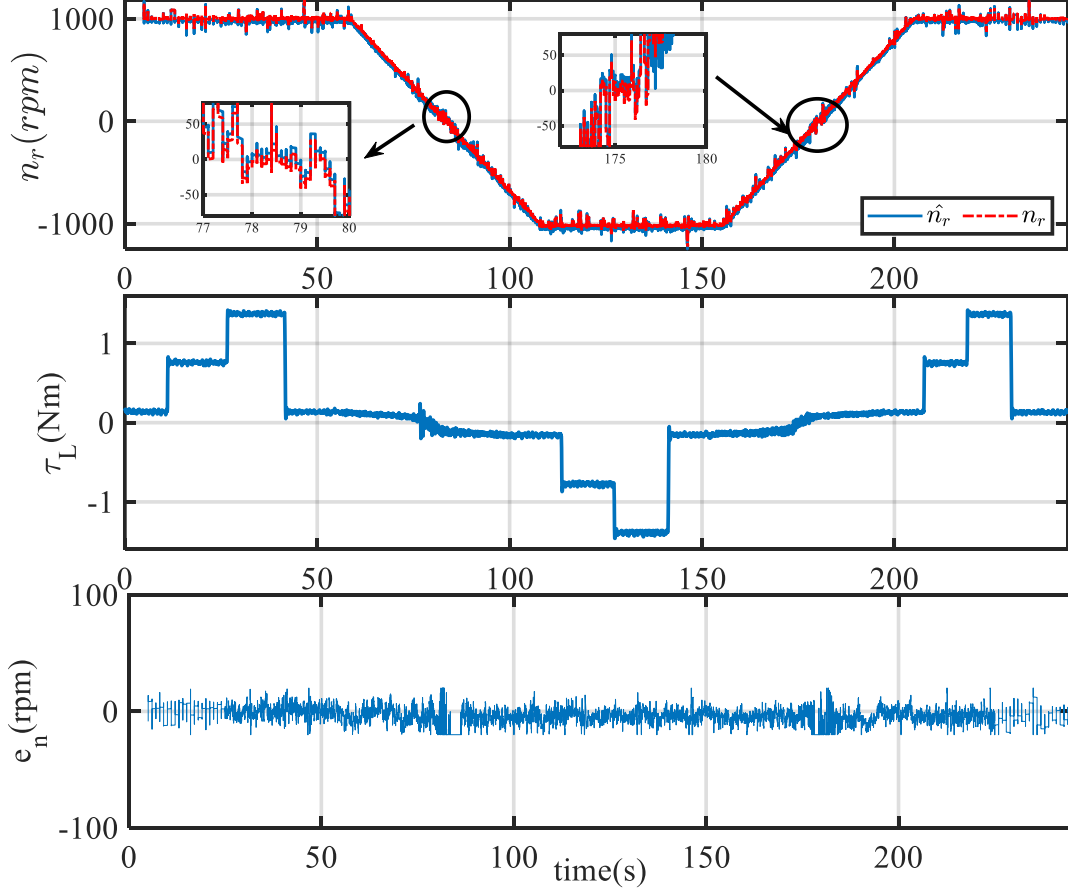


Figure 6.13 Real-time speed estimation performance of the proposed voltage injection-based SMO&PLL for Scenario 1. (a) variations of the estimated and actual speeds (b) variation of the load torque, and (c) variation of the speed estimation error.

Scenario 2 (Figure 6.14): Performances for very low and zero speed reversal under step-like variations in the load torque

This scenario aims to test the estimation performance of the proposed voltage injection-based SMO&PLL under the instant load torque variations at very low and zero speed which is the most challenging scenario for the observers using the fundamental model of IPMSM in the literature. Therefore, the IPMSM is operated by injecting the HF voltage of $25 \cos(1000\pi t)$ on d -axis at zero and 100 rpm in the directions of both CW and CCW under no-load and the instant load torque variations, as demonstrated in Figure 6.14. In spite of the most difficult scenario, the proposed HF voltage injection based SMO&PLL

is able to satisfactorily estimate the measured rotor speed, as seen in Figure 6.14 (c). It should be also noted that real-time experimental results in Figure 6.14, which are the first in the current literature associated with the position sensorless AC drive using only the fundamental model of IPMSMs, are obtained due to the voltage injection based SMO proposed by this work.

