

Sensorless control for electric power steering in automotive applications: a systems engineering perspective

by

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Abstract

With the rapid expansion of vehicle electrification and automated driving, Electric Power Steering (EPS) has become a foundational safety-critical subsystem in modern vehicles. Unlike hydraulic systems, EPS relies on electronic control of a Permanent Magnet Synchronous Motor (PMSM) to generate assist torque, enabling improved energy efficiency, packaging flexibility, and software-defined steering characteristics. As EPS systems increasingly operate under ISO 26262 (functional safety) and ISO/SAE 21434 (cybersecurity) constraints—particularly for ASIL-D applications—the architectural robustness of motor control becomes a system-level concern rather than a purely algorithmic one. Sensorless motor control has emerged as a promising approach to reduce dependency on hardware position sensors such as resolvers or Hall devices, thereby lowering cost, complexity, and failure exposure.

This thesis investigates the structural feasibility of a hybrid sensorless control architecture for PMSM-based EPS, combining Sliding Mode Observer (SMO)-based back-EMF estimation in the medium-to-high speed region with High-Frequency Injection (HFI)-based saliency tracking for low-speed observability. Quantitative validation is performed in a MATLAB/Simulink environment using an IPMSM model with saliency. The results demonstrate stable and robust SMO-based estimation in the medium-to-high speed region, with consistent speed tracking and torque ripple characteristics under load variation. While full closed-loop low-speed hybrid validation remains future work, the HFI signal-processing path and architectural integration are implemented and structurally verified.

Beyond algorithmic comparison, this study conducts an explicit EPS-oriented cross-evaluation of four major sensorless control strategies (EKF, SMO, MRAS, PLL) and establishes that the SMO-HFI hybrid offers the most favorable robustness-to-complexity tradeoff for sedan-class production EPS systems. The work adopts a systems engineering perspective, cascading vehicle-level safety, hardware, software, and cost constraints down to observer design. This traceability from system objectives to signal-level control structure constitutes the primary contribution of the thesis and defines a realistic pathway toward future steer-by-wire and higher-automation steering architectures. Most conventional studies on sensorless control focus narrowly on specific algorithmic techniques, such as SMO-based rotor position estimation or high-frequency signal injection methods, without explicitly addressing their integration into full vehicle steering architectures. In contrast, the present work adopts a systems engineering perspective grounded not only in academic research but also in practical development experience. Rather than optimizing an observer in isolation, this thesis incorporates higher-level EPS system requirements—including functional safety, production feasibility, computational constraints, and vehicle software-hardware integration—into the architectural decision process. These top-level requirements are systematically cascaded down to the estimator design through a structured top-down methodology, forming a key distinguishing feature of this study. By embedding cross-domain constraints into

the SMO–HFI hybrid architecture, the proposed framework achieves more than signal-level estimation accuracy. It maintains coherence with implementation realities encountered in production-grade automotive systems, such as limited control-cycle budgets, inverter nonidealities, diagnostic requirements, and sensorless operational boundaries. This explicit traceability from system-level objectives to observer-level structure differentiates the work from conventional control-centric approaches and enables a balanced design that integrates academic rigor with industrial feasibility.

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Chapter 1. The Structure of the Thesis

1.1 Background and objectives of research

In recent years, the advancement of vehicle electrification and automated driving technologies has significantly increased the importance of Electric Power Steering (EPS) systems in modern automobiles. Unlike conventional hydraulic power steering systems, EPS utilizes an electronic control unit (ECU) to process inputs from sensors such as steering torque sensors and angle sensors and provides assistive torque via a permanent magnet synchronous motor (PMSM) or a brushless DC motor. EPS contributes to improved fuel economy, packaging efficiency, and customizable assist characteristics through software control, and its installation rate in new passenger vehicles has reached virtually 100%.

However, as steering systems are safety-critical components, they are subject to strict requirements for functional safety in accordance with ISO 26262, as well as cybersecurity standards such as ISO/SAE 21434. Moreover, in anticipation of ASIL-D/SAE Level 3 and above autonomous driving, EPS systems must support fail-operational capability, meaning they must retain a minimum level of steering assistance even in the event of failure. Against this backdrop, sensorless control techniques for motor drive systems have gained considerable attention.

Currently, motor position sensors such as Hall-effect sensors and resolvers are commonly used in EPS to detect rotor position. However, these sensors impose constraints in terms of hardware cost, installation space, environmental durability, and failure risk. In high-end systems such as redundant EPS or Steer-by-Wire (SBW), complete sensor duplication is required, which significantly impacts cost and complexity. In addition, the growing demand for compact and cost-sensitive electric vehicles has motivated the removal or simplification of not only *motor position sensors*, but also *steering angle and torque sensors*. In this context, sensorless control technologies offer the potential to simultaneously reduce system cost and enhance reliability.

This study proposes a novel sensorless control strategy for PMSM-based EPS systems, targeting only the elimination of *motor position sensors* while maintaining high control performance. The proposed method combines back-electromotive force (back-EMF) estimation for mid-to-high-speed operation with high-frequency signal injection techniques for accurate rotor position detection in the low-speed and standstill regions. In particular, a hybrid control algorithm incorporating a sliding mode observer (SMO) is employed to extract rotor saliency information from the current response to high-frequency d- or q-axis signal injection, enabling robust position estimation across a wide speed range. MATLAB/Simulink simulations using a salient IPMSM model confirm that SMO-based estimation is stable and robust at medium-to-high speeds, even under

varying loads. While full hybrid closed-loop testing is planned for the future, the HFI signal logic and system architecture have been successfully implemented and verified.

It should be added that sensorless control strategies for steering angle and torque are important research topics for future electric power steering (EPS) systems. However, these techniques have not yet been implemented in practice or widely adopted due to concerns about control stability, functional safety, and computational burden. Additionally, steering angle and torque sensors are tightly integrated with the advanced driver assistance systems (ADAS); for example, lane-keeping assist and steering-based autonomous driving require high-precision steering angle and torque feedback. Removing these sensors through sensorless methods could compromise control quality, thus requiring further technical validation and reliability engineering. In light of these challenges, this thesis focuses on motor-side sensorless control, which is currently feasible to implement and does not hinder ADAS functionality. Looking ahead, achieving sensorless steering angle and torque estimation that remains compatible with ADAS will hinge on advances in high-precision estimation algorithms, robust safety diagnostics, and integrated ECU architectures.

In conclusion, the sensorless EPS control system - based on the integration of a sliding mode observer (SMO) based back-EMF estimation and high-frequency signal injection – is proposed in this thesis and offers a promising solution for next-generation EPS systems that require high reliability, cost efficiency, and safety. This work contributes to the field by providing a comprehensive study of sensorless control theory, algorithm design, and performance validation, thereby highlighting a viable direction for the future of steer-by-wire and autonomous vehicle steering technologies.

1.2 The Structure of the Thesis

The present thesis is organized into 8 main chapters, each of which is designed to progressively guide the reader from the foundational context of Electric Power Steering (EPS) systems to the advanced implementation and evaluation of a hybrid sensorless control architecture. The organization follows a top-down logic consistent with systems engineering principles, ensuring that high-level system requirements are systematically translated into low-level software/algorithmic and hardware design considerations. This structure also facilitates traceability between functional safety compliance, control performance objectives, and practical feasibility for mass production in the automotive domain.

Chapter 1 The Structure of the Thesis

The first chapter sets forth the motivation, objectives, and scope of the research. It outlines the industrial and technological context in which EPS has emerged as a key enabling technology for vehicle electrification and automated driving. The introduction situates the study within the broader landscape of functional safety (ISO

26262) and cybersecurity (ISO/SAE 21434), highlighting the inherent challenges posed by traditional sensor-based motor control systems in terms of cost, packaging, redundancy, and reliability. It also presents the core research hypothesis: that a hybrid sensorless control method—combining back-electromotive force (back-EMF) estimation with high-frequency injection (HFI) using a Sliding Mode Observer (SMO)—can achieve high position estimation accuracy, low torque ripple, and robust low-speed performance while satisfying production-level constraints. The chapter concludes by stating the key contributions of the work and their intended impact on both academic research and industrial EPS development.

Chapter 2 Overview of Electric Power Steering (EPS)

The second chapter provides a comprehensive overview of EPS technology from a systems engineering perspective. It begins with the historical evolution from manual and hydraulic steering to fully electronic power assist, describing the mechanical layouts (Column-Assist, Pinion-Assist, and Rack-Assist) and their respective advantages and trade-offs. It then examines the electrical hardware architecture in detail, including ECU design, microcontroller requirements, power management, gate driver ICs, and power semiconductor selection. Special emphasis is placed on power electronics control methods, comparing Sinusoidal PWM (SPWM) and Space Vector PWM (SVPWM). This chapter also discusses functional safety architectures, redundancy strategies, and electromagnetic compatibility measures, linking them to market trends, and emerging Steer-by-Wire (SbW) concepts. The purpose of this chapter is to equip the reader with a holistic understanding of EPS subsystems and their interactions, thereby establishing the design constraints that sensorless control must satisfy.

Chapter 3 Literature Review: Field-Oriented Control for PMSM Drives

Chapter 3 serves as the theoretical foundation for the motor control strategy employed in this thesis. It presents the mathematical modeling of Permanent Magnet Synchronous Machines (PMSMs) in both stationary and synchronously rotating reference frames, along with Clarke and Park transformations. The derivations of torque production equations, saliency effects, and d-q axis dynamic models are provided to support the later integration of sensorless estimation methods. The chapter also explains the structure of Field-Oriented Control (FOC), including the inner torque/current loop, outer speed loop, and PWM generation. This theoretical framework ensures that the subsequent discussion of sensorless control methods is firmly grounded in established control theory.

Chapter 4 Literature Review: Various Sensorless Control Strategies for PMSM

In Chapter 4, the focus shifts to existing rotor position estimation methods. Four principal strategies—Extended Kalman Filter (EKF), Sliding Mode Observer (SMO), Model Reference Adaptive System (MRAS), and Phase-Locked Loop (PLL)—are described in terms of estimation accuracy, computational requirements, sensitivity to parameter variation, and robustness under low-speed or transient conditions. Hybrid methods that combine high-frequency signal injection with back-EMF estimation are also examined, as they address the known limitations of purely back-EMF-based techniques at low speeds. In the following chapter5, the advantages and disadvantages of each method will be compared in detail, leading to a conclusion on the most suitable approach for sensorless EPS systems.

Chapter 5 The System-Level Superiority of the SMO–HFI Hybrid Sensorless Architecture for EPS

Chapter 5 consolidates the thesis’s final decision logic by moving one step beyond the literature review of Chapter 4. While Chapter 4 introduces four representative sensorless estimation families—EKF, SMO, MRAS, and PLL—primarily from a control-theory standpoint, Chapter 5 reframes them under EPS-specific system constraints and performs a cross-domain, application-driven comparison. In particular, the evaluation emphasizes requirements that are distinctive for steering systems, including standstill/parking operation, deterministic real-time execution on automotive ECUs, sensitivity of steering feel to torque ripple and phase delay, and the need for safety-oriented plausibility monitoring.

Based on this EPS-oriented comparison, Chapter 5 establishes that no single back-EMF-based observer can satisfy the full operating envelope of an EPS actuator, because observability degrades at low speed and standstill. Consequently, the chapter identifies a hybrid architecture as structurally necessary: saliency-based high-frequency injection (HFI) for low/zero-speed observability, combined with a robust and computationally efficient back-EMF observer for the medium-to-high speed region. Among back-EMF-based candidates, the analysis concludes that the sliding mode observer (SMO) offers the most favorable robustness-to-complexity tradeoff for sedan-class production EPS, while remaining compatible with embedded timing constraints and practical diagnostic integration.

Importantly, Chapter 5 frames this selection not merely as an algorithmic preference but as a system-level architectural choice that remains coherent across mechanical NVH constraints, inverter signal integrity realities, AUTOSAR-based software partitioning, and ISO 26262 safety reasoning. The chapter therefore positions the SMO–HFI hybrid as the most defensible baseline strategy for sensorless EPS, and it closes with a roadmap focusing on full low-speed closed-loop validation, transition-region optimization, and extension toward higher-redundancy steering architectures such as steer-by-wire.

Chapter 6 Systems Engineering Impact Analysis of Transitioning from Sensor-Based to Sensorless EPS

Chapter 6 analyzes the architectural consequences of removing rotor position sensors using a structured systems-engineering lens. Instead of treating sensor elimination as a control-layer modification, the chapter demonstrates that the sensor virtualization method that was selected in the chapter 5, affects five interdependent subsystems: mechanics and dynamics, electrical hardware, control architecture, embedded software structure, and functional safety.

The analysis reveals that rotor angle estimation becomes a software-derived safety-relevant variable whose quality depends on inverter signal integrity, timing determinism, and estimator robustness. Mechanical NVH behavior becomes more sensitive to estimator-induced excitation, inverter nonidealities become part of the sensing chain, and AUTOSAR scheduling must explicitly account for estimator timing constraints. Furthermore, safety arguments shift from physical redundancy toward analytical redundancy and estimator plausibility monitoring.

Chapter 6 therefore bridges vehicle-level requirements and low-level observer design, defining the architectural envelope within which sensorless EPS remains technically coherent and production-feasible.

Chapter 7 Simulation-Based Validation of SMO Sensorless Control

Chapter 7 presents the modeling and validation methodology adopted in the selected control strategy in the chapter 5. A MATLAB/Simulink-based framework, derived from a validated reference FOC model, was adapted to represent an IPMSM-based EPS actuator with saliency characteristics. The model incorporates SVPWM-based modulation, a sliding mode observer for medium-to-high speed estimation, and a structured HFI signal-processing path prepared for future low-speed integration.

While the initial research vision included full SMO–HFI closed-loop validation across the entire speed range, the present study focuses quantitatively on SMO-based estimation performance in the medium-to-high speed region. Performance metrics include rotor position estimation stability, speed tracking accuracy under load variation, and q-axis current ripple as a torque smoothness proxy.

The HFI path was implemented structurally but not fully validated in closed-loop low-speed operation. This scoping ensures that all reported results are reproducible and aligned with the validated simulation environment.

Modeling and Test Results

Simulation results demonstrate stable SMO-based sensorless control in the medium-to-high speed region. Speed tracking remains consistent under nominal and increased load conditions, and q-axis current ripple remains bounded even under $3\times$ torque disturbance. No estimator-induced instability is observed within the defined evaluation window. The thesis does not claim complete validation of low-speed hybrid operation. Instead, it establishes that the proposed architecture structurally supports such extension without requiring fundamental redesign.

The results confirm that SMO-based sensorless control satisfies actuator-level EPS requirements in the medium-to-high speed region while maintaining computational efficiency compatible with automotive ECUs. The architecture is consistent with AUTOSAR integration and functional safety reasoning.

However, limitations remain. Closed-loop HFI validation is incomplete, inverter nonlinearity sensitivity requires further compensation strategies, and calibration robustness across manufacturing tolerances must be addressed for production deployment. Future work includes low-speed hybrid stabilization, transition-region optimization, adaptive gain scheduling, and integration with vehicle-level steering dynamics models.

Chapter 8 Conclusion

This thesis concludes that the SMO–HFI hybrid sensorless architecture constitutes a technically coherent and systemically justified solution for sedan-class EPS applications when evaluated from a systems engineering perspective. The contribution lies not in isolated algorithm derivation, but in demonstrating how control theory, inverter constraints, AUTOSAR-based software architecture, and ISO 26262 safety reasoning can be integrated into a unified architectural argument.

The validated medium-to-high speed SMO performance, combined with the structured impact analysis across mechanical, electrical, software, and safety domains, supports the feasibility of sensorless EPS as a scalable and cost-effective production strategy. The roadmap toward full-speed hybrid validation and future steer-by-wire architectures is clearly defined, ensuring continuity of development while maintaining functional safety and industrial scalability.

Chapter 2. Overview of EPS (Electric Power Steering)

2.1 History of EPS

2.1.1 What is electric power steering and its advantage over conventional hydraulic systems.

EPS “Electric power steering” (Figure 2.1.1) uses an electric motor to assist steering force, and speed-responsive control changes the amount of assistance according to the vehicle speed, allowing for light steering at low speeds and stable steering at high speeds. Compared to engine-driven hydraulic systems, there is no need to operate the pump all the time, reducing engine load and contributing to improved fuel efficiency. In addition, the high system controllability provided by electronic control allows for precise adjustment of steering characteristics and advanced integration with other vehicle control systems (Figure 2.1.2). [2-15]

Characteristics

Hitachi Automotive Systems Americas, Inc. Belt Drive Electric Power Steering System with enhanced steering feel and ISO 26262 ASIL-D compliant design.