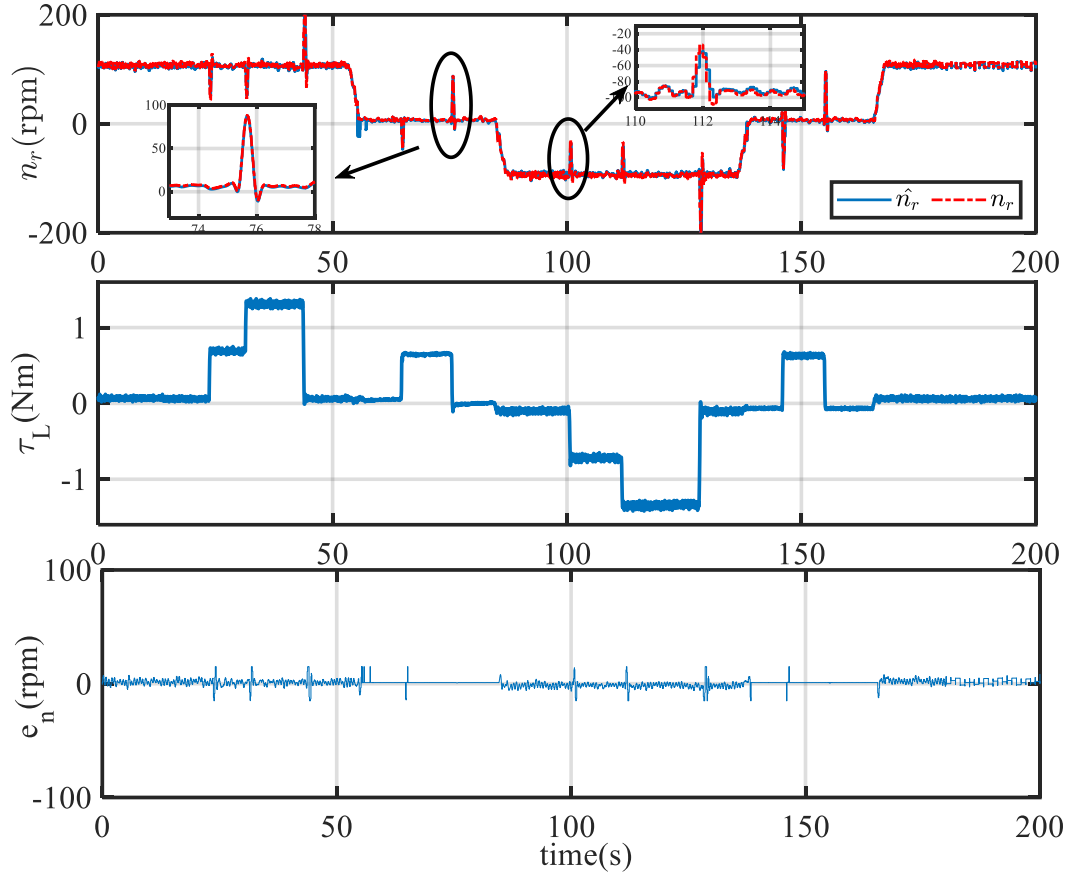
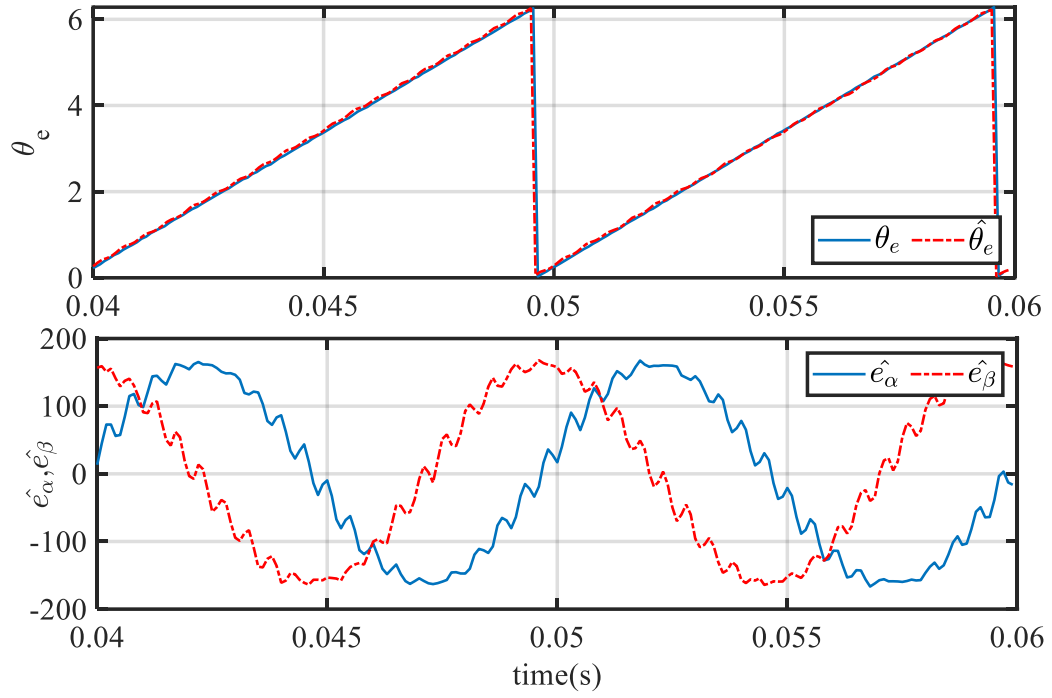


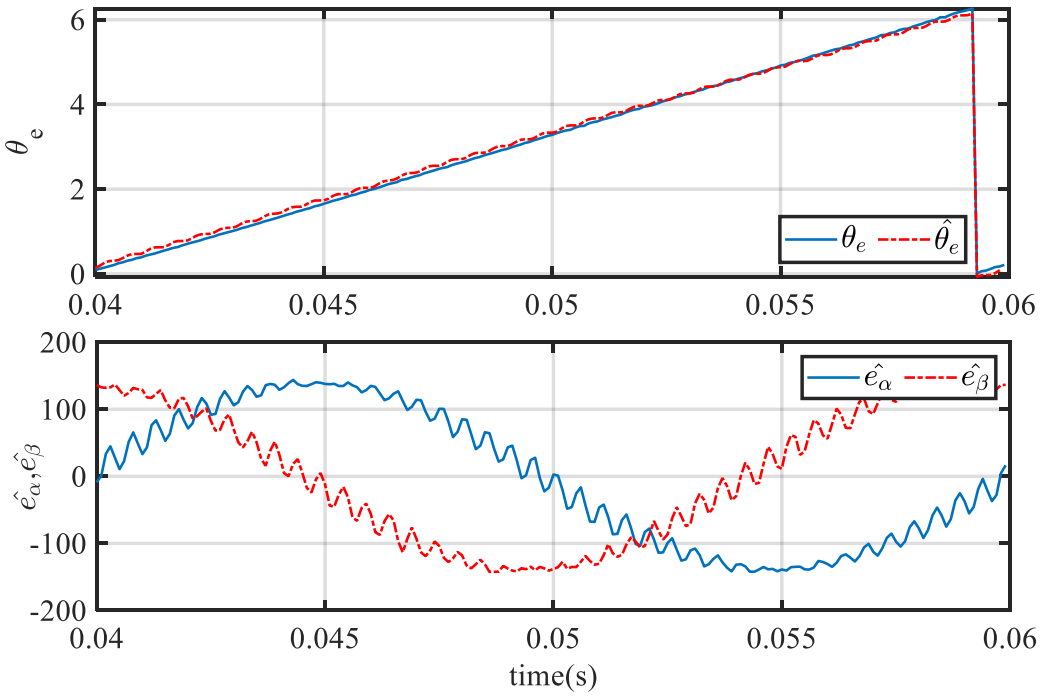
Figure 6.14 Real-time speed estimation performance of the proposed voltage injection-based SMO&PLL for Scenario 2. (a) variations of the actual and estimated speeds (b) variation of the load torque, and (c) variation of the speed estimation error.