Features & Benefits

- Advanced safety with Limp Home and ISO 26262 ASIL-D compliant design
- High reliability due to elimination of mechanical relays & ribbon cables
- Enhanced steering feel is achieved by high accuracy rack, ball screw, pinion gear and ball screw grade control, all made in-house by Hitachi Automotive Systems Americas, Inc. [2-15]
- Full development capability from control logic to system development.
- **Torque Sensor:** Non-contact Hall effect
Redundant
Digital communication
- **Absolute Steering Angle Sensor:** Twin Gear with non-contact GMR plus Motor Position Sensor
Redundant and Digital communication
- **Limp Home function** is realized by the addition of redundant sensors

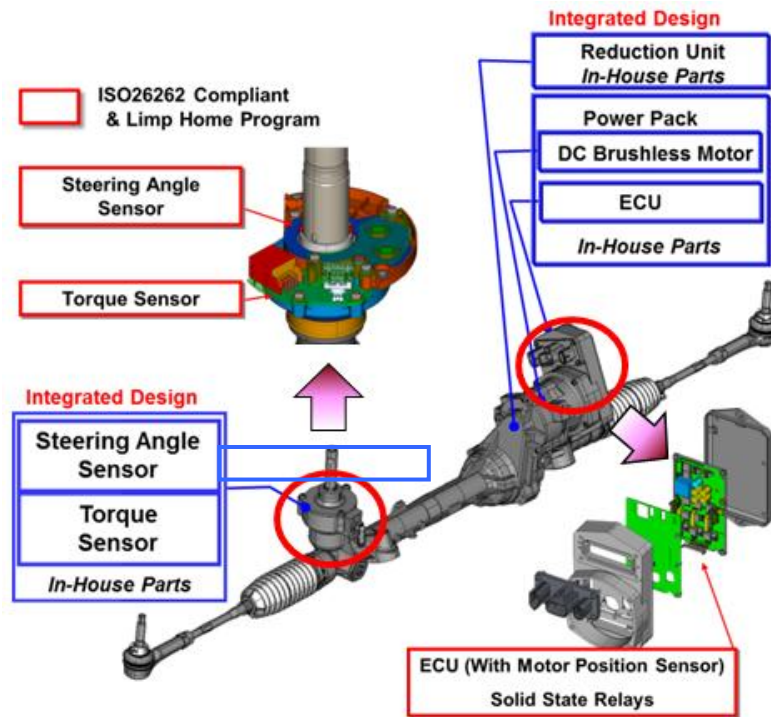


Figure 2.1.1 Hitachi Astemo Americas, Inc., “Steering System: Electric Power Steering (EPS) [2-15]

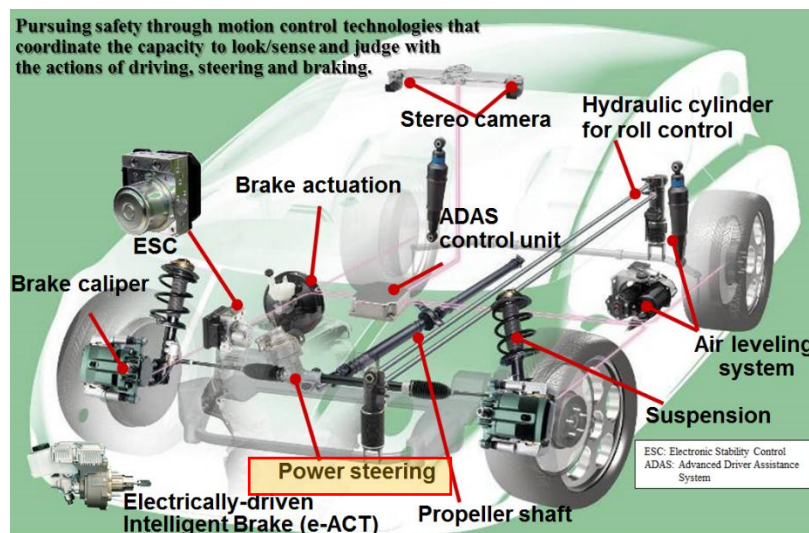


Figure 2.1.2 EPS in vehicle; Hitachi Astemo Americas, Inc., “Steering System: Electric Power Steering (EPS) [2-15]

2.1.2 Technological evolution and emergence of sensorless system

Backgrounds and demands for EPS

With the increasing size and weight of vehicles, manual steering became difficult, leading to the emergence of hydraulic power steering (HPS) systems in the mid-20th century. While HPS significantly reduced steering effort at low speeds, its engine-driven pump operated continuously, resulting in poor energy efficiency and increased fuel consumption. Additionally, hydraulic systems faced maintenance issues such as oil leakage.

In response, Electric Power Steering (EPS) systems were introduced in the late 1980s. EPS only provides assist torque when necessary and can shut down the motor during straight-line driving, offering considerable energy savings. Its power consumption is about 1/10 that of HPS, and the simplified structure allows for weight reduction, space efficiency, and reduced environmental impact.

The emergence and adoption of electronic control further enabled variable assist based on vehicle speed and integration with functions such as automated parking. EPS can also operate in hybrid and electric vehicles where the engine may be off, making it essential for modern vehicle platforms.

Global Adoption and Spread

The world's first mass-produced EPS-equipped vehicle was the 1988 Suzuki Cervo. Through the 1990s, EPS gained traction among automakers, although early applications were limited to compact cars due to constraints in motor output and semiconductor performance. By the 2000s, stricter environmental regulations and technical advancements accelerated global adoption. With increasing motor output, EPS became applicable to SUVs, luxury vehicles, and heavier segments.

Technological Advancements: Motors, Control, Sensorless Techniques

Early EPS systems used brushed DC motors with simple electronic control, but due to issues like wear and slower response, the industry shifted to brushless motors such as PMSMs. These feature stator windings and permanent magnet rotors, enabling high torque, responsiveness, and durability while minimizing rotor inertia.

EPS control is based on *torque assist computation*, where sensors detect steering torque and vehicle speed, allowing the ECU to calculate the required assist torque and issue commands to the inverter. This real-time assist modulation ensures that driver inputs are supported precisely, enhancing stability and steering feel and is designed to *reduce manual steering effort* during low-speed maneuvers.

In recent years, sensorless control techniques have gained attention - especially methods estimating rotor position using back-electromotive force (Back-EMF). These enable high-precision control without Hall sensors, reducing cost and improving reliability. As sensorless EPS systems become more practical, it is

increasingly necessary to re-examine system-level functional safety and redundancy, especially in the absence of traditional position sensors.

Current EPS System Configurations

EPS systems can be broadly classified into three types based on motor and gear positioning:

- Column-Assist EPS (C-EPS): Simple design suitable for compact vehicles; low-cost with in-cabin motor placement offering good thermal and waterproofing characteristics.
- Pinion-Assist EPS (P-EPS): Motor applies torque at the rack's pinion gear. Offers excellent steering feel and is widely used in mass-produced vehicles. Dual-pinion configurations improve crash safety and decouple assist from steering mechanics.
- Rack-Assist EPS (R-EPS): Motor directly actuates the rack. Though more expensive, it provides superior torque, responsiveness, and precision, making it ideal for SUVs, sports cars, and premium vehicles.

Future Outlook: Autonomous Driving, Steer-by-Wire, and AI

EPS plays a vital role as an enabler of steer-by-wire (SbW), a key technology for autonomous vehicles. SbW eliminates mechanical links between the steering wheel and road wheels, relying entirely on sensors and actuators. This enables variable steering ratios, flexible interior layouts, and improved ride comfort by electronically filtering out vibrations.

However, the absence of mechanical fallback necessitates extremely high reliability of electronics, requiring redundant architecture such as dual power supplies, communication paths, and motor systems. These requirements align with safety levels defined by ISO 26262 and are essential for safe deployment in automated driving scenarios.

AI-driven smart steering is also under development. It dynamically adjusts steering characteristics based on road conditions and driver preferences, and may eventually integrate with other vehicle dynamics systems (e.g., braking and suspension) to create a unified dynamic control platform.

Present Challenges and Functional Safety Considerations

While EPS has matured technologically, several challenges remain from the perspectives of control quality, safety, and cost:

Steering Feel and Responsiveness: Despite improvements, some drivers claim EPS lacks the road feedback of hydraulic systems. For sensorless EPS and SbW, reproducing realistic steering feel remains a technical challenge.

Functional Safety and Redundancy: Especially in SbW, there must be no single point of failure that can lead to steering loss. Dual electrical systems, control units, and power supply paths are essential. These redundant designs, however, increase cost and weight, requiring intelligent optimization.

Sensorless Control and Functional Safety Reevaluation: Although eliminating sensors improves reliability and reduces costs, it also requires new approaches to ensure safety. Redundant signal paths and fault detection, formerly reliant on sensors, must be replaced by alternative mechanisms.

High Output and Heavy Vehicle Applications: High-output EPS using 48V systems or electro-hydraulic hybrids are being developed, but fully electrifying heavy vehicles still demands improvements in motor durability, gearbox design, and system robustness.

Cost Constraints: High-performance magnets and power semiconductors dominate EPS system cost. Alternative materials and component standardization are essential to support future innovations such as SbW without compromising safety.

In summary, EPS has evolved as a highly efficient, responsive, and scalable steering technology and now serves as the technological bridge to autonomous driving and steer-by-wire architectures. However, the transition to sensorless control and redundant safety mechanisms introduces new engineering challenges. Going forward, a redefinition of EPS architecture that integrates functional safety, cost efficiency, and advanced control will be indispensable. These background and technological considerations form the foundation for further discussion in the following chapters of this thesis.

2.2 EPS systems

2.2.1 Mechanics and dynamics

Figure 2.2.1 illustrates a streamlined configuration of the EPS system, which comprises two terminal points of the torque bar—one linked to the steering wheel, and the other to the steering column. A torque sensor mounted along the torque bar identifies deformation, which acts as an input to the ECU controller. The ECU then governs the motor to deliver supplementary assist torque. This assist torque is subsequently enhanced by the reduction gearing, enabling it to counterbalance both internal friction and resistance torque from the road surface.

The mathematical formulation describing the torque at the upper end of the torque bar is given by [2-1], [2-2]:

$$T_{sw} - T_s = c_{sw}\omega + J_{sw}a_{sw}, \quad (2.1)$$

where T_{sw} denotes the torque applied at the steering wheel; T_s refers to the torque transmitted through the torque bar, which is detected by the torque sensor; c_{sw} stands for the damping coefficient of the steering wheel; ω is the angular velocity of the steering wheel; J_{sw} is the rotational inertia associated with the wheel; and a_{sw} represents the angular acceleration of the steering wheel.

In practice, the values of J_{sw} and c_{sw} are sufficiently small to be neglected in modeling. As a result, equation (2.1) may be approximated by the simplified form; [2-3]:

$$T_s \approx T_{sw} \quad (2.2)$$

The torque relationship at the lower terminal of the torque bar can be formulated as:

$$GT_e + T_s = J_p a_p + c_p \omega_p + T_r \quad (2.3)$$

where G represents the gear ratio of the reduction mechanism; T_e denotes the electromagnetic torque produced by the motor; J_p corresponds to the rotational inertia of the steering column; a_p stands for the angular acceleration of the column; c_p is the damping coefficient attributed to the steering column; ω_p signifies the angular velocity of the steering column; and T_r accounts for the torque generated by road resistance. Moreover, the expression for the motor's electromagnetic torque is given by [2-4], [2-5]:

$$T_e = K_m \cdot i, \quad (2.4)$$

where K_m refers to the torque constant of the motor, and i represents the motor current.

Motor torque T_e is directly related to steering assist torque and plays a central role in providing steering assistance and serves as the vital core of the EPS control system, as it eliminates the need for excessive manual steering effort. Precisely regulated based on both driving inputs and road conditions, T_e enhances driver comfort and facilitates ease of maneuverability through sophisticated motor assist control.

Since the motor is directly coupled with the reduction mechanism, the ω_m corresponds to the angular velocity of the motor, which leads to the following relationship for ω_p :

$$\omega_m = G \omega_p \quad (2.5)$$

Therefore, by substituting equations (2.2) through (2.5), the steering wheel torque can be formulated as:

$$T_{sw} = -GK_m i + T_r + J_p a_p + c_p \frac{\omega_m}{G} \quad (2.6)$$

It is evident that the torque experienced at the steering wheel is significantly influenced by both external road disturbances and the inherent damping characteristics of the system. The external disturbance input may be represented as a time-dependent function, $a(t)$ [2-2], [2-3]:

To achieve consistent and stable steering torque performance, the effects of these disturbance factors must be mitigated through precise regulation of the motor current.

The motor's behavior, under such control, can be represented using the following simplified dynamic model [2-4], [2-5]:

$$u = Ri + pL_m i + K_e \omega_m, \quad (2.7)$$

where u denotes the voltage applied to the motor, R represents the equivalent resistance of the motor's armature winding, p is the differential operator, L_m corresponds to the motor's inductance, and K_e stands for the back electromotive force (EMF) constant of the motor.

The reference principal parameters included in the above formulation are detailed in Table 2.2.1.

Parameters	Unit	Value
L_m	mH	2.5
R	Ω	0.5
K_e	V·s/rad	0.06
K_m	N·m/A	0.06
J_p	kg·m ²	0.05
c_p	N·m·s/rad	0.3
G	—	18

Table 2.2.1 *Selected Parameters* of the Steering Mechanism [2-1], [2-2]

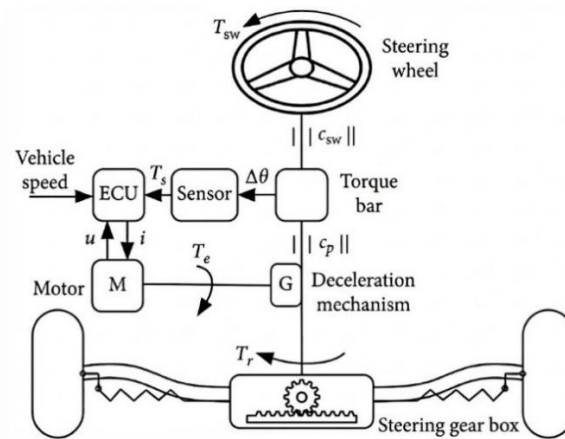


Figure 2.2.1 EPS system model (simplified) [2-1], [2-2]

2.2.2 Electrical Hardware System (ECU)

2.2.2.1 Key Circuit Components and Their Functions

Figure 2.2.2 show system block diagram for electric circuit board of EPS module (ECU). This section describes the major component and its function of the circuit.

Microcontroller Unit (MCU):

Serves as the brain of the EPS control board, executing control algorithms and managing communication with other ECUs. Automotive-grade 32-bit MCUs with high-performance CPU cores, ample memory, dedicated timers, and high-speed, high-precision ADCs are required to implement sophisticated control such as vehicle-speed-dependent steering angle modulation. Functional safety mechanisms such as lockstep dual-core architecture and memory ECC are integrated for ASIL compliance. [2-6]

Power Management IC (PMIC):

Converts 12V car battery input into regulated voltages (e.g., 5V, 3.3V) for the MCU, gate drivers, and sensors. Automotive PMICs include undervoltage/overvoltage detection, watchdog timers, and monitoring circuits necessary for ASIL-D compliant designs. Pairing with dedicated MCUs reduces part count and board size while ensuring high reliability. [2-6]

Gate Driver IC:

Interfaces MCU's control signals (PWM, etc.) to high-current motor-driving MOSFETs. Typically includes three half-bridge drivers for 3-phase PMSC/BLDC motors, bootstrap circuits for high-side drive, and integrated protection/diagnostics (overcurrent, thermal shutdown, undervoltage). Infineon's TLE9189, for example, features adaptive MOSFET control to reduce EMI and switching losses. [2-7]

Power Semiconductors (MOSFETs, etc.):

The 3-phase bridge for BLDC motor drive utilizes high-current, high-voltage-rated MOSFETs or IGBTs. Devices must endure transient torque demands and thermal shocks. Key selection criteria include low $R_{ds(on)}$, high avalanche rating, and high pulse current tolerance. AEC-Q101-compliant packages such as D2PAK or TO-220 are preferred for automotive reliability. [2-7]

Design Considerations: Safety, Redundancy, EMC

Compliance with Functional Safety (ISO 26262):

As steering control is classified under the highest functional safety level (ASIL-D), the electronic circuitry must be designed to maintain safety even in the event of a failure. Specifically, MCUs are equipped with error

detection mechanisms such as lockstep dual-core CPUs and ECC-protected memory, enabling high diagnostic coverage. ECC-protected Flash and RAM are typically employed in ASIL-D EPS ECUs to detect and correct single-bit memory errors and detect multi-bit corruption, thereby preventing silent data corruption in safety-critical control loops.

Critical sensor signals—such as those from torque sensors—are duplicated to allow for mutual comparison, enabling fault detection and isolation in case of abnormalities. Furthermore, mechanical backups (e.g., rack-and-pinion systems) are sometimes retained to ensure that steering force can still be transmitted by the driver in the event of total control loss. [2-6]

Redundant Monitoring and Fail-Safe Design:

To prevent loss of assist functionality due to a single point of failure, system architectures incorporating mutual monitoring between major components are adopted. For example, Texas Instruments (TI) proposes architecture in which the safety MCU and the power management IC monitor each other's operation via watchdog timers and voltage monitors. This configuration enables immediate transition to a fail-safe state - such as cutting motor drive output or issuing warnings—when an MCU fault or power failure is detected. Gate driver ICs also include redundant clock monitoring and built-in self-test (BIST) capabilities (ti.com), allowing for the detection of latent internal faults and a controlled transition to a safe state.

EMC/EMI Countermeasures:

To prevent electromagnetic noise from interfering with in-vehicle devices and wireless systems due to high-current switching, compliance with EMC standards such as CISPR 25 is required. In PCB layout, switching loops must be minimized, and decoupling capacitors and noise filters must be properly placed. Techniques such as adjusting switching speed and suppressing transient voltages with gate resistors or snubber circuits are effective. In practical circuit design, the insertion of gate resistors or gate-to-gate capacitors helps moderate the signal's rising edge, while RC snubbers and Schottky diodes may be used in parallel to suppress large reverse-voltage spikes. Furthermore, advanced approaches such as Infineon's adaptive gate control dynamically tune switching waveforms to reduce EMI while maintaining response performance (infineon.com). [2-7]

Development Kits and Technical Notes by Major Vendors

Infineon Technologies:

Offers comprehensive EPS solutions centered around AURIX™ MCUs and MOTIX™ 3-phase smart gate drivers. The MOTIX TLE9189 series is ASIL-D compliant and suitable for steer-by-wire systems. Infineon

provides application notes, design tools, and evaluation boards supporting MOSFET/gate driver co-evaluation.

Renesas Electronics:

Renesas supplies RH850/P1M-E MCU series (40nm, ASIL-D ready) and companion PMICs (e.g., RAA270005) tailored for EPS. These combinations simplify power design while ensuring safety. Renesas also provides torque sensors like the inductive IPS series and comprehensive evaluation kits.

Texas Instruments (TI):

TI's Hercules™ safety MCU (dual-core Cortex-R5) and DRV32xx gate drivers enable ASIL-compliant EPS designs. The “Electric Power Steering Design Guide with DRV3205-Q1” offers detailed failure modes, protections, and system diagrams. TI provides EVMs, whitepapers, and safety architecture examples highlighting redundancy and robust protection.