Scenario 3 (Figure 6.15): Performances in low speed region for Estimated position and EMF under loaded condition

This scenario aims to show the estimation performance of the proposed voltage injection-based SMO&PLL at low speeds illustrating the effect of the proposed method on the estimation of electromotive force (EMF) and position. For this purpose, the IPMSM is operated by injecting the HF voltage on d -axis at 300 rpm and 150 rpm in the CW direction under loaded condition, as depicted in Figure 6.15. The obtained results indicate that the proposed voltage injection-based SMO&PLL presented in this paper is capable of accurately tracking the actual position, with acceptable estimation performance.



(a)



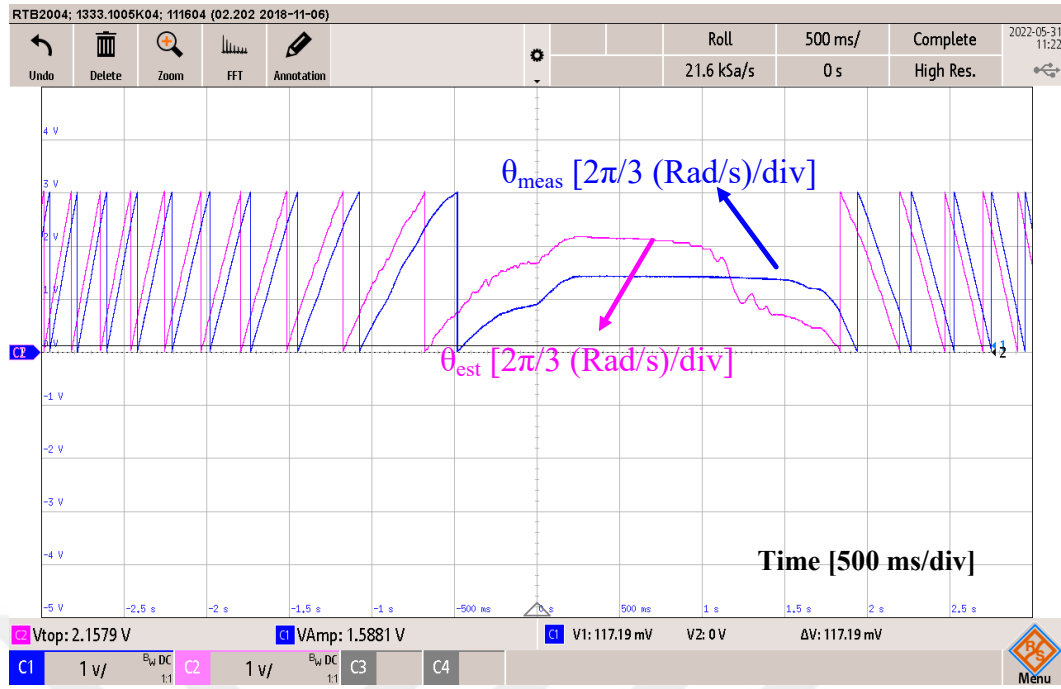
(b)

Figure 6.15 EMF and position estimation performance of the proposed voltage injection-based SMO&PLL for Scenario 3. (a) 300 rpm speed, and (b) 150 rpm speed.

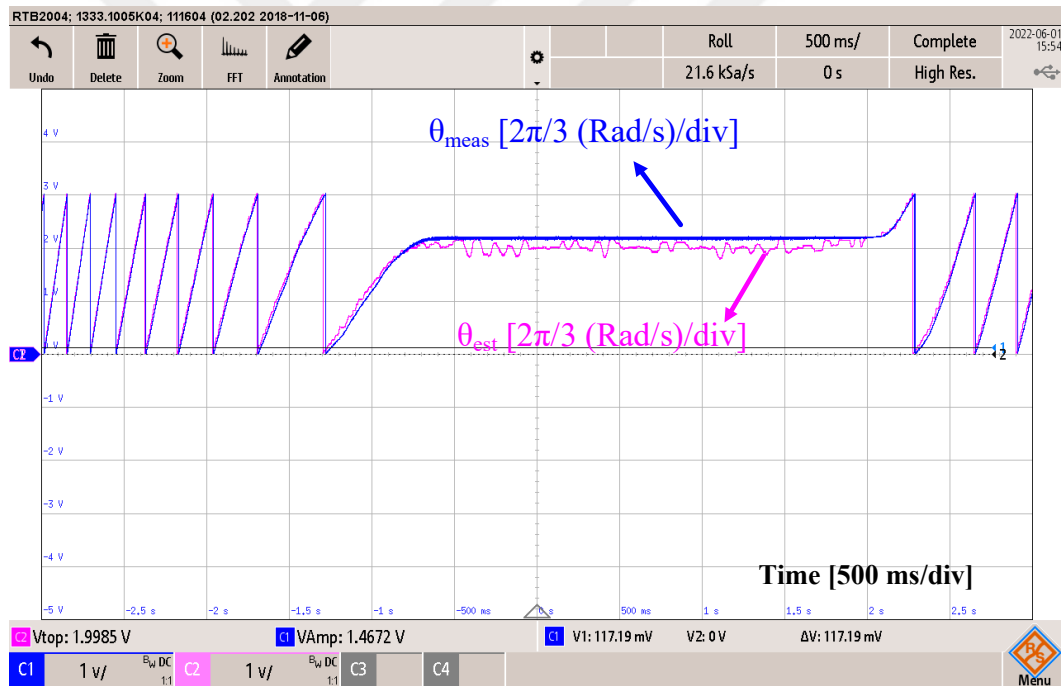
Scenario 4 (Figure 6.16): Performance Comparison of the proposed voltage injection-based SMO&PLL and the conventional SMO&PLL for very low and zero speed reversals. This scenario is designed to evaluate and compare the accuracy of rotor position estimation achieved using the proposed High-Frequency (HF) voltage injection-based Sliding Mode Observer (SMO) combined with a Phase-Locked Loop (PLL) against the conventional SMO&PLL approach. The focus is on performance under challenging conditions, specifically at very low and zero speeds, where accurate rotor position estimation is critical but inherently more difficult due to reduced signal strength and increased susceptibility to noise and disturbances. The evaluation is conducted in both clockwise (CW) and counterclockwise (CCW) rotational directions to ensure a comprehensive analysis of performance across all operating scenarios. By highlighting the advantages of the proposed method - such as improved robustness, enhanced noise immunity, and greater precision in low-speed operation - the results aim to clearly demonstrate its superiority over the conventional approach, offering a reliable solution for applications demanding precise rotor position control under demanding conditions. Thus,

- Initially, the Interior Permanent Magnet Synchronous Motor (IPMSM) operates using the actual rotor position within a vector control (VC) system to ensure accurate and stable performance. In this setup, the conventional Sliding Mode Observer (SMO) combined with a Phase-Locked Loop (PLL) functions by processing the measured stator voltages and currents from the VC system. However, the conventional SMO&PLL operates in an open-loop configuration where its estimated rotor position is not fed back into the VC system. Instead, it is solely monitored for comparison purposes. This configuration is necessary because the conventional SMO&PLL method struggles to perform effectively at very low and zero speeds. These limitations arise due to insufficient signal strength and the challenges of accurately estimating the rotor position under such conditions. As a result, the conventional method cannot provide the precision required for closed-loop operation at these critical speeds. By monitoring its performance in this open-loop setup, the study aims to highlight the differences between the conventional approach and advanced methods that are capable of reliable operation even at near-zero or zero speeds. This analysis underscores the importance of innovative rotor position estimation techniques for improving the operational range and robustness of IPMSM systems.

- later, the IPMSM is operated by using the proposed HF voltage injection-based SMO&PLL in the closed loop configuration, as shown in Figure 5.2.



(a)



(b)

Figure 6.16 The real-time comparison results of the rotor position estimations during low and zero speed region for Scenario 3. (a) the conventional SMO&PLL (without voltage signal injection) working in the open loop, (b) the proposed HF voltage injection based SMO&PLL operated in the closed loop VC system.

Finally, the real-time experimental results related to the estimated and the actual rotor position of both conventional and proposed methods in this paper are obtained for the cases the rotor speed approaches zero in the CW direction, stays at zero speed for a

duration of time, and accelerates in the CCW direction, as given in Figure 6.16. Examining the obtained results in Figure 6.16. (a) and 6.16(b), the following observations can be made:

- The rotor position estimation error of the conventional SMO&PLL increases and results in the loss of the phase alignment as the rotor velocity approaches zero; Thus, it cannot be utilized as feedback in the closed loop VC system, which is expected and common result as reported by the literature.