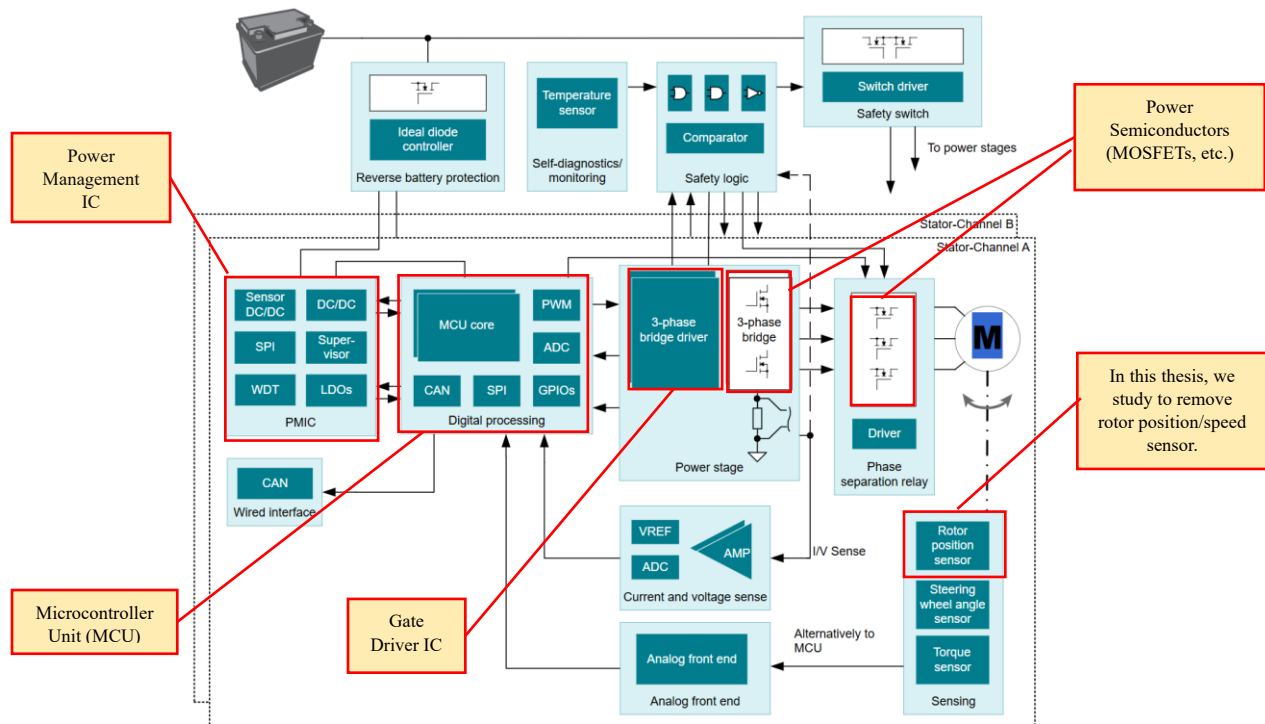


Figure 2.2.2. System block diagram for electric circuit board of EPS module (ECU) (Texas Instruments). [2-6], [2-7]

2.2.2.2 Power Electronics - SVPWM/SPWM for EPS

2.2.2.2.1 Introduction

Electric Power Steering (EPS) systems have become the dominant steering technology in modern passenger vehicles due to their high energy efficiency, compact packaging, and compatibility with advanced driver-assistance systems (ADAS). In contrast to conventional hydraulic power steering, EPS relies on an electric motor drive to generate assist torque, making the performance of the motor control system a critical factor in steering feel, noise, vibration, and harshness (NVH), and overall system efficiency.

Among the various elements of motor control, the pulse width modulation (PWM) strategy used to drive the inverter plays a fundamental role. In EPS applications, the PWM method directly influences torque ripple, acoustic noise, switching losses, and low-speed controllability. Traditional sinusoidal PWM (SPWM) has been widely used due to its conceptual simplicity; however, it suffers from limited DC bus voltage utilization and relatively high harmonic distortion. These limitations become particularly critical in EPS systems, where smooth torque generation at low speed and high efficiency under thermal constraints are required. [2-17], [2-13]

Space Vector Pulse Width Modulation (SVPWM) has therefore been widely adopted in automotive EPS motor drives. By exploiting the space vector representation of three-phase voltages, SVPWM achieves a higher linear modulation range and improved voltage utilization compared with SPWM. In addition, SVPWM enables a reduction in current harmonic distortion, leading to lower torque ripple and improved NVH characteristics—key performance metrics for steering systems that are directly perceived by the driver.

This chapter presents a detailed analysis of SVPWM as applied to automotive EPS systems. The focus is placed on the mathematical formulation of SVPWM, its equivalent implementation using common-mode voltage injection, and its suitability for EPS applications. Particular attention is given to the relationship between the classical space-vector-based formulation and practical implementations used in industry-standard motor control platforms. The discussion is limited to the linear modulation region, which is typically adopted in EPS systems to ensure predictable torque generation and superior steering feel. [2-12], [2-13]

2.2.2.2.2 Mathematical Background of Motor Voltage Representation

To establish the theoretical foundation for SVPWM, this section reviews the mathematical representation of three-phase motor voltages and the associated coordinate transformations commonly used in field-oriented control (FOC). These transformations provide the basis for expressing inverter output voltages as space vectors and enable an intuitive interpretation of SVPWM.

2.2.2.2.1 Three-Phase Voltage System

Consider a three-phase voltage system consisting of phase voltages v_a, v_b, v_c . For a balanced system without zero-sequence components, the phase voltages satisfy [2-8]:

$$v_a + v_b + v_c = 0 \quad (2.8)$$

In practical motor drives, these voltages are generated by a voltage-source inverter (VSI) supplied by a DC bus voltage V_{dc} . The objective of the PWM strategy is to synthesize the desired phase voltages by appropriately switching the inverter power devices.

2.2.2.2.2 Clarke Transformation [2-8]

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} \quad (2.9)$$

The inverse Clarke transformation, which reconstructs the three-phase voltages from the stationary reference frame, is given by [2-8]:

$$\begin{aligned} v_a &= v_\alpha \\ v_b &= -\frac{1}{2}v_\alpha + \frac{\sqrt{3}}{2}v_\beta \\ v_c &= -\frac{1}{2}v_\alpha - \frac{\sqrt{3}}{2}v_\beta \end{aligned} \quad (2.10)$$

The $\alpha\beta$ representation allows the three-phase voltage system to be treated as a single complex vector, which forms the basis of space vector analysis.

2.2.2.2.3 Park Transformation

In field-oriented control, the stationary reference frame is further transformed into a rotating reference frame aligned with the rotor magnetic field. This transformation, known as the Park transformation, is defined as [2-9]:

$$\begin{aligned} v_d &= v_\alpha \cos \theta + v_\beta \sin \theta \\ v_q &= -v_\alpha \sin \theta + v_\beta \cos \theta \end{aligned} \quad (2.11)$$

where θ denotes the electrical rotor position. The inverse Park transformation converts the voltage commands from the rotating frame back to the stationary frame [2-9]:

$$\begin{aligned} v_\alpha &= v_d \cos \theta - v_q \sin \theta \\ v_\beta &= v_d \sin \theta + v_q \cos \theta \end{aligned} \quad (2.12)$$

In EPS motor drives, the d -axis voltage is typically regulated to control the magnetic flux, while the q -axis voltage determines the electromagnetic torque. The resulting voltage vector (v_α, v_β) represents the desired inverter output voltage in the stationary frame.

2.2.2.2.4 Space Vector Representation

Using the Clarke transformation, the three-phase voltages can be expressed as a single complex space vector:

$$\vec{V}_{ref} = v_\alpha + jv_\beta \quad (2.13)$$

This space vector representation provides a geometric interpretation of the inverter output voltage as a vector rotating in the $\alpha\beta$ plane. SVPWM exploits this representation to optimally synthesize $\vec{V}_{ref}$ using the discrete switching states of the inverter.

The space vector formulation is particularly well suited for EPS applications, as it allows a unified treatment of voltage magnitude, phase, and inverter constraints. In the following section, the principle of SVPWM is derived based on this space vector representation, and the limitations imposed by the DC bus voltage are discussed. [2-5]

2.2.2.2.3 Principle of Space Vector Pulse Width Modulation

Space Vector Pulse Width Modulation (SVPWM) is a modulation technique that synthesizes the desired inverter output voltage by considering the inverter as a single entity rather than treating each phase independently. By exploiting the space vector representation introduced in the previous section, SVPWM enables optimal utilization of the DC bus voltage and improved waveform quality compared with conventional sinusoidal PWM.

2.2.2.2.3.1 Voltage Source Inverter and Space Vectors

A three-phase voltage source inverter (VSI) consists of three inverter legs, each of which can connect the corresponding motor phase either to the positive or the negative DC bus. As a result, the inverter has eight possible switching states. Six of these states generate non-zero output voltages, referred to as *active voltage vectors*, while the remaining two states produce zero output voltage, referred to as *zero voltage vectors*.

Using the space vector representation, each active switching state corresponds to a fixed voltage vector in the $\alpha\beta$ plane. These six active vectors are evenly spaced by 60 electrical degrees and form the vertices of a regular

hexagon. The two zero vectors are located at the origin. The set of inverter voltage vectors can therefore be expressed as [2-12]:

$$\vec{V}_k = \frac{2}{3} V_{dc} e^{j(k-1)\frac{\pi}{3}}, \quad k = 1, 2, \dots, 6 \quad (2.14)$$

where V_{dc} denotes the DC bus voltage.

This hexagonal structure defines the fundamental voltage limitation of a three-phase VSI. Any realizable output voltage vector must lie within the hexagon formed by the six active vectors. [2.12]

2.2.2.2.3.2 Linear Modulation Region

The objective of SVPWM is to synthesize a reference voltage vector $\vec{V}_{ref}$ by appropriately combining the inverter switching states over one PWM period T_s . In the linear modulation region, $\vec{V}_{ref}$ is located strictly inside the hexagon defined by the active voltage vectors.

The maximum magnitude of the reference voltage vector that can be synthesized without entering overmodulation is given by:

$$|\vec{V}_{ref}| \leq \frac{\sqrt{3}}{2} V_{dc} \quad (2.15)$$

This condition defines the linear operating range of SVPWM. Compared with sinusoidal PWM, which is limited to a maximum reference magnitude of $V_{dc}/2$, SVPWM provides approximately 15% higher voltage utilization. This increased utilization is particularly beneficial in EPS applications, where the available DC bus voltage may be constrained and high torque output is required at low motor speeds.

In automotive EPS systems, operation is typically restricted to the linear modulation region to ensure predictable torque generation, low harmonic distortion, and superior NVH performance. Consequently, overmodulation and six-step operation are generally avoided. [2-17], [2-13]

2.2.2.2.3.3 Synthesis of the Reference Voltage Vector

Within the linear modulation region, the reference voltage vector $\vec{V}_{ref}$ can be synthesized by time-averaging two adjacent active voltage vectors and the zero vector over one switching period. Assuming that $\vec{V}_{ref}$ lies in a given sector of the hexagon, it can be expressed as:

$$\vec{V}_{ref}T_s = T_1\vec{V}_k + T_2\vec{V}_{k+1} + T_0\vec{V}_0 \quad (2.16)$$

where:

- $\vec{V}_k$ and $\vec{V}_{k+1}$ are the two adjacent active voltage vectors defining the sector,
- $\vec{V}_0$ denotes the zero-voltage vector,
- T_1, T_2 , and T_0 represent the corresponding dwelling times, and
- $T_s = T_1 + T_2 + T_0$ is the switching period.

The dwelling times T_1 and T_2 are determined by the magnitude and angular position of $\vec{V}_{ref}$ within the sector, while T_0 represents the remaining time allocated to the zero vectors. In classical SVPWM implementations, the zero-vector time is typically symmetrically distributed at the beginning and end of the switching period to minimize switching harmonics.

Although the explicit computation of T_1, T_2 , and T_0 provides clear physical insight, it is not always necessary to calculate these quantities directly in practical implementations. As discussed in the following section, an equivalent formulation based on common-mode voltage injection can achieve the same average output voltage without explicit sector identification. [2-12], [2-13]

2.2.2.2.3.4 Implications for EPS Applications

The space vector formulation offers several advantages that make SVPWM particularly suitable for automotive EPS systems. First, the increased DC bus voltage utilization allows higher assist torque to be generated without increasing the inverter voltage rating. Second, the reduced harmonic distortion of the phase currents leads to lower torque ripple, which directly improves steering smoothness and NVH performance.

Furthermore, the unified treatment of the three-phase system simplifies the integration of SVPWM with field-oriented control. Since EPS systems operate over a wide range of speeds and loads, including prolonged operation at very low speeds, the predictable behavior of SVPWM in the linear region is a key advantage. These characteristics motivate the widespread adoption of SVPWM in modern EPS motor drives.

In the next section, an equivalent implementation of SVPWM based on common-mode voltage injection is presented. This formulation avoids explicit sector determination and dwelling time calculation, while remaining mathematically equivalent to the classical space vector approach described above. [2-17], [2-13]

2.2.2.2.4 Equivalent SVPWM Implementation Using Common-Mode Voltage Injection

Although the classical formulation of SVPWM described in Section 2.2.2.2.3 provides a clear physical interpretation of the inverter switching process, practical implementations often avoid explicit sector identification and dwelling time computation. Instead, an equivalent formulation based on common-mode voltage injection is widely adopted in modern digital motor control systems. This approach yields the same average output voltage as the classical SVPWM method, while offering improved numerical robustness and implementation simplicity. [2-13], [2-14]

2.2.2.2.4.1 Common-Mode Voltage Injection Principle

Starting from the reference phase voltages $v_a, v_b,$ and v_c obtained by the inverse Clarke transformation, SVPWM can be equivalently realized by adding a common-mode voltage v_{cm} to all three phases. The modified phase voltages are expressed as [2-13]:

$$\begin{aligned} v'_a &= v_a - v_{cm} \\ v'_b &= v_b - v_{cm} \\ v'_c &= v_c - v_{cm} \end{aligned} \quad (2.17)$$

The key requirement is that the resulting phase voltages $v'_a, v'_b,$ and v'_c remain within the inverter voltage limits imposed by the DC bus. This condition can be satisfied by selecting the common-mode voltage as [2-13], [2-14]:

$$v_{cm} = \frac{\max(v_a, v_b, v_c) + \min(v_a, v_b, v_c)}{2} \quad (2.18)$$

This choice shifts the three-phase voltage set such that the maximum and minimum phase voltages are symmetrically positioned with respect to the DC bus midpoint. As a result, the available DC bus voltage is fully utilized without violating the linear modulation constraint.

2.2.2.2.4.2 Mathematical Equivalence to Classical SVPWM

The equivalence between the common-mode voltage injection method and the classical space-vector-based SVPWM can be demonstrated by examining the resulting line-to-line voltages. Since the same common-mode voltage is added to all three phases, the line-to-line voltages remain unchanged [2-13]:

$$\begin{aligned} v'_{ab} &= v'_a - v'_b = v_a - v_b \\ v'_{bc} &= v'_b - v'_c = v_b - v_c \\ v'_{ca} &= v'_c - v'_a = v_c - v_a \end{aligned} \quad (2.19)$$

Because the electromagnetic torque of a three-phase motor depends on the line-to-line voltages or, equivalently, on the $\alpha\beta$ components of the voltage vector, the injected common-mode voltage does not alter the torque-producing components. Consequently, the average output voltage synthesized by the inverter is identical to that obtained using the classical SVPWM approach based on space vector dwelling times.

From a geometric perspective, the common-mode voltage injection effectively centers the reference voltage vector within the hexagonal voltage boundary defined by the inverter switching states. This operation implicitly accounts for the zero-vector time allocation that would otherwise be computed explicitly in the classical formulation. Therefore, the common-mode voltage injection method is mathematically equivalent to SVPWM in the linear modulation region. [2-14], [2-20]

2.2.2.2.4.3 Advantages for Digital Implementation

The common-mode voltage injection formulation offers several practical advantages that make it particularly attractive for digital control implementations in automotive EPS systems. First, it eliminates the need for explicit sector determination and trigonometric calculations associated with the computation of T_1 , T_2 , and T_0 . This simplification reduces computational complexity and improves numerical stability, especially under rapid transient conditions.

Second, the use of simple maximum and minimum operations facilitates efficient implementation on fixed-point processors and microcontrollers commonly used in automotive electronic control units (ECUs). This property is particularly important for EPS applications, where real-time performance, determinism, and robustness are critical.

Finally, the common-mode voltage injection approach naturally lends itself to code generation and model-based design environments. Industry-standard platforms, such as MathWorks Motor Control Blockset, adopt this formulation to ensure consistency between simulation models and embedded implementations. As a result, the behavior of the simulated SVPWM closely matches that of the deployed control software. [2-13], [2-14] [2-12]

2.2.2.2.4.4 Relevance to EPS Operating Conditions

In automotive EPS systems, SVPWM is typically operated strictly within the linear modulation region to guarantee smooth torque generation and minimal acoustic noise. The common-mode voltage injection method is well suited to this operating regime, as it provides continuous and smooth modulation waveforms without introducing discontinuities associated with overmodulation or six-step operation.

Moreover, EPS systems frequently operate at low motor speeds with relatively high torque demand, particularly during parking maneuvers. Under these conditions, precise voltage synthesis and low harmonic

distortion are essential to prevent torque ripple and steering vibration. The common-mode voltage injection formulation ensures that the reference voltage is optimally mapped onto the inverter voltage limits, thereby contributing to improved NVH performance and enhanced steering feel. [2-17], [2-18]

2.2.2.2.5 SVPWM Implementation in Automotive Electric Power Steering Systems

In practical EPS motor drive systems, the theoretical formulation of SVPWM must be translated into a numerically robust and computationally efficient implementation. This section describes a representative SVPWM implementation widely used in automotive EPS applications, with particular emphasis on its correspondence to the common-mode voltage injection formulation discussed in Section 2.2.2.2.4.

2.2.2.2.5.1 Structure of the Space Vector Generator

In a typical EPS motor control architecture based on field-oriented control, the voltage commands generated by the current controllers are first expressed in the rotating dq reference frame. These commands are then transformed into the stationary reference frame using the inverse Park transformation, yielding the voltage components v_α and v_β .

The Space Vector Generator converts these stationary-frame voltage components into three-phase modulation signals. This process consists of the following steps:

1. Inverse Clarke Transformation:

The stationary-frame voltage vector (v_α, v_β) is transformed into three-phase phase voltage references v_a, v_b , and v_c using the inverse Clarke transformation described in Section 2.2.2.2.2.

2. Common-Mode Voltage Computation:

The common-mode voltage v_{cm} is computed according to 2.2.2.2.4.1, based on the maximum and minimum values of the phase voltage references.

3. Phase Voltage Adjustment:

The common-mode voltage is subtracted from each phase voltage to obtain the modified phase voltages v'_a, v'_b , and v'_c , which satisfy the inverter voltage constraints.

This sequence directly corresponds to the equivalent SVPWM formulation presented in the previous section and avoids explicit computation of sector information or dwelling times. [2-20]

2.2.2.2.5.2 Normalization and Duty Cycle Generation

The modified phase voltages v'_a , v'_b , and v'_c obtained after common-mode voltage injection are normalized with respect to the DC bus voltage to form the modulation signals. In the linear modulation region, these normalized modulation signals are bounded within the interval $[-1,1]$.

To generate the PWM duty cycles for each inverter leg, the normalized modulation signals are linearly mapped to the range $[0,1]$ using the following relationship:

$$D_{a,b,c} = 0.5 v'_{a,b,c} + 0.5 \quad (2.20)$$

where D_a , D_b , and D_c , denote the duty cycles of the corresponding inverter phases. This transformation ensures that a modulation signal of zero corresponds to a duty cycle of 0.5, while positive and negative modulation values shift the duty cycle toward the upper or lower limits, respectively. The resulting duty cycles can be directly applied to the PWM hardware module of the inverter. [2-13]

2.2.2.2.5.3 Implementation Considerations for EPS Systems

In automotive EPS applications, the SVPWM implementation is designed with several system-level considerations in mind.

First, EPS systems prioritize smooth torque generation and low acoustic noise over maximum voltage utilization. As a result, the modulation index is deliberately limited to remain within the linear SVPWM region. This design choice avoids the nonlinear behavior associated with overmodulation and six-step operation, which can introduce torque ripple and audible noise.

Second, the computational simplicity of the common-mode voltage injection method is well suited to automotive ECUs. The use of basic arithmetic operations, such as maximum, minimum, addition, and scaling, enables efficient execution on fixed-point processors while maintaining deterministic real-time performance.

Third, the implementation integrates seamlessly with model-based design workflows. In industry-standard environments, such as MathWorks Motor Control Blockset, the same SVPWM structure is used in both simulation and embedded code generation. This consistency ensures that the control behavior observed during simulation accurately reflects the behavior of the deployed system, reducing calibration effort and development risk. [2-14], [2-21]

2.2.2.2.5.4 Consistency with EPS Operating Requirements

The SVPWM implementation described in this section satisfies the key operating requirements of automotive EPS systems. By ensuring continuous and smooth modulation waveforms, it minimizes current harmonic distortion and torque ripple, thereby improving steering smoothness and NVH performance.

Moreover, the predictable behavior of SVPWM in the linear modulation region supports stable control at very low motor speeds, which are commonly encountered during parking and low-speed maneuvers. This characteristic is particularly important in EPS systems, where abrupt torque variations can be directly perceived by the driver through the steering wheel.

In summary, the common-mode voltage injection implementation of SVPWM provides an effective and practical solution for automotive EPS motor drives, combining theoretical optimality with implementation robustness. [2-17], [2-22]

2.2.2.2.6 Discussion: Suitability of SVPWM for Automotive EPS Applications

The selection of a PWM strategy for automotive Electric Power Steering systems must be guided not only by theoretical optimality, but also by practical considerations related to steering feel, NVH performance, and system robustness. This section discusses the suitability of SVPWM for EPS applications, with particular emphasis on its advantages over alternative modulation methods such as six-step operation and overmodulation-based schemes. [2-17], [2-21], [2-22]

2.2.2.2.6.1 Comparison with Six-Step Operation

Six-step operation represents the extreme case of inverter modulation, in which the zero voltage vectors are eliminated and each phase voltage is clamped to the positive or negative DC bus for extended intervals. While this approach maximizes voltage utilization, it produces highly non-sinusoidal phase currents and introduces significant harmonic distortion.

In EPS systems, such behavior is undesirable for several reasons. First, the harmonic-rich current waveforms associated with six-step operation generate pronounced torque ripple, which is directly transmitted to the steering wheel and perceived by the driver as vibration or roughness. Second, the abrupt switching transitions inherent in six-step operation excite structural resonances in the steering column and gearbox, leading to increased acoustic noise.

By contrast, SVPWM operating in the linear modulation region produces smooth, near-sinusoidal current waveforms with substantially lower harmonic content. As demonstrated in the preceding sections, SVPWM achieves higher voltage utilization than SPWM while maintaining continuous modulation signals. This balance between voltage efficiency and waveform quality makes SVPWM fundamentally more suitable for EPS systems than six-step operation. [2-18], [2-23]

2.2.2.2.6.2 Avoidance of Overmodulation in EPS Systems

Overmodulation extends the voltage utilization of SVPWM beyond the linear region by intentionally distorting the modulation signals. Although this technique is commonly employed in traction drives and high-speed motor applications, it is generally avoided in EPS systems.

The primary reason lies in the sensitivity of steering feel to torque nonlinearity. In the overmodulation region, the relationship between the commanded voltage and the resulting motor torque becomes nonlinear and discontinuous. Such behavior can lead to unpredictable torque responses, particularly during low-speed maneuvers where EPS systems are most actively engaged.

Furthermore, overmodulation increases low-order current harmonics, which can amplify acoustic noise and mechanical vibration. Since EPS systems typically operate with sufficient DC bus voltage margin and prioritize smooth torque generation over maximum voltage output, the benefits of overmodulation are outweighed by its negative impact on NVH performance. Consequently, EPS controllers are commonly designed to restrict SVPWM operation to the linear modulation region. [2-17]

2.2.2.2.6.3 Impact on NVH Performance

Noise, vibration, and harshness are critical performance metrics for EPS systems, as any disturbance generated by the motor or inverter can be directly transmitted to the driver through the steering mechanism. The use of SVPWM contributes to NVH reduction in several ways.

First, the reduced harmonic distortion of phase currents lowers electromagnetic torque ripple, which is a primary source of vibration in electric motor drives. Second, the smooth variation of duty cycles inherent in SVPWM minimizes abrupt changes in inverter switching states, thereby reducing high-frequency excitation of mechanical components.

Additionally, the common-mode voltage injection formulation ensures continuous modulation waveforms without discontinuities at sector boundaries. This property further suppresses tonal noise components that could otherwise be amplified by structural resonances. As a result, SVPWM provides a favorable compromise between efficiency and acoustic performance for EPS applications. [2-18], [2-22]

2.2.2.2.6.4 Low-Speed Steering Performance

Low-speed operation represents one of the most demanding operating conditions for EPS systems. During parking maneuvers and low-speed steering corrections, the motor is required to generate high assist torque at near-zero speed. Under these conditions, any torque ripple or modulation-induced disturbance becomes highly perceptible.

SVPWM operating in the linear region enables precise voltage synthesis even at low electrical frequencies, ensuring stable current regulation and smooth torque output. The predictable behavior of the modulation scheme supports accurate torque control, which is essential for maintaining consistent steering feel during fine steering adjustments.

Moreover, the absence of discontinuities in the modulation signals reduces the likelihood of stick-slip phenomena and steering vibration. These characteristics make SVPWM particularly well suited for the low-speed operating regime that dominates EPS usage. [2-21]

2.2.2.2.6.5 Compatibility with ADAS and Future Steering Functions

Modern EPS systems increasingly serve as actuators for advanced driver-assistance systems, such as lane-keeping assistance and automated parking. These functions impose stringent requirements on steering precision, smoothness, and repeatability.

SVPWM supports these requirements by providing a linear and continuous mapping between voltage commands and inverter output. The resulting smooth torque response facilitates seamless interaction between driver input and automated steering interventions. In addition, the computational efficiency and determinism of the common-mode voltage injection implementation make SVPWM compatible with the real-time constraints of automotive control architectures.

As EPS systems continue to evolve toward higher levels of automation, the advantages of SVPWM in terms of controllability and NVH performance are expected to become increasingly important. [2-12], [2-21]

2.2.2.2.7 Summary and Comparative Discussion with SPWM

To clearly position the contribution of this chapter, a direct comparison between sinusoidal pulse width modulation (SPWM) and space vector pulse width modulation (SVPWM) is presented from the perspective of automotive Electric Power Steering (EPS) applications. The comparison highlights the fundamental differences in modulation characteristics and their practical implications for steering performance.

Characteristic	SPWM	SVPWM	Relevance to EPS
DC bus voltage utilization	Up to $V_{dc}/2$	Up to $\sqrt{3}/2 \cdot V_{dc}$	Higher assist torque without increasing DC voltage
Linear modulation range	Narrow	Wide	Stable torque control over wider range
Harmonic distortion	Relatively high	Lower	Reduced current harmonics

Torque ripple	Higher	Lower	Improved steering smoothness
Common-mode voltage handling	Implicit / not optimized	Explicitly optimized	Improved NVH characteristics
Low-speed performance	Sensitive to distortion	Stable and smooth	Critical for parking maneuvers
NVH characteristics	Inferior	Superior	Reduced audible noise and vibration
Implementation complexity	Low	Moderate	Acceptable trade-off for EPS
Computational suitability	Very low load	Slightly higher load	Well within ECU capability
Industrial adoption	Legacy / limited	Widely adopted	Standard choice for modern EPS

Table 2.2.2 Comparison of SPWM and SVPWM for Automotive EPS Applications [2-13], [2-17], [2-12] [2-24]

1) Discussion Based on the Comparative Analysis

As summarized in Table 2.2.2, SPWM offers simplicity but suffers from limited voltage utilization and relatively high harmonic distortion. These limitations are particularly problematic in EPS systems, where torque smoothness and NVH performance are directly perceived by the driver. In contrast, SVPWM provides superior voltage utilization and reduced harmonic content by exploiting the space vector representation of the inverter output voltage. [2-17], [2-13]

The wider linear modulation range of SVPWM allows EPS motor drives to operate with increased torque capability while maintaining linear and predictable control behavior. This characteristic is essential for EPS applications, which typically avoid overmodulation and six-step operation in order to preserve steering feel and minimize acoustic noise. By operating strictly within the linear SVPWM region, the motor drive achieves smooth torque generation without introducing nonlinearities that could degrade driver perception.

Furthermore, the explicit handling of common-mode voltage in SVPWM enables optimal centering of the phase voltages within the DC bus constraints. This property reduces low-order current harmonics and suppresses torque ripple, contributing to improved NVH characteristics. Such improvements are especially important at low motor speeds, where EPS systems deliver maximum assist torque and any disturbance is readily transmitted to the steering wheel. [2-14], [2-13]

From an implementation standpoint, the slightly increased computational complexity of SVPWM is justified by its performance benefits. Modern automotive ECUs are fully capable of executing the required calculations within real-time constraints, particularly when the common-mode voltage injection formulation is employed. This formulation avoids explicit sector determination and dwelling time computation, making SVPWM both robust and efficient for embedded implementation. [2-13]

2) Final Conclusion for EPS Applications

Based on the comparative analysis and the theoretical and practical considerations discussed throughout this chapter, SVPWM is clearly more suitable than SPWM for automotive EPS applications. SVPWM offers a superior balance between voltage utilization, torque smoothness, and NVH performance while maintaining predictable and stable behavior in the linear modulation region.

The SVPWM implementation adopted in this study, which is mathematically equivalent to the classical space-vector-based formulation and consistent with the MathWorks reference model, provides a validated and industry-aligned modulation framework. Its advantages directly address the key requirements of EPS systems, including smooth low-speed operation, reduced torque ripple, and compatibility with advanced steering and driver-assistance functions.

Therefore, SVPWM is selected as the modulation strategy for the EPS motor drive considered in this thesis, forming a robust foundation for the subsequent modeling, analysis, and control development. [2-13], [2-17]

2.2.3 Control

2.2.3.1 EPS control system block

In modern EPS configurations, permanent magnet synchronous motors (PMSMs) are increasingly preferred due to their enhanced performance efficiency and robust operational reliability. To achieve optimal torque output while mitigating non-linear behavior in motor torque characteristics, an advanced control strategy known as **vector control** is employed. This technique necessitates real-time computation involving both rotor position and stator phase current measurements. [2-20], [2-25]

Figure 2.2.3 depicts the core structure of the steering control unit and highlights its principal functionalities:

- Regulating and supervising supply voltages for all internal subsystems
- Acquiring and conditioning data from the steering torque and steering angle sensors
- Accepting signals related to vehicle and engine speed through CAN or FlexRay™ interfaces
- Interfacing with other systems such as the brake controller for additional command inputs

- Computing the assistive torque or force necessary for steering support
- Handling motor signals and executing torque control through vector-based methods [2-26]
- Generating and shaping PWM outputs to drive MOSFETs commonly utilized in 3-phase motor power stages

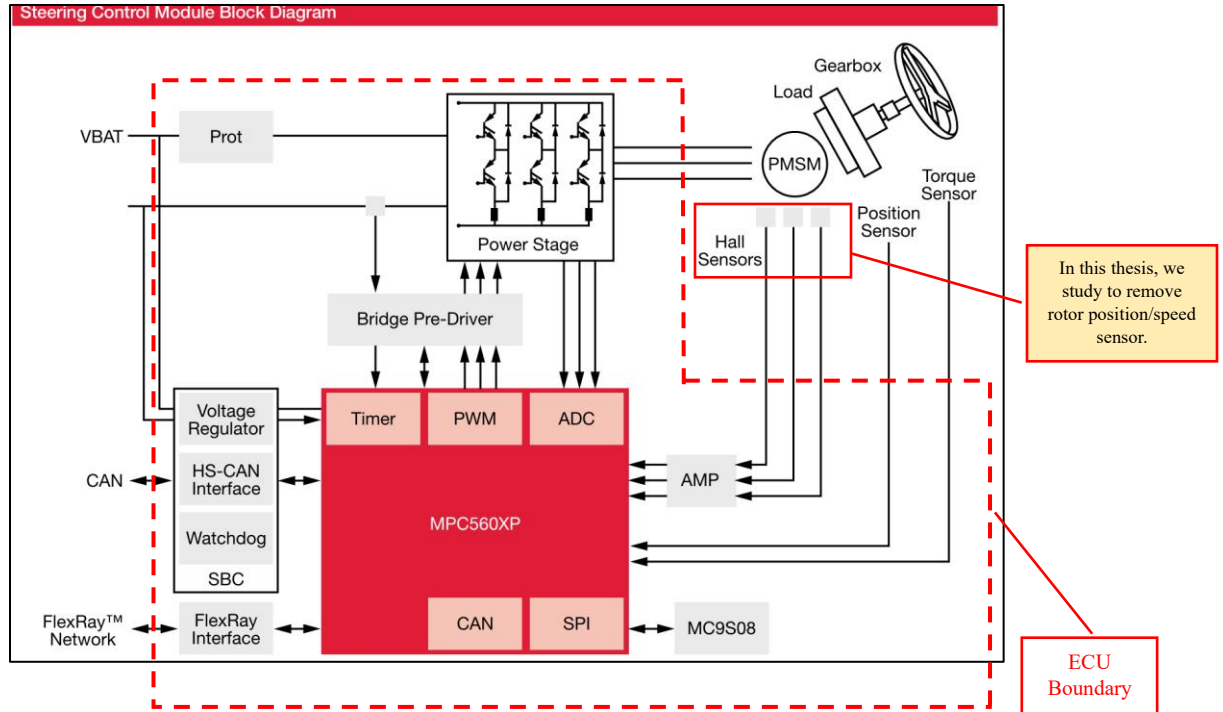


Figure 2.2.3 Complete block diagram of electric power steering (EPS) “NXP Semiconductors” [2-26], [2-27], [2-28]

2.2.3.2 Assist control summary - Function and Characteristics of Assist Curve (Boost Curve) in Passenger Vehicle EPS

What is the Assist Control and Assist Curve in EPS?

Figure 2.2.4 shows the assist curve (boost curve) in Electric Power Steering (EPS) systems represents the relationship between the steering torque applied by the driver (input torque) and the assist torque generated by the EPS motor. This characteristic curve is *fundamental* to EPS control, enabling a balance between steering ease and road feedback through appropriate design. Generally, the boost curve is designed to be symmetric about zero, ensuring consistent characteristics in both left and right steering directions.

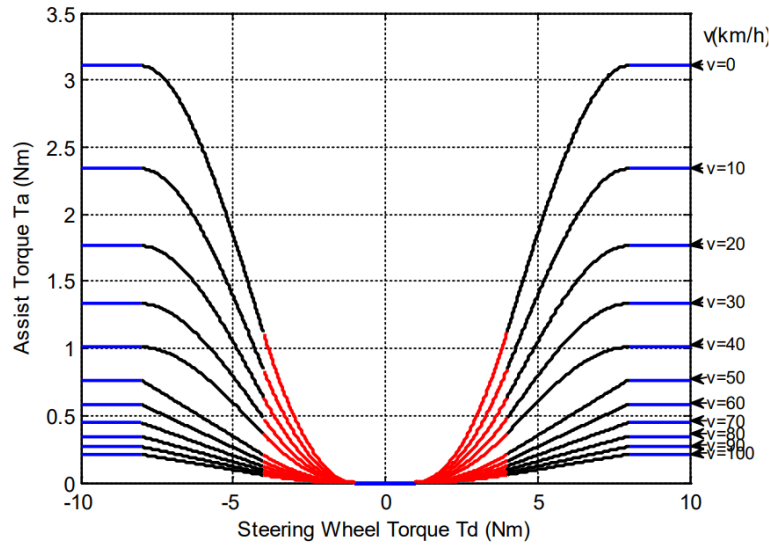


Figure 2.2.4 Conceptual Diagram of Assist Curve in EPS [2-26], [2-27], [2-28] [2-29]

In this diagram, the horizontal axis represents the driver's applied steering torque T_d , and the vertical axis represents the assist torque output by the EPS motor (T_{assist}). In the low steering torque region, there is a deadband T_{d0} where no assist is provided ($T_{assist} = 0$). Beyond this threshold, the assist torque increases linearly with the input torque until it reaches a saturation point (T_{sat}) corresponding to the motor's maximum torque. This saturation level varies with vehicle speed, providing strong assist at low speeds (e.g., during parking) and reduced assist at high speeds for better road feel and stability.