Table 6.2 Error Comparisons for position estimation performance

Error Calculations Estimation Methods	MAE	IE	IAE	ISE	ITAE
Conventional SMO	0.9014	0.0907	1.0347	3.0638	-1.1830
Proposed SMO	0.1579	0.0599	0.1564	0.3243	-0.1469

The results presented in Table 6.2 highlight a comprehensive comparison of position estimation performance between the proposed SMO and the conventional SMO. The proposed SMO consistently demonstrates superior performance across all evaluated error metrics, including the Mean Absolute Error (MAE), Integral Error (IE), Integral of Absolute Error (IAE), Integral of Squared Error (ISE), and Integral of Time-weighted Absolute Error (ITAE). These metrics collectively provide a thorough assessment of the accuracy, robustness, and responsiveness of the estimation methods.

The consistent reduction in all error metrics achieved by the proposed SMO indicates its ability to enhance estimation accuracy and improve the overall quality of the position estimation process. The lower error values suggest that the proposed method effectively minimizes deviations, whether measured in terms of absolute error, accumulated error, or error magnitude weighted by time. Furthermore, the reduced squared error highlights the proposed method's ability to handle larger errors more effectively, while the improvement in time-weighted absolute error demonstrates its responsiveness and ability to adapt to time-varying dynamics.

The results underscore the robustness and reliability of the proposed SMO in mitigating estimation errors and optimizing performance. This makes it a promising solution for applications requiring precise position estimation, particularly in systems sensitive to estimation inaccuracies. Overall, the findings validate the superiority of the

proposed approach over the conventional method, reflecting its potential to significantly improve position estimation accuracy and system efficiency.

Table 6.3 Error Comparisons for speed estimation performance

Error Calculations Estimation Methods	MAE	IE	IAE	ISE	ITAE
Conventional SMO	1.3024	0.6765	1.6168	17.2459	-54.2377
Proposed SMO	0.4657	0.1698	1.0255	4.7834	-7.9390

- It is very clear that the proposed HF voltage injection-based SMO&PLL is able to precisely estimate the actual rotor position at very low and zero speeds due to the injected HF voltage on d - axis amplifying the attenuated EEMF signal, which is the first according to the literature utilizing the fundamental model of IPMSMs.

As conclusion, the obtained comparison results confirm the superior performance of the proposed HF voltage injection-based SMO&PLL over the conventional SMO&PLL.

6.5 Conclusion

This chapter presented the simulation and experimental validation of the proposed sensorless control strategy for IPMSMs. The results illustrated the effectiveness and robustness of the HF voltage injection-based SMO integrated with a PLL in precisely estimating rotor position and velocity across various operating conditions.

The simulation phase provided valuable insights into the system's performance under controlled conditions, highlighting its ability to handle noise, parameter variations, and low-speed challenges. These findings were further substantiated through experimental validation using an IPMSM test bench, where the proposed method exhibited high accuracy, stability, and computational efficiency. The alignment of simulation and experimental results reinforced the practical feasibility of the approach, demonstrating its superiority over conventional sensorless control techniques.

Overall, this chapter confirmed that the proposed sensorless control strategy is a reliable, robust, and efficient solution for real-world applications. Its adaptability to diverse operating conditions and minimal hardware requirements makes it a promising candidate for deployment in industrial and automotive systems, setting a strong foundation for future advancements in IPMSM control technology.

Chapter 7

Conclusions and Future Prospects

7.1 Conclusions

In this paper, we introduce for the first time a SMO based on HF voltage signal injection combined with a PLL estimator to address the challenges of rotor position estimation in IPMSMs. This novel approach is aimed at resolving issues typically associated with observer-based position estimation, especially at extremely low and zero speeds, where attenuated and deteriorated BEMF signals are highly susceptible to noise. By leveraging the HF voltage injection method, our design avoids reliance on additional signal processing algorithms, simplifying the design process and reducing the computational load in position sensorless VC systems.