The assist curve is typically implemented within the EPS control unit as a **lookup table (characteristic map)**. Based on the detected steering torque from torque sensors and vehicle speed signals, the system references this table to determine the target assist torque. The motor then receives the appropriate current to generate the assist torque through a reduction mechanism, aiding the steering system. Unlike hydraulic power steering (HPS), where the boost curve is mechanically determined by valve design, EPS allows for flexible adjustment of the curve shape through software, enabling fine-tuning and optimization for different vehicle models. [2-26]

Speed-Dependent Assist Characteristics and Their Rationale

EPS systems adjust the assist curve based on vehicle speed. Typically, as vehicle speed increases, the assist torque decreases to maintain stable steering feel at high speeds. Conversely, at low speeds (e.g., during parking or low-speed cornering), higher assist torque is provided to reduce the driver's steering effort. This

speed-sensitive control ensures that steering feels "**light**" at low speeds and "**firm**" at high speeds, balancing safety and maneuverability. [2-27], [2-28]

Specifically, the EPS control map includes different boost curves for various speed ranges, such as "parking speed," "urban speed," and "highway speed." For instance, at low speeds, even small steering inputs result in significant assist torque, while at high speeds, the same input yields reduced assist, enhancing road feel. Some studies have presented typical boost curves under different conditions, showing that assist torque is higher during parking and significantly limited during highway driving. By interpolating between these curves, a continuous three-dimensional assist characteristic is achieved across all speed ranges. [2-27]

Manufacturers often define target assist gains (or maximum assist torque values) at specific speeds (e.g., 0 km/h and 100 km/h) and adjust the assist level linearly between these points based on vehicle speed. The primary reason for implementing speed-dependent control is to prevent excessive steering lightness at high speeds, which could lead to over-sensitive vehicle responses and reduced straight-line stability. Additionally, excessive assist can diminish road feedback, making it harder for drivers to sense tire-road interactions. [2-28]

Therefore, the boost curve's slope (gain) and saturation level are reduced at higher speeds to retain appropriate steering resistance. Conversely, at low speeds, higher assist torque allows for effortless steering, even in heavy passenger vehicles. Most commercial EPS systems adopt this "full boost at low speeds, restrained at high speeds" strategy.

Key Points in Assist Curve Design

1. **Balance of Steering Feel:** As previously mentioned, the assist curve must balance "steering lightness" and "road feedback," which are often conflicting requirements. Designers adjust the curve shape to ensure easy maneuverability during parking and stable steering feel at high speeds. Ideally, the assist gain is high at low speeds and reduced at high speeds, maintaining consistent steering feel across different speeds. Additionally, on-center steering feel is crucial. To reduce play and enhance straight-line stability, assist is often suppressed near the center position (deadband setting). Conversely, if internal friction causes heavy steering, slight assist may be added even in low torque regions to compensate (friction compensation). Adjusting gain separately for on-center and off-center positions helps balance straight-line stability and agility during turns. [2-26], [2-28]
2. **Smooth Characteristics:** The boost curve should be continuous and smooth, avoiding abrupt gradient changes or discontinuities. Traditional piecewise linear approximations can cause sudden changes in assist torque at segment functions, leading to unnatural steering feel. Modern EPS

controls employ quadratic curves or spline interpolation to create smoother curves, ensuring natural steering feel across all steering inputs. [2-26] For example, some patents propose constructing the boost curve using a combination of quadratic and linear functions, defined by just six parameters, allowing for smooth curves with minimal memory usage and facilitating continuous shape changes with vehicle speed.

3. **Safety and Stability:** In assist curve design, it's crucial to avoid excessive assist. If the EPS motor's assist torque exceeds the steering system's restoring torque (e.g., from tire self-aligning torque), it may cause the steering wheel to return abruptly or amplify vibrations. Therefore, the assist torque's upper limit is set below the steering system's resistance torque. Additionally, when the driver releases the steering wheel after a sharp turn, the assist torque should be limited to a level that does not exceed the inherent restoring torque of the steering system, in order to prevent unstable steering behavior and ensure safe and predictable steering response. [2-28]

2.2.3.3 Motor control block

Figure 2.2.5 shows motor control block diagram of EPS. Vector-based motor control aims to synchronize the magnetic fields of the stator and rotor in a Permanent Magnet Synchronous Motor (PMSM) to maximize torque generation. The most effective configuration occurs when these two fields maintain a perpendicular orientation — precisely 90 degrees apart. The method used to enforce and maintain this optimal 90-degree alignment is widely known as Field Oriented Control (FOC). [2-30]

In standard PMSM control systems, the typical execution loop operates on a 50-microsecond interval. [2-30] Within each of these control cycles, the following key operations are usually carried out:

- Acquisition and computation of phase current and DC bus current signals
- Interpretation of encoder or resolver feedback to determine rotor position
- Conversion of motor current components (i_d , i_q) through Clarke and Park transforms
- Execution of the current control logic
- Regeneration of updated PWM signals via inverse Park and Clarke transforms

These control processes — combined with compliance requirements for functional safety systems — present several design challenges for engineers developing electric power steering (EPS) control units. [2-30]

High-speed, accurate state variable acquisition

Because of the tight timing constraints imposed by the PWM loop, the system must incorporate an analog-to-digital converter (ADC) that can deliver rapid and accurate current measurements. The ADC should be capable of capturing both DC bus current and phase currents with sufficient speed to perform oversampling within a single PWM cycle. Common specifications for such ADCs include a maximum conversion time of 1.5 microseconds or less, with a resolution of at least 9 bits. [2-31]

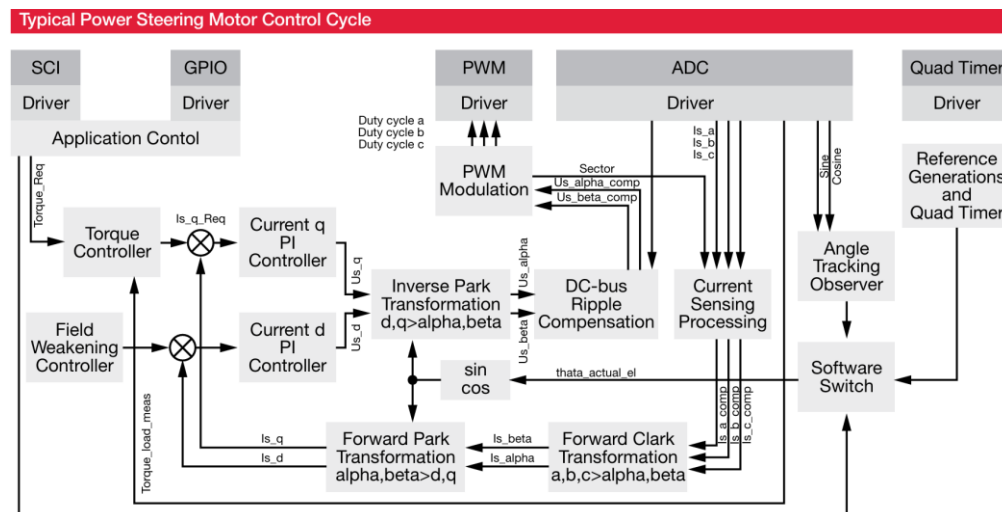


Figure 2.2.5. Motor control block diagram of EPS “NXP Semiconductors” [2-30]

2.2.4 Software architecture

Electric Power Steering (EPS) is a system driven by an electric motor that assists steering operations, featuring lighter weight and higher efficiency compared to traditional hydraulic power steering systems. Since the world's first mass production of EPS in 1988, the system has contributed significantly to fuel efficiency improvements and the realization of advanced driver assistance functions, serving as a critical component in performing the vehicle's basic functions of “driving, steering, and stopping”.

In particular, because the steering system may cause severe hazards in the event of malfunction, the EPS ECU (Electronic Control Unit) is subjected to the highest safety requirements. According to the international functional safety standard ISO 26262, Automotive Safety Integrity Level D (ASIL-D) is applied [2-6]. In reality, EPS is one of the typical systems where malfunctions could result in extremely hazardous situations, thus requiring compliance with the most stringent ASIL-D criteria.

Meanwhile, in recent automotive system developments, not only functional safety but also cybersecurity considerations have become indispensable. Even in control systems prioritizing safety, such as steering, countermeasures against cyber threats are now required. [2-40]

This study focuses on in-vehicle EPS systems, and based on academic literature, conducts a comprehensive review from the following four perspectives: (1) The development process based on Automotive SPICE (ASPICE) and its application examples for EPS, (2) Implementation and utilization examples of standardized operating systems and development environments using AUTOSAR, (3) The role and function of EPS systems within the CAN network, (4) Challenges and countermeasures regarding cybersecurity.

2.2.4.1 Development Process Based on ASPICE and Application Cases in EPS

Automotive SPICE (ASPICE) is a process assessment framework specifically developed for automotive software and system development. Rather than prescribing technical solutions, ASPICE defines how development activities should be structured, executed, and documented in order to ensure process quality, traceability, and repeatability across organizations. It is widely used by OEMs and Tier 1 suppliers as a benchmark for evaluating supplier process maturity.

In the context of Electric Power Steering (EPS), ASPICE compliance is particularly significant. EPS is classified as a safety-critical system—often up to ASIL-D under ISO 26262—and therefore requires rigorous development discipline. Process consistency, requirement traceability, verification planning, and defect management must be demonstrable across the entire lifecycle. For this reason, ASPICE-aligned development has become effectively mandatory in many automotive programs. [2-6], [2-32] (Appendix A) (Appendix E)

2.2.4.1.1 Process Reference Model and V-Model Integration

ASPICE organizes development activities into a structured Process Reference Model (PRM). Processes are grouped into three major categories:

1. Primary life cycle processes
(e.g., system requirements, software requirements, design, integration, testing)
2. Organizational life cycle processes
(e.g., quality management, project management)
3. Supporting life cycle processes
(e.g., configuration management, problem resolution)

Each process includes:

- A defined purpose statement
- Expected outcomes

- Base practices that must be performed
- Work products that must be produced

ASPICE aligns naturally with the V-model development structure, which is commonly used in safety-critical automotive systems. In EPS development, this structure can be summarized as follows:

Left side of the V (decomposition and specification):

- SYS.1 / SYS.2: System requirements analysis
- SYS.3: System architectural design
- SWE.1: Software requirements specification
- SWE.2 / SWE.3: Software architectural and detailed design

Right side of the V (integration and validation):

- SWE.4 / SWE.5: Software integration and testing
- SYS.4 / SYS.5: System integration and system testing

This structure ensures that every requirement defined at the system level (SYS) is cascaded down to the software level (SWE), implemented, and then verified at the corresponding level of integration.

For EPS systems, this cascading process is particularly important. Steering assist behavior, fault detection thresholds, torque limits, and safety reactions originate at the vehicle/system level and must be traceable down to software design and code implementation. ASPICE therefore enforces bidirectional traceability from system requirements to software implementation and back to verification evidence.

Capability Levels and Maturity Evaluation

ASPICE evaluates process capability across maturity levels from Level 0 (Incomplete) to Level 5 (Optimizing).

For most automotive suppliers:

- Level 2 (Managed) indicates that processes are planned and monitored.
- Level 3 (Established) indicates standardized and organization-wide defined processes.

In modern EPS development, achieving at least ASPICE Level 2 or 3 is often required for supplier qualification. This means that beyond technical correctness, development artifacts must demonstrate:

- Defined requirements
- Traceability matrices

- Verification evidence
- Controlled configuration management
- Structured defect handling

Relevance to EPS Architecture and Safety

Because EPS is an ASIL-D system, ASPICE process compliance complements ISO 26262 safety lifecycle activities. While ISO 26262 defines safety goals and safety mechanisms, ASPICE ensures that the development process itself is capable of reliably implementing those safety requirements.

In practice, this means:

- Hazard Analysis and Risk Assessment (HARA) results are translated into system requirements.
- System requirements are decomposed into software requirements.
- Software architecture reflects safety allocation.
- Verification activities are planned and documented.
- All artifacts are traceable and auditable.

Thus, ASPICE provides the structural backbone that ensures EPS control software—including sensorless estimation logic—can be developed, verified, and maintained in a systematic and reproducible manner.

For the process dimension, the Automotive SPICE process reference model provides the set of processes shown in Figure 2.2.6 “Automotive SPICE process reference model – Overview”

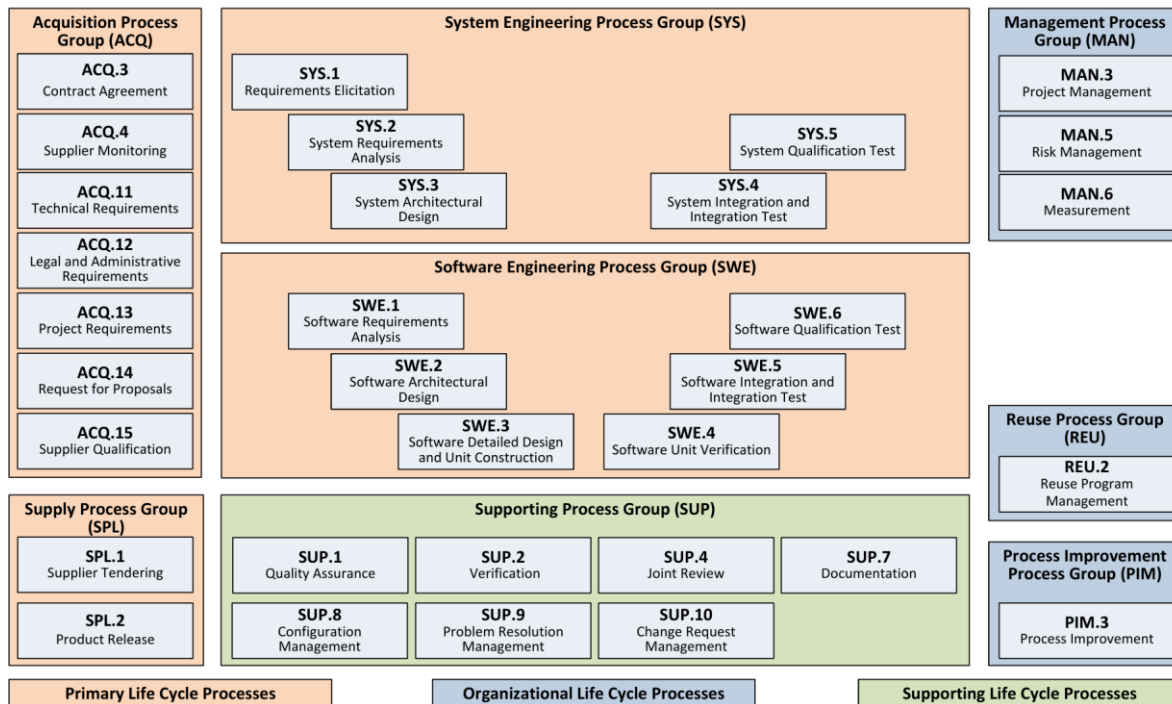


Figure 2.2.6 Automotive SPICE process reference model – Overview

[2-6], [2-32] (Appendix A) (Appendix E)

Example of Development Process Application:

In the EPS development process of European major supplier HELLA (now part of the FORVIA Group), system development was conducted in accordance with the ASPICE process reference model. With the support of items, independent activities such as system requirements analysis (SYS.2), architectural design (SYS.3), and safety analyses using FMEA/FTA were carried out, achieving a methodologically consistent development approach across the process. In this case, ISO 26262 (Functional Safety) and ISO/SAE 21434 (Cybersecurity) were also taken into consideration as requirements, and the EPS development process was integrally managed from both safety and security perspectives. Similarly, Hitachi Astemo, a major Japanese automotive parts supplier, has introduced ASPICE-compliant processes in EPS product development, contributing to the "prevention of rework caused by missing requirements" and the "promotion of safe and reliable product development". Thus, the application of ASPICE-based development processes is a key factor in enhancing the quality and reliability of EPS development and meeting assessment requirements from European OEMs.

2.2.4.2 Implementation and Application Cases of Standard OS and Development Environment and Platform with AUTOSAR:

Modern EPS ECUs increasingly adopt software architectures conforming to the AUTOSAR Classic Platform. AUTOSAR is a standardized framework for automotive software, providing layers from Microcontroller Abstraction Layer (MCAL) to Basic Software (BSW), standard OS (OSEK-based real-time OS), and Runtime Environment (RTE). [2-33] The adoption of a standard OS allows for unified implementation of task scheduling, memory protection, etc., enabling developers to concentrate on implementing application functionalities. Moreover, compliance with AUTOSAR facilitates the integration of software components across different ECUs, improving reusability and scalability. (Appendix B)

Application Cases in EPS Development:

The AUTOSAR-compliant platform "Julinar," developed by Japanese startup Advanced Platform Technology (APT), was adopted by JTEKT's EPS and installed in the steering control system of a commercial SUV (North American market, new Nissan Rogue model). Julinar achieves highly accurate inter-software synchronization and coordination with low overhead and is equipped with a high-performance real-time OS. By combining JTEKT's functional safety expertise with APT's AUTOSAR platform technology, an EPS control software system was realized that meets stringent safety requirements (ASIL-D) and seamlessly integrates with other vehicle modules. As a result, issues such as long processing times were addressed, contributing to the performance and reliability improvement of EPS. [2-16], [2-10]

Additionally, Tata Elxsi in India developed a Virtual ECU (V-ECU) for EPS using Vector's MICROSAR Basic Software, enabling early validation in a Software-in-the-Loop (SIL) environment. In this process, an AUTOSAR OS and BSW configured to run on a PC were used to execute application code identical to that on the real ECU at Level-3 V-ECU. Hardware-dependent modules such as MCAL and OS parts were replaced with simulation modules, and settings for scheduling and CAN communication were imported via AUTOSAR tools (dSPACE SystemDesk), achieving behavior closely resembling an actual ECU. This virtual EPS ECU demonstrates realistic responses on the CAN bus, including standby behavior where activation only begins upon receiving a Network Management (NM) message. Consequently, compared to methods using real ECU binaries, significantly faster simulations were made possible, scaling up the validation of EPS control functions. From these cases, it is evident that the use of AUTOSAR standard OS and development environments greatly contributes to improving software design efficiency and ensuring reliability in EPS development. Figure 2.2.7 shows AUTOSAR/ MICROSAR software stack layer/architecture.

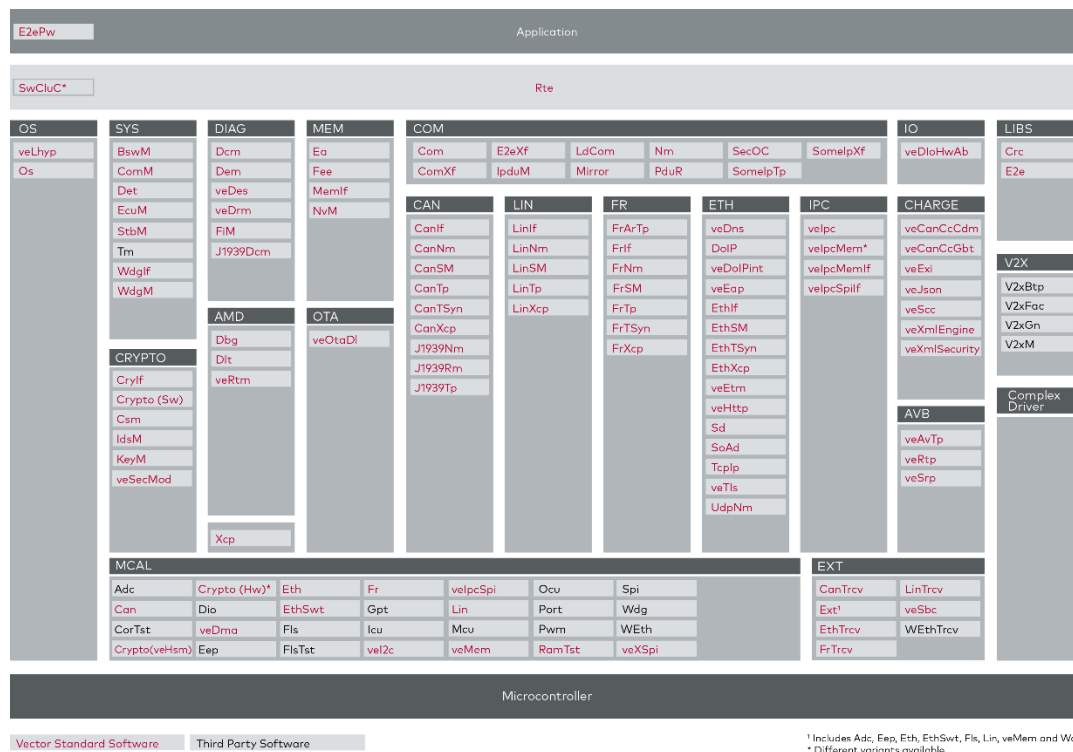


Figure 2.2.7 MICROSAR Architecture/Embedded-Software from Vector Available for AUTOSAR Classic Platform 4.3.0 [2-34]

2.2.4.3 The Role and Utilization of CAN Network Systems in EPS and In-Vehicle Networks:

Elements constituting the Electric Power Steering system include the motor, various sensors (such as motor angle sensors and torque sensors), the ECU itself, and various signals transmitted via CAN. [2-36]

The EPS ECU is connected to other ECUs within the vehicle through the CAN bus, receiving and utilizing necessary information for steering control (e.g., vehicle speed, steering angle) from them. In fact, typical EPS control logic modulates assist force according to vehicle speed, with vehicle speed signals provided via CAN messages from ECUs like ABS/ESC (Brake Control). (Appendix C)

In EPS integrated with Advanced Driver Assistance Systems (ADAS) such as lane-keeping support, steering torque commands are also sent via CAN from the ADAS ECU. [2-37]

Thus, the CAN network serves as a lifeline for EPS to obtain external environmental information and higher-level control commands.

Furthermore, EPS is responsible for transmitting its own status information to the entire vehicle network via CAN. Failures or abnormalities detected internally by the EPS ECU are notified to the vehicle system

controller or instrument cluster as DTCs (Diagnostic Trouble Codes), triggering driver warnings such as warning light illumination.

Actual studies identify the following as critical communications within EPS systems:

- Signals from steering input torque sensors to EPS ECU and fault detection signals
- Brake pedal position signals (for EPS with 4-wheel steering)
- Torque command request signals from other ECUs (ADAS or vehicle stability control)
- Vehicle speed signals or wheel speed signals
- Yaw rate sensor signals
- EPS diagnostic fault signals from EPS ECU to vehicle system controller
- Driver warning signals (e.g., warning lamp control)
- Low voltage anomaly signals and communication bus error detection signals

These examples illustrate that signals related to EPS over CAN communication have extremely high importance, directly affecting vehicle safety and control.

Therefore, ensuring communication reliability is indispensable. For example, Bosch's latest EPS product, Servolectric®, supports extended specifications like CAN-FD and high-speed buses like FlexRay, allowing flexible network selection based on communication bandwidth and redundancy requirements. [2-37]

Moreover, EPS ECUs participate in the vehicle's Network Management (NM) protocol, staying in low-power mode until an NM message is received on the bus and controlling activation accordingly. [2-39], [2-38]

In case of communication interruptions or anomalies on the CAN bus, the EPS ECU is designed to shift to a fail-safe mode, maintaining minimum mechanical steering functionality. On the other hand, detecting unauthorized communications as part of cybersecurity measures has become a new challenge (see 2.2.4.4).

2.2.4.4 Cybersecurity Related to EPS (Security Requirements, Risks, and Measures)

Intersection of Safety and Security:

EPS, being an actuator that directly controls vehicle motion, poses extreme risks if maliciously interfered with externally. For instance, if an attacker sends unauthorized steering commands to the vehicle's CAN, unintended steering actions could occur, potentially leading to serious accidents. [2-41]

Thus, today's automakers and suppliers strictly impose cybersecurity assurance as a mandatory requirement in the development of safety-critical systems like EPS. [2-40]

The international standard ISO/SAE 21434:2021 establishes a cybersecurity risk management process in vehicle development, requiring a complete sequence of activities, including item definition, Threat Analysis and Risk Assessment (TARA), and definition of security requirements for systems like EPS [2-40]. (Appendix D)

Anticipated Risks and Requirements:

Typical cybersecurity risks associated with EPS include "unauthorized commands via the in-vehicle network" and "tampering of ECU firmware."

For the former, if external access to the CAN bus is obtained, there is a risk of EPS control messages being forged or altered. To address this, the AUTOSAR standard introduces Secure Onboard Communication (SecOC), standardizing mechanisms that append Message Authentication Codes (MAC) to CAN messages to verify sender authenticity. In the 2021 Toyota RAV4 Prime, EPS supports SecOC, with internal EPS ECU processing of MACs based on AES encryption reported. [2-41], [2-42], [2-43]

For the latter risk — ECU tampering — there is concern about unauthorized access and reverse engineering of the EPS ECU, potentially leading to extraction and modification of internal encryption keys and programs. Indeed, a white-hat hacker group exploited vulnerabilities in the Renesas RH850 microcontroller's bootloader, successfully extracting SecOC keys via memory dumps. Consequently, they demonstrated the ability to forge steering control messages with valid MACs, breaching EPS communication protection. [2-43], [2-44]

This case highlights the critical importance of ensuring robustness in automotive ECU bootloaders and cryptographic implementations, including Secure Boot and the use of Hardware Security Modules (HSMs). [2-40]

Countermeasures and Trends:

To address these risks, multi-layered cybersecurity measures are being implemented in EPS development. Besides SecOC-based message authentication, the introduction of Intrusion Detection Systems (IDS) to detect abnormal communication patterns is under consideration.

The U.S. NHTSA (National Highway Traffic Safety Administration) also recommends that communication buses in EPS systems "detect unauthorized access and notify the EPS control module" as a functional safety requirement. [2-45]

This reflects the growing need to detect cyberattack signs and ensure safe system operation in modern automotive systems where safety and security boundaries are increasingly blurred.

Furthermore, during ECU software updates, secure boot and OTA (Over-the-Air) update mechanisms with digital signature verification have become mandatory, and EPS ECUs must be designed to prevent the activation of tampered firmware.

Additionally, with the advent of Level 3 and above autonomous driving systems, integrated safety and security designs are being pursued, including adoption of Steer-by-Wire systems and redundant designs. Specific methods include fail-operational architectures allowing safe control even if one path is attacked or fails, and storing cryptographic keys securely within HSMs. [2-47], [2-40]

In conclusion, cybersecurity for EPS is now as critical as functional safety in development, and the application of standardized practices and the latest protective technologies is rapidly advancing.

2.2.5 Functional safety-A survey of passenger vehicle EPS functional safety and ISO 26262 requirements

2.2.5.1 System Architecture and Control Structure of EPS

Electric Power Steering (EPS) is a system that assists steering operations using an electric motor, replacing conventional hydraulic power steering driven by an engine-mounted pump. EPS typically consists of the following major components:

Steering Wheel and Column: These transmit the driver's steering input. A torsion bar inside the column measures the steering torque via mechanical twist.

Torque Sensor: Mounted on the steering column, this sensor converts the torsional deformation of the torsion bar into an electrical signal representing steering torque. Dual-sensor configurations are commonly used to ensure redundancy for functional safety.

Electronic Control Unit (ECU): The ECU, centered around a microcontroller, processes steering torque and angle signals, and receives additional information such as vehicle speed and engine RPM from other ECUs via CAN or FlexRay. It calculates the optimal assist torque and generates PWM signals to control the motor through an integrated inverter. Typically, field-oriented control (FOC) of brushless motors is used. Rotor position feedback is obtained via Hall sensors or magnetic encoders. The ECU may also coordinate with ADAS functions such as lane-keeping assist or parking assist.

Electric Motor: A brushless DC or PMSM motor provides assist torque to the steering mechanism. The motor is connected to the steering rack via reduction gears (e.g., worm gear, belt drive, or rack-and-pinion). The ECU is often integrated into motor housing for compact packaging.

Communication Interface: The EPS ECU connects to the vehicle network (e.g., CAN FD or FlexRay) to receive external signals (e.g., vehicle speed, steering requests from ADAS) and to transmit diagnostic information.

As a drive-by-wire system, EPS plays a key role in vehicle maneuvering and is now widely adopted in modern vehicles due to its high energy efficiency and controllability. Furthermore, EPS is highly compatible with ADAS features, enabling precise steering support during autonomous or semi-autonomous driving.

[2-45]

2.2.5.2. Functional Safety Goals and Safety Mechanisms Based on ISO 26262

Excessive or Insufficient Assist: A deviation between intended and actual assist torque. For example, too much assist may cause oversteering, while too little assist may lead to driver fatigue or delayed response. These faults are typically classified as ASIL C.

Sudden Loss of Assist: The EPS system unexpectedly stops assisting. Although manual steering remains possible, the abrupt change in steering effort—especially at high speed—can be dangerous. This may be rated between ASIL B and C, depending on detectability and controllability.

Among these, unintended steering and steering lock are considered the most critical hazards and are almost universally addressed at the ASIL D level.

To achieve compliance with these safety goals, EPS systems incorporate several safety mechanisms:

Fault-Tolerant Architecture: EPS designs follow ISO 26262 principles by incorporating redundancy to tolerate single-point failures. For example, torque sensors are often duplicated, and dual-core or dual-MCU setups are used for mutual monitoring.

Use of High-Reliability Components: Safety-critical elements such as sensors and semiconductors are selected for their reliability and fail-safe behavior under harsh conditions. Mechanical parts, like gears and shafts, are also designed to avoid jamming or breakage throughout their service life.

Fault Detection and Fail-Safe Behavior: The ECU continuously monitors sensors and internal states. Upon detecting faults such as sensor mismatch, motor current anomalies, or internal errors—the system transitions to a safe state, limits assist torque, logs the event, and notifies the driver.

Functional Safety-Compliant Microcontrollers: Modern EPS ECUs use ISO 26262-qualified MCUs featuring lock-step cores, ECC-protected memory, and integrated diagnostic circuits. Software is developed in compliance with ASIL-D processes and includes safeguards such as torque limiters, relay-controlled power cutoffs, and redundant signal paths.

These design strategies ensure that EPS systems meet stringent ASIL D requirements. According to industry documentation (e.g., Analog Devices), meeting such targets requires system-level compliance involving redundant signal paths, cross-checking mechanisms, and robust monitoring from sensor input to torque output.

2.2.5.3. Mainstream Redundancy Architectures in EPS Systems

To satisfy ISO 26262 functional safety requirements - especially for ASIL D targets - redundancy is essential in EPS design. Modern EPS systems widely adopt redundant architectures across sensors, ECUs, power

stages, and actuators. Typical redundancy (Figure 2.2.9 Redundancy Architectures -Full Dual Architecture) implementations include:

Redundant Sensors: Steering torque sensors commonly employ dual sensing elements—either two identical sensors or heterogeneous types (e.g., magnetoresistive and Hall-effect). Their outputs are constantly cross-checked to detect discrepancies. Steering angle may also be validated by comparing motor angle feedback with torque-derived estimates. Vehicle speed and other critical signals are often received via dual communication paths for redundancy.

Redundant ECU Architectures: Microcontrollers often feature dual cores operating in lockstep mode or fully independent dual CPUs. Mismatches in computation results trigger fault responses. Dual power supply paths ensure that even if one regulator fails, the system can continue operating. Some EPS ECUs also implement dual CAN channels to maintain communication or separate safety messages.

Redundant Power Stages: The power inverter driving the motor is a key point of redundancy. One common design is a dual inverter with dual windings, where a single motor has two isolated stator windings, each driven by an independent inverter. This allows continued operation at reduced assist levels (typically ~50%) in the event of a fault in one channel.

Dual Motor Configurations: Another approach uses two independent motors—either both on the rack (dual-pinion EPS) or with one on the column and one on the rack. If one motor fails, the other can still provide assist torque. This physical redundancy ensures that no single fault leads to total assist loss.

Fail-Operational Control Strategies: Using the above hardware redundancies, EPS systems are designed to continue operating (at reduced capability) even after a fault is detected. For example, if one torque sensor fails, the system may fall back to the other sensor while alerting the driver. Similarly, if one inverter path fails, the remaining channel can still provide partial assist. These strategies help avoid abrupt changes in steering feel and allow the driver to maintain control until safe stopping.

The degree of redundancy depends on system requirements and cost targets. Some designs employ full duplication of ECUs and actuators, while others rely on partially redundant cores and sensors. In all cases, the design must demonstrate that required ASIL levels can be achieved with technical independence between primary and backup systems. [2-47], [2-19]

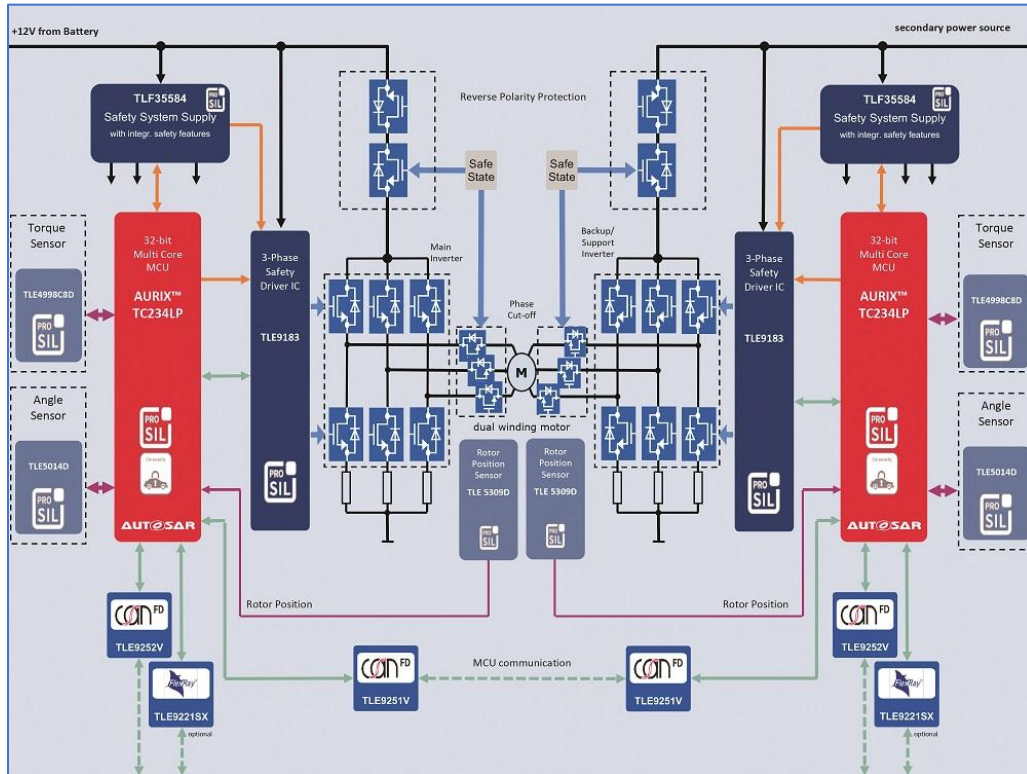


Figure 2.2.9 “Infineon” EPS Redundancy Architectures -Full Dual Architecture [2-19]

2.2.5.4. EPS-Specific Functional Safety Challenges and Mitigation Strategies

As EPS directly interfaces with the driver's steering input, maintaining consistency with the driver's intent is a critical safety requirement. EPS systems face unique challenges in functional safety due to their integration with human control and advanced driver assistance systems (ADAS).

Key challenges and countermeasures include:

Coordination with ADAS Steering Intervention:

Modern features like Lane Keeping Assist (LKA) and automated parking require EPS to apply steering torque autonomously (torque overlay). However, unintended or excessive intervention may conflict with the driver's intent and pose safety risks.