The real-time experimental validation, conducted on an 8-pole, 0.4 kW IPMSM under no-load and step-like load torque variations, demonstrates the system's effectiveness across a wide speed range. The experimental results further underscore the practical applicability and robustness of the proposed system, highlighting its ability to maintain accurate rotor position estimation under dynamic operating conditions and varying load demands. Specifically, the results confirm that:

The closed-loop VC system, utilizing the proposed voltage signal injection-based SMO and PLL, operates reliably at high, very low, and zero speeds, overcoming limitations inherent in conventional SMO- and PLL-based estimators. By achieving a consistent rotor position estimate, even during zero-speed conditions where back EMF is minimal, the system addresses one of the most challenging aspects of IPMSM control, where traditional observer methods often fail or exhibit instability.

The proposed SMO and PLL design exhibits improved robustness, accuracy, and stability compared to conventional methods, offering substantial advancements in sensorless control techniques within the literature. This improved robustness is evident in the observer's resilience to parameter variations, noise, and load disturbances, which can

otherwise impact estimation accuracy. Overall, the proposed method represents a significant contribution to the field, setting a new standard for efficient and reliable sensorless control in IPMSM applications, particularly suited for applications requiring high-precision performance across a broad operating range.

7.2 Societal Impact and Contribution to Global Sustainability

This work makes a substantial contribution to the advancement of more efficient, reliable, and cost-effective electric motor control systems, which are critical components across diverse applications, including industrial machinery, electric vehicles (EVs), and household appliances. By eliminating the need for physical position sensors, the proposed sensorless control method has the potential to reduce manufacturing costs, minimize maintenance demands, and improve the operational longevity of IPMSM systems. Physical position sensors not only add complexity to motor control systems but also introduce potential failure points. By replacing these sensors with a robust sensorless control method, our approach enables a more streamlined design, reducing both initial costs and the likelihood of performance degradation over time.

This research aligns with the ongoing global shift toward improved energy efficiency and sustainable technology. The high accuracy and reliability of the sensorless IPMSM drive system reduce energy losses commonly associated with less precise motor control, which, in turn, can decrease overall energy consumption. Given that electric motors account for a significant portion of worldwide electricity usage, enhancing the energy efficiency of motor-driven systems is crucial. The reduction in energy demand resulting from the proposed method can contribute to lower greenhouse gas emissions, thus supporting sustainability goals and the push for greener industrial processes.

In addition to its industrial applications, this work holds promise for advancing EV technology, where IPMSMs are widely employed due to their compact size, high efficiency, and superior torque characteristics. Reliable sensorless control at low and zero speeds achieved through the proposed sliding mode observer and phase-locked loop combination could prove transformative for EV drivetrains. By enabling smooth, precise control under diverse driving conditions, this method can enhance both the performance and affordability of EVs. The reduced manufacturing and maintenance costs further contribute to lowering the overall price of EVs, making them more accessible to a broader

range of consumers. Consequently, this innovation supports the global shift towards eco-friendly transportation solutions by facilitating the broader adoption of EVs, which play a key role in reducing urban pollution and greenhouse gas emissions. The proposed control method, therefore, not only advances motor control technology but also has a meaningful impact on environmental and sustainability efforts across industries.

7.3 Future Prospects

Building on the current work, our future studies will focus on improving the dynamic performance of the position sensorless VC system for IPMSMs. Key areas of interest include refining the observer's response to rapid speed and load changes, enhancing noise rejection at low and zero speeds, and optimizing the design for various IPMSM configurations and power ratings. Additionally, we plan to investigate adaptive techniques that could allow the observer to adjust to changes in motor parameters over time, further improving robustness and accuracy in long-term operation. We are also interested in exploring real-time implementation on high-performance embedded platforms, facilitating practical applications in industrial and automotive contexts.

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C1) E. Ateş, B. Tekgün, and G. Ablay. (2021) “Sensorless Position and Speed Control of IPMSM with Sliding Mode Observer and Voltage Signal Injection,” presented at the 3rd Global Power, Energy and Communication Conference (GPECOM), Antalya, Türkiye.

C2) E. Ateş, B. Tekgün, and G. Ablay. (2020). Sliding Mode Control of a Switched Reluctance Motor Drive with Four-Switch Bi-Directional DC-DC Converter for Torque Ripple Minimization. Presented at the 2020 International Conference on Smart Energy Systems and Technologies (SEST), İstanbul, Türkiye.