To address this, EPS systems limit the torque magnitude accepted from external systems and prioritize driver override. In practice, many production vehicles limit ADAS steering torque to around 2–3 Nm, ensuring the driver can easily counteract it.

While this conservative design improves safety, it may restrict ADAS functionality. As a solution, driver interaction detection and dynamic torque limit adjustment are under development to provide strong assist when needed and immediate handover to the driver otherwise.

A core principle is to ensure that the driver always retains ultimate steering authority.

Steering Feel and Human-Machine Interface (HMI):

EPS affects the tactile feedback the driver receives through the steering wheel. An unnatural or inconsistent feel can reduce trust and lead to unsafe driving. EPS control must filter out unnecessary vibrations while preserving essential road feedback.

During ADAS interventions, EPS should apply torque smoothly, avoiding abrupt or unpredictable movements. The emerging ISO 21448 standard (SOTIF) addresses such interactions by focusing on safety of the intended function, including usability and perception-related risks. [2-46]

Maintaining Drivability in Fault Conditions:

In the event of EPS failure, the driver must still be able to steer the vehicle. Most EPS designs retain a mechanical link, allowing manual steering even if assist is lost.

However, sudden loss of assist—especially at high speeds—can cause dangerous handling changes. Therefore, EPS systems aim to gradually reduce assist or use backup channels to maintain limited torque output.

For example, fail-operational EPS can continue delivering ~50% assist through the redundant path, allowing smooth transition and safe stopping. Even in non-redundant systems, the ECU may issue warnings and apply speed-based assist limits to support manual control.

Mechanical lock-up is categorically avoided through robust design of shafts, gears, and motor power-off mechanisms.

In summary, EPS development must ensure seamless coordination between human steering behavior and electronic control, even in degraded or autonomous modes. Simulation and HIL testing are widely employed to validate not only fault detection and fail-safe behavior, but also driver acceptance, comfort, and safety perception under various scenarios.

2.2.5.5. Industry Trends, Standardization, and Academic Research

The development of EPS systems is evolving rapidly in response to growing demands for functional safety, redundancy, and autonomous driving readiness.

Recent industry trends include:

Fail-Operational EPS for Autonomous Driving:

Autonomous vehicles at SAE Level 3 or higher require steering systems that remain functional even after a fault. In 2017, Bosch introduced the world's first fail-operational EPS, capable of maintaining ~50% assist via redundant actuation paths when one system fails. The system has since been commercialized as a modular kit.

Japanese supplier JTEKT has also developed a dual-redundant EPS control unit (FOPS) with independent ECUs and redundant motors to meet similar safety goals. [2-47]

Evolving Standards and Broader Safety Scope:

The second edition of ISO 26262, released in 2018, includes enhanced guidance on semiconductor safety and ADAS applications. EPS development follows this standard with structured safety lifecycle activities: hazard analysis (HARA), functional safety concept, technical safety requirements, and verification.

Moreover, EPS design is increasingly influenced by ISO 21448 (SOTIF) for non-fault-based risks and ISO/SAE 21434 for cybersecurity, especially in connected and autonomous scenarios.

Patent Activity:

Recent patents focus on safe steering intervention algorithms and redundant motor control architectures. For example, Volvo has filed patents for dynamically limiting ADAS steering torque to around 0.5 Nm by default, with controlled escalation based on driver monitoring and environment assessment.

Multi-phase motor designs (e.g., 6-phase) have also been patented to ensure fault-tolerant torque generation even when some windings are damaged.

Academic Research Directions:

Studies are addressing model-based fault detection and fault-tolerant control, such as estimating torque using motor currents when torque sensors fail.

Another research focus is steer-by-wire (SBW), a system without mechanical linkage. SBW is seen as a long-term solution for fully autonomous vehicles but poses higher safety challenges.

Also notable is research on multi-phase motor applications, improving reliability and enabling continued torque output during partial faults.

In conclusion, EPS is no longer a standalone actuator, but a safety-critical subsystem deeply integrated with autonomy, human-machine interaction, and vehicle-level risk management. Ongoing technological and regulatory advancements continue to reshape the safety requirements and design practices for EPS.

Chapter 3. Literature Review: Field Oriented Control

This chapter establishes the mathematical framework for the motor control strategy through three key areas:

- **Mathematical Modeling:** Detailed PMSM dynamics in stationary and rotating frames using Clarke and Park transformations.
- **Dynamic Equations:** Derivation of torque production, saliency effects, and d-q axis models specifically to support subsequent sensorless estimation.
- **FOC Architecture:** Analysis of Field-Oriented Control, covering the inner current loop, outer speed loop, and PWM generation.

The mathematical formulations presented in this chapter follow established structures in the cited literature and are restated here in a unified framework to support EPS-oriented system-level comparison and architectural decision-making.

3.1 Introduction to Permanent Magnet Synchronous Machines (PMSMs) and its Mathematical Model

Permanent Magnet Synchronous Machines (PMSMs) are widely adopted in variable-speed applications due to their high efficiency and torque density. In automotive Electric Power Steering (EPS) systems, PMSMs are particularly suitable because they provide smooth torque generation at low speeds and enable precise torque control required for steering assistance. [3-1], [3-2], [3-4]

In particular, PMSMs are highly suitable for automotive Electric Power Steering (EPS) systems, where precise torque control, low acoustic noise, and high reliability are essential. Their ability to provide consistent assist torque at near-zero speeds makes them ideal for steering applications, especially under parking or idle conditions. [3-8], [3-9], [3-10]

A PMSM typically comprises a rotor equipped with embedded permanent magnets and a stator carrying three-phase windings interfaced with an inverter optimized for automotive power systems. This section develops the mathematical model of the PMSM in the context of EPS applications, followed by the introduction of Field-Oriented Control (FOC) methodology.

FOC is especially advantageous for EPS applications due to its decoupled control of torque and flux, allowing the motor to generate assist torque even when the vehicle is stationary. Additionally, its ability to deliver high bandwidth current regulation ensures responsive steering performance, while minimizing torque ripple and acoustic noise. These features, along with FOC's inherent compatibility with current feedback-based safety

diagnostics, make it an optimal solution for modern EPS systems. [3-1], [3-5], [3-8], [3-11]

The electrical dynamics of a three-phase PMSM can be captured using the following set of equations. [3-2], [3-3], [3-4] Here, u_{abc} denotes the vector of three-phase stator voltages, which depends on the stator resistance r_s , the three-phase stator currents i_{abc} , and the stator flux linkages λ_{abc} . Moreover, λ_{abc} is a combination of the inductive properties of the stator, represented by the inductance matrix L_s , and the magnetic flux produced by the permanent magnets, indicated by λ_{rabc} .

The governing equations are:

$$[u_{abc}] = r_s[i_{abc}] + \frac{d}{dt}[\lambda_{abc}] \quad (3.1)$$

$$[\lambda_{abc}] = [L_s][i_{abc}] + [\lambda_{rabc}] \quad (3.2)$$

The rotor flux linkage vector λ_{rabc} is expressed in terms of the rotor flux magnitude λ_r , the rotor mechanical angle θ_m , and the pole pair number n_p , as follows [3-2], [3-3]:

:

$$\begin{bmatrix} \lambda_{ra} \\ \lambda_{rb} \\ \lambda_{rc} \end{bmatrix} = \lambda_r \begin{bmatrix} \cos(n_p \theta_m) \\ \cos(n_p \theta_m - \frac{2\pi}{3}) \\ \cos(n_p \theta_m + \frac{2\pi}{3}) \end{bmatrix} \quad (3.3)$$

The inductance matrix L_s characterizes the mutual interactions between the stator phases and is structured as:

$$[L_s] = \begin{bmatrix} L_{aa} & M_{ab} & M_{ac} \\ M_{ba} & L_{bb} & M_{bc} \\ M_{ca} & M_{cb} & L_{cc} \end{bmatrix} \quad (3.4)$$

Additionally, L_s can be decomposed into two components: one representing the average inductance and the other reflecting the saliency effects of the motor:

$$[L_s] = [L_0] + [L_\delta] \quad (3.5)$$

$$[L_0] = \begin{bmatrix} l_0 & m_0 & m_0 \\ m_0 & l_0 & m_0 \\ m_0 & m_0 & l_0 \end{bmatrix} \quad (3.6)$$

and the saliency-dependent term is given by:

$$[L_\delta] = l_\delta \begin{bmatrix} \cos(2n_p\theta_m) & \cos(2n_p\theta_m - \frac{2\pi}{3}) & \cos(2n_p\theta_m + \frac{2\pi}{3}) \\ \cos(2n_p\theta_m - \frac{2\pi}{3}) & \cos(2n_p\theta_m + \frac{2\pi}{3}) & \cos(2n_p\theta_m) \\ \cos(2n_p\theta_m + \frac{2\pi}{3}) & \cos(2n_p\theta_m) & \cos(2n_p\theta_m - \frac{2\pi}{3}) \end{bmatrix} \quad (3.7)$$

Substituting these into the original voltage equation, the stator voltage can now be rewritten as:

$$[u_{abc}] = r_s [i_{abc}] + \frac{d}{dt} ([L_s][i_{abc}] + [\lambda_{rac}]) \quad (3.8)$$

This foundational three-phase model sets the stage for further simplification using Clarke and Park transformations, leading to a two-phase equivalent representation. [3-2], [3-3]

3.2 Clarke and Park Transformations

To transition the three-phase representation into a two-axis stationary frame, Clarke and Park transformations are applied sequentially.

First, the Clarke Transformation projects the three-phase quantities into the two-phase α - β stationary frame [3-1], [3-3], [3-16]:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (3.9)$$

Similarly, the α - β frame voltage dynamics are described by:

$$[\Lambda_{\alpha\beta}] = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} [L_s] \frac{2}{3} \begin{bmatrix} 1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \quad (3.10)$$

Expanding, $[\Lambda_{\alpha\beta}]$ has the structure:

$$[\Lambda_{\alpha\beta}] = \begin{bmatrix} L_\alpha & L_{\alpha\beta} \\ L_{\alpha\beta} & L_\beta \end{bmatrix} \quad (3.11)$$

The time derivative of the two-phase inductance matrix is given by [3-3], [3-2]:

$$\frac{d[\Lambda_{\alpha\beta}]}{dt} = l_\delta n_p \omega_m \begin{bmatrix} -\sin(2n_p\theta_m) & \cos(2n_p\theta_m) \\ \cos(2n_p\theta_m) & \sin(2n_p\theta_m) \end{bmatrix} \quad (3.12)$$

Meanwhile, the rotor flux linkage in the α - β frame can be described by:

$$\begin{bmatrix} \lambda_{r\alpha} \\ \lambda_{r\beta} \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \lambda_r \begin{bmatrix} \cos(n_p \theta_m) \\ \cos(n_p \theta_m - \frac{2\pi}{3}) \\ \cos(n_p \theta_m + \frac{2\pi}{3}) \end{bmatrix} \quad (3.13)$$

The time evolution of the rotor flux components is [3-3], [3-2]
:

$$\begin{bmatrix} \frac{d\lambda_{r\alpha}}{dt} \\ \frac{d\lambda_{r\beta}}{dt} \end{bmatrix} = \begin{bmatrix} -n_p \omega_m \lambda_{r\beta} \\ n_p \omega_m \lambda_{r\alpha} \end{bmatrix} \quad (3.14)$$

3.3 PMSM Model in the Stationary Frame

The dynamic behavior of the PMSM in the stationary two-axis reference frame (α - β) can be described by the following set of differential equations:

$$\begin{bmatrix} \frac{d[i_{\alpha\beta}]}{dt}, \frac{d[\lambda_{r\alpha\beta}]}{dt} \end{bmatrix}^T = [\Lambda_{\alpha\beta}]^{-1} \left\{ - \left(r_s + \frac{d[\Lambda_{\alpha\beta}]}{dt} \right) [i_{\alpha\beta}] + n_p \omega_m \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix} [\lambda_{r\alpha\beta}] + [u_{\alpha\beta}] \right\} \quad (3.15)$$

In this formulation:

- r_s denotes the stator resistance,
- ω_m is the rotor mechanical angular velocity,
- n_p is the number of pole pairs,
- $[\Lambda_{\alpha\beta}]$ is the two-phase stator inductance matrix,
- $[\lambda_{r\alpha\beta}]$ is the rotor flux linkage vector projected in the α - β frame,
- $[u_{\alpha\beta}]$ represents the stator voltages in the α - β frame.

If the rotor saliency effect is negligible (i.e., the inductance matrix is constant and independent of the rotor position), the above dynamic model simplifies significantly to [3-2], [3-3], [3-4]:

Where l_s is the scalar stator self-inductance:

$$[l_s] \begin{bmatrix} \frac{d[i_{\alpha\beta}]}{dt}, \frac{d[\lambda_{r\alpha\beta}]}{dt} \end{bmatrix}^T = \left\{ -r_s [i_{\alpha\beta}] + n_p \omega_m \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix} [\lambda_{r\alpha\beta}] + [u_{\alpha\beta}] \right\} \quad (3.16)$$

Furthermore, the final form of the PMSM model in the stationary α - β reference frame is presented through Equation (3.16). This model incorporates the two-phase stator inductance matrix, $\Lambda_{\alpha\beta}$.

Nevertheless, if rotor saliency effects are absent, a considerable simplification can be achieved, leading to the reduced model shown in Equation (3.16).

3.4 Transformation to the Synchronously Rotating Frame (d-q Frame)

Once the PMSM model has been formulated in the stationary frame, it is practical to convert it into a synchronously rotating frame using a Park Transformation.

The Park Transformation effectively maps the α - β frame onto the rotating direct (d) and quadrature (q) axes, which rotate at the same electrical angular velocity as the rotor.

Equation (3.17) provides the mathematical expression for transitioning from the two-phase stationary α - β frame to the two-phase rotating d-q frame:

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos \theta_e & \sin \theta_e \\ -\sin \theta_e & \cos \theta_e \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (3.17)$$

Where θ_e denotes the electrical rotor position angle. Similarly, a direct Park transformation from the three-phase (abc) system into the rotating d-q frame can be derived as shown in Equation (3.18):

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos(\theta_e) & \cos(\theta_e - \frac{2\pi}{3}) & \cos(\theta_e + \frac{2\pi}{3}) \\ -\sin(\theta_e) & -\sin(\theta_e - \frac{2\pi}{3}) & -\sin(\theta_e + \frac{2\pi}{3}) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (3.18)$$

In this rotating frame, the direct axis (d-axis) is oriented along the rotor magnetic flux, whereas the quadrature axis (q-axis) lies orthogonal to it. In these transformations:

- i_d corresponds to the component aligned with the magnetic flux axis (direct axis),
- i_q corresponds to the quadrature axis component responsible for torque generation.

Thus, the decomposition into d-q components enables a decoupled control of torque and flux, which is a fundamental principle underlying Field-Oriented Control (FOC). [3-1], [3-3], [3-15]

The relationship between the electrical angle ϕ_e and the mechanical angle θ_m is expressed by:

$$\phi_e = n_p \theta_m \quad (3.19)$$

where n_p denotes the number of pole pairs. [3-2], [3-3]

3.5 PMSM Dynamic Model in d-q Frame

Once the Park Transformation has been applied, the dynamic behavior of the PMSM in the rotating d-q reference frame can be modeled as follows [3-1], [3-2], [3-4]:

$$\begin{bmatrix} \frac{di_d}{dt}, \frac{di_q}{dt}, \frac{d\omega_r}{dt} \end{bmatrix}^T = \begin{bmatrix} -\frac{r_s}{l_s}i_d + n_p\omega_r i_q + \frac{v_d}{l_s} \\ -\frac{r_s}{l_s}i_q - n_p\omega_r i_d + \frac{v_q}{l_s} - \frac{n_p\omega_r \lambda_r}{l_s} \\ -\frac{f_r}{J}\omega_r + \frac{n_p\lambda_r}{J}i_q - \frac{1}{J}T_l \end{bmatrix} \quad (3.20)$$

Where:

- r_s is the stator resistance,
- l_s is the stator self-inductance,
- λ_r is the rotor permanent magnet flux linkage,
- f_r is the viscous friction coefficient,
- J is the moment of inertia,
- T_l is the external load torque,
- ω_r is the rotor mechanical speed.

This dynamic model forms the basis for the development of the control strategy used in EPS motor drives.

3.6 Electromagnetic Torque in d-q Frame

The electromagnetic torque T_e produced by the PMSM in the d-q frame can be formulated as:

$$T_e = \frac{3}{2}n_p\lambda_r i_q \quad (3.21)$$

Here, the torque is directly proportional to the q-axis current component i_q and the rotor flux λ_r , emphasizing the decoupled control strategy between torque and flux under Field-Oriented Control (FOC). [3-2], [3-13]

3.7 Impact of Saliency and IPMSM Modeling

To incorporate the effects of saliency into the PMSM model, it becomes necessary to differentiate between the d-axis inductance L_d and the q-axis inductance L_q . [3-2], [3-3], [3-13]

The saliency-related inductance matrices can be expressed as:

$$\begin{aligned} l_\delta &= \frac{1}{3}(L_d - L_q) \\ l_0 &= \frac{1}{3}(L_d + L_q) \end{aligned} \quad (3.22)$$

The two-phase stationary frame inductance matrix $[\Lambda_{\alpha\beta}]$ can then be represented by:

$$[\Lambda_{\alpha\beta}] = \frac{1}{2}(L_d - L_q) \begin{bmatrix} \cos(2\phi_e) & \sin(2\phi_e) \\ \sin(2\phi_e) & \cos(2\phi_e) \end{bmatrix} + \frac{1}{2}(L_d + L_q) \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad (3.23)$$

Where ϕ_e denotes the rotor electrical angle.

3.8 Dynamic Model of IPMSM in the Rotating Frame

In the rotating d-q frame, the IPMSM dynamic equations considering saliency become:

$$\begin{bmatrix} \frac{di_d}{dt}, \frac{di_q}{dt}, \frac{d\omega_r}{dt} \end{bmatrix}^T = \begin{bmatrix} -\frac{r_s}{L_d}i_d - n_p\omega_r\frac{L_q}{L_d}i_q + \frac{v_d}{L_d} \\ -\frac{r_s}{L_q}i_q + n_p\omega_r\frac{L_d}{L_q}i_d + \frac{v_q}{L_q} - n_p\omega_r\frac{\lambda_r}{L_q} \\ -\frac{f_r}{J}\omega_r + \frac{n_p}{J}((L_d - L_q)i_di_q + \lambda_r i_q) - \frac{1}{J}T_l \end{bmatrix} \quad (3.24)$$

For an IPMSM, the electromagnetic torque T_{em} includes two contributing components: one due to the permanent magnets and the other due to magnetic saliency (reluctance torque). [3-2], [3-13]

The torque expression becomes:

$$T_{em} = \frac{3}{2}n_p(L_d - L_q)i_di_q + \frac{3}{2}n_p\lambda_r i_q \quad (3.25)$$

Here, the first term represents **reluctance torque** due to saliency, and the second term is due to the permanent magnets. The first term represents the additional torque generated due to the difference between L_d and L_q , while the second term corresponds to the torque from the permanent magnet flux linkage.

In Saliency Effects in PMSMs, the physical structure of the rotor significantly influences the magnetic saliency characteristics of a PMSM. In the case of Surface-mounted PMSMs (SPMSMs), permanent magnets are attached to the outer surface of the rotor, resulting in an air gap that is spatially uniform. Consequently, the stator inductances along the direct (d) and quadrature (q) axes remain equivalent, rendering the motor effectively non-salient. [3-2], [3-14]

By contrast, Interior PMSMs (IPMSMs) feature permanent magnets embedded within the rotor body. This construction leads to a non-uniform magnetic circuit, causing the d-axis inductance to differ from the q-axis inductance. Such magnetic anisotropy enables the exploitation of reluctance torque and enhances torque

controllability. These characteristics are beneficial for automotive electric power steering (EPS) systems, where accurate torque generation, rapid dynamic response, and efficient operation are required.

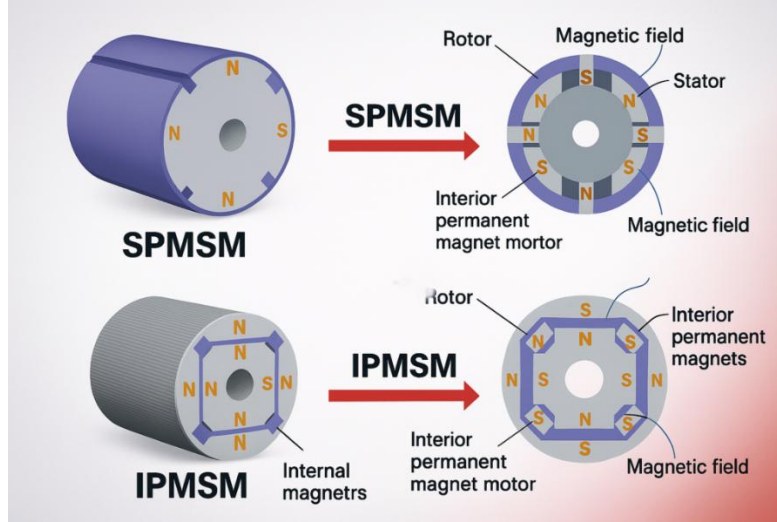


Figure 3.1. The structural differences between SPMSMs and IPMSMs [3-18]

3.9 Motor Control Strategy for IPMSM Drives

With the established dynamic model of the IPMSM, a Field-Oriented Control (FOC) approach is developed to independently regulate torque and flux.

Depending on the system requirements, a **speed reference** may or may not be necessary:

- In many electric drive systems, including traction motors, torque commands are often used as the primary control input. Similarly, in EPS systems, the assist torque is determined based on steering torque demand from the driver. [3-5], [3-1]
- A **speed control** loop is added only when cruise control or regulated-speed modes are active.

The mechanical rotational speed n of the IPMSM relates to the stator electrical frequency f_e

By [3-3], [3-2]:

$$n = \frac{60f_e}{n_p} \quad (3.26)$$

Where:

- n_p is the number of pole pairs,
- n is expressed in revolutions per minute (RPM).

3.10 Torque Regulation Procedure in FOC

The regulation of motor torque under FOC follows these steps:

1. **Measurement of Three-Phase Currents:**

The stator currents i_a, i_b, i_c are measured by external current sensors.

2. **Transformation to d-q Frame and Error Computation:**

The measured currents are transformed into the d-q rotating frame.

Error signals are generated by comparing the measured i_d, i_q with their respective reference values.

3. **Control Voltage Generation and Modulation:**

The error signals feed into PI controllers for the d-axis and q-axis currents.

The PI controllers output reference voltages v_d, v_q , which are modulated by Space Vector PWM (SVPWM) to drive the inverter. [3-6], [3-7]

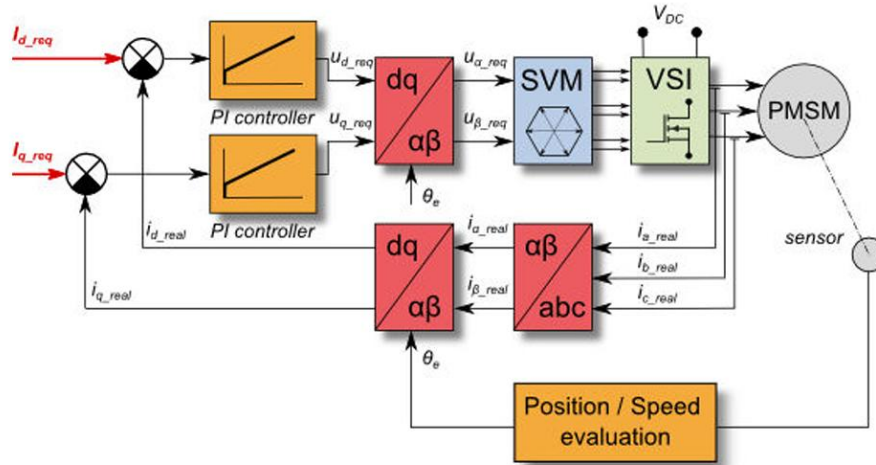


Figure 3.2. FOC structure for an IPMSM. [3-19]

3.11 Field-Oriented Control System Architecture

In EPS applications, the fast response of the inner current loop is particularly important because the steering assist torque must respond rapidly to driver input. The components and functionalities of the FOC system include:

- **Current Regulation Inner Loop:**

Separate PI controllers for i_d , and i_q provide fast torque control response. [3-5], [3-1]

- **Optional Speed Regulation Outer Loop:**
If speed control is required, an external speed PI controller adjusts the i_q current reference dynamically.
- **Reference Frame Transformations:**
Park and Clarke transformations manage the conversion between the stationary and rotating frames.
- **Space Vector PWM (SVPWM):**
Generates optimized inverter switching patterns to achieve the commanded voltages.
- **Protection and Monitoring Functions:**
Ensure safe and reliable motor operation under all conditions.

3.12 Torque Loop (Inner Control Loop)

The **Torque Loop**, forming the innermost control layer in FOC, consists of:

- Sampling of stator phase currents,
- Clarke and Park transformations to extract i_d , and i_q .
- Current error generation,
- Current PI control for each axis,
- Voltage reference generation for SVPWM.

3.13 Speed Control (Outer Loop)

If implemented, the speed PI controller adjusts the i_q current reference to ensure the motor speed n follows the commanded speed n_{ref} . Changes in the commanded speed are translated into changes in i_q thereby modulating the generated torque and ultimately the rotor speed.

Thus, the **outer speed control loop** governs the dynamic behavior over longer time constants, while the **inner current/torque loop** handles faster torque dynamics.

Although speed control loops are commonly used in general motor drives, EPS systems primarily rely on torque control. A speed loop may still be implemented for diagnostic or operational purposes, but the dominant control objective remains torque regulation.

3.14 PWM Signal Generation

Finally, the d-q voltage references generated by the current controllers are transformed back into the three-phase system using inverse Park transformation. These voltage references are then modulated through **Space Vector PWM (SVPWM)** to produce appropriate gate signals for the inverter's semiconductor switches,

ensuring precise application of the desired stator voltages. The use of SVPWM also helps reduce current ripple and acoustic noise, which is important for maintaining quiet operation in EPS systems.

Chapter 4. Literature Review: Overview of 4 control strategies for sensorless control of PMSM

This chapter outlines the fundamental principles of the four major sensorless control methods for PMSMs. The mathematical formulations presented in this chapter follow established structures in the cited literature and are restated here in a unified framework to support EPS-oriented system-level comparison and architectural decision-making.

In the following chapter5, the advantages and disadvantages of each method will be compared in detail, leading to a conclusion on the most suitable approach for sensorless EPS systems.

4.1 Extended Kalman Filter (EKF)

4.1.1. Mathematical Model of Permanent Magnet Synchronous Machines

Permanent Magnet Synchronous Machines (PMSMs) have emerged as a popular choice in various applications, and are increasingly adopted in automotive systems such as Electric Power Steering (EPS), owing to their superior efficiency, elevated power density, and rapid dynamic response.

However, the inherent nonlinear, multivariable, and tightly coupled nature of PMSMs presents significant challenges for implementing control algorithms. Accurately modeling the motor's behavior is fundamental for developing effective control strategies. Consequently, a detailed and thorough examination of a PMSM's mathematical model becomes essential prior to crafting its control system. The electrical voltage relationships of a PMSM in the stationary α - β coordinate frame are given as [4-1], [4-7]:

$$\begin{cases} v_a = R_s i_a + L_s \frac{di_a}{dt} - \omega_e \lambda_m \sin \theta_e \\ v_b = R_s i_b + L_s \frac{di_b}{dt} - \omega_e \lambda_m \cos \theta_e \end{cases} \quad (4.1.1)$$

By rearranging the above expressions to formulate the current dynamics, we obtain:

$$\begin{cases} \frac{di_a}{dt} = -\frac{R_s}{L_s} i_a + \omega_e \frac{\lambda_m}{L_s} \sin \theta_e + \frac{v_a}{L_s} \\ \frac{di_b}{dt} = -\frac{R_s}{L_s} i_b - \omega_e \frac{\lambda_m}{L_s} \cos \theta_e + \frac{v_b}{L_s} \end{cases} \quad (4.1.2)$$

Here, i_a and v_a represent the phase current and voltage along the α -axis, whereas i_b and v_b denote those along the β -axis. R_s stands for the stator resistance, and L_s symbolizes the stator inductance. The rotor's electrical angular speed is ω_e , its angular position is θ_e , and λ_m denotes the flux linkage from the permanent magnet. Both voltages v_a, v_b and currents i_a, i_b can be measured directly. Acknowledging that the mechanical time constant significantly exceeds the electrical time constant for a PMSM, it is reasonable to

assume that ω_e remains effectively constant within each control sampling period. Therefore, the following simplified dynamic relationships hold [4-1], [4-7]:

$$\begin{cases} \frac{d\omega_e}{dt} = 0 \\ \frac{d\theta_e}{dt} = \omega_e \end{cases} \quad (4.1.3)$$

$$\dot{z} = \frac{dz}{dt} = g(z) + Dv \quad (4.1.4)$$

$$h = Cz \quad (4.1.5)$$

where: [4-1], [4-7]

$$z = \begin{bmatrix} i_a \\ i_b \\ \omega_e \\ \theta_e \end{bmatrix}, \quad v = \begin{bmatrix} v_a \\ v_b \end{bmatrix}, \quad h = \begin{bmatrix} i_a \\ i_b \end{bmatrix} \quad (4.1.6)$$

The system functions are given as:

$$g(z) = \begin{bmatrix} -\frac{R_s}{L_s} i_a + \omega_e \frac{\lambda_m}{L_s} \sin \theta_e \\ -\frac{R_s}{L_s} i_b - \omega_e \frac{\lambda_m}{L_s} \cos \theta_e \\ 0 \\ \omega_e \end{bmatrix} \quad (4.1.7)$$

$$D = \begin{bmatrix} \frac{1}{L_s} & 0 \\ 0 & \frac{1}{L_s} \\ 0 & 0 \\ 0 & 0 \end{bmatrix}, \quad C = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \quad (4.1.8)$$

Since both (4.1.4) and (4.1.5) are nonlinear, they must be linearized for the purpose of digital implementation.

Thus, the discrete-time form of the model becomes:

$$z(k+1) = g[z(k)] + D(k)v(k) + \Xi(k) \quad (4.1.9)$$

$$h(k) = C(k)z(k) + \Upsilon(k) \quad (4.1.10)$$

Here, $\Xi(k)$ represents the process noise, and $\Upsilon(k)$ signifies the measurement noise. Assuming that both $\Xi(k)$ and $\Upsilon(k)$ are zero-mean Gaussian white noise processes, we have [4-6], [4-2]:

$$\mathbb{E}\{\Xi(k)\} = 0, \quad \mathbb{E}\{\Upsilon(k)\} = 0 \quad (4.1.11)$$

where $\mathbb{E}\{\cdot\}$ denotes the mathematical expectation operator.

Within the extended Kalman filtering (EKF) framework, neither $\Xi(k)$ nor $Y(k)$ is utilized directly; instead, their covariance matrices Q and R are employed, which will be defined explicitly later. [4-6], [4-2]

$$E\{\omega\omega^T\} = Q, \quad E\{vv^T\} = R \quad (4.1.12)$$

In the relations above, it is assumed that the process noise $\omega(k)$ and the measurement noise $v(k)$

are mutually independent, and that the initial state vector $z(0)$ is a random vector independent of both $\omega(k)$ and $v(k)$.

4.1.2 Realization of the Extended Kalman Filter

The procedure of state estimation using the EKF consists of two sequential phases: the prediction stage and the correction stage. [4-6]

4.1.2.1 Prediction of the State Vector:

Based on the previous state estimate $\hat{z}(k)$ and the input $\mu(k)$, the prediction of the state at the next time step $(k + 1)$ is conducted according to [4-6], [4-2]:

$$\tilde{z}(k + 1) = \hat{z}(k) + \Delta t [g(\hat{z}(k)) + D(k)\mu(k)] \quad (4.1.13)$$

Here, (4.1.13) constitutes the Forward Euler discretization and Δt is the sampling time interval, $\hat{z}$ stands for the estimated state, and $\tilde{z}$ indicates the predicted state.

4.1.2.2 Prediction of the output:

The corresponding predicted output $\tilde{h}(k + 1)$ is determined by [4-6], [4-2]:

$$\tilde{h}(k + 1) = C\tilde{z}(k + 1) \quad (4.1.14)$$

At this point, the EKF calculates the anticipated measurement $\tilde{h}(k + 1)$ at step $k + 1$, which is utilized for subsequent update and refinement of the state estimate.

4.1.2.3 Prediction of the Error Covariance:

The error covariance propagation is given by [4-6], [4-2]:

$$\tilde{P}(k + 1) = \hat{P}(k) + \Delta t [M(k)\hat{P}(k) + \hat{P}(k)M^T(k)] + Q \quad (4.1.15)$$

This is the discretized form of the continuous-time Riccati equation. The process noise covariance Q encapsulates model uncertainties and disturbances, typically determined from empirical modeling knowledge or real-world experience.

The Jacobian matrix $M(k)$, representing the system sensitivity, is computed by [4-6], [4-2]:

$$M(k) = \left. \frac{\partial(g)}{\partial z} \right|_{z=\hat{z}(k)} \quad (4.1.16)$$

Explicitly, $M(k)$ is structured as [4-6], [4-2]:

$$M(k) = \begin{bmatrix} -\frac{r}{L_m} & 0 & (\phi_m/L_m) \sin \hat{\phi}_r(k) & \omega_r(k)(\phi_m/L_m) \cos \hat{\phi}_r(k) \\ 0 & -\frac{r}{L_m} & -(\phi_m/L_m) \cos \hat{\phi}_r(k) & \omega_r(k)(\phi_m/L_m) \sin \hat{\phi}_r(k) \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix} \quad (4.1.17)$$

4.1.2.4 Calculation of the Kalman Gain:

The Kalman gain $G(k + 1)$ is evaluated by [4-6], [4-2]:

$$G(k + 1) = \tilde{P}(k + 1)C^T [C\tilde{P}(k + 1)C^T + R]^{-1} \quad (4.1.18)$$

This gain matrix is employed to adjust the prediction by minimizing the deviation between predicted and observed outputs.

4.1.2.5 State Update through Feedback:

The refined state estimate at $(k + 1)$ is updated via [4-6], [4-2]:

$$\hat{z}(k + 1) = \tilde{z}(k + 1) + G(k + 1) [h(k + 1) - \tilde{h}(k + 1)] \quad (4.1.19)$$

This phase corresponds to the corrected prediction, aiming to adjust the estimated state values so that they align more closely with the actual measured outputs. During this phase, the previously forecasted states are evaluated against the real-time observations, and these observations are utilized to update and refine the state estimations for improved accuracy. In the corresponding formula, $h(k + 1)$ denotes the observed output applied to amend the state estimate at time instant $(k+1)$.

Typically, this observed value is acquired via measurement equipment, and in this context, following Equation (4.1.5), $h(k + 1)$ specifically refers to the actual current value at step $(k+1)$ as captured by the hardware sampling interface and processed through the data handling program.

4.1.2.6 Update of the Error Covariance Matrix:

The corrected error covariance for the next iteration is obtained as [4-6], [4-2]:

$$\hat{P}(k+1) = \tilde{P}(k+1) - G(k+1)C\tilde{P}(k+1) \quad (4.1.20)$$

The above methodology is firmly rooted in the concept of the extended Kalman filter, where feedback-driven adjustments ensure improved estimation accuracy, uncertainty suppression, and optimized system control.

According to Figure 4.1.1 EKF Block diagram, the EKF achieves this by computing the difference between the observer's predicted output and the actual system output, applying the Kalman gain G to correct the observer's state, and dynamically tuning G based on real-time covariance data. [4-4]

In the context of motor control discussed herein, the EKF is deployed to monitor the rotor speed and rotor angle of the PMSM. This is achieved by gathering signals such as terminal voltages and currents, constructing an appropriate motor model, and employing the EKF to perform online state estimation, thus yielding accurate speed and position data for the motor. The procedural flow of the EKF system designed for the PMSM application is presented in Figure 4.1.2. EKF algorithm for Permanent Magnet Synchronous Motor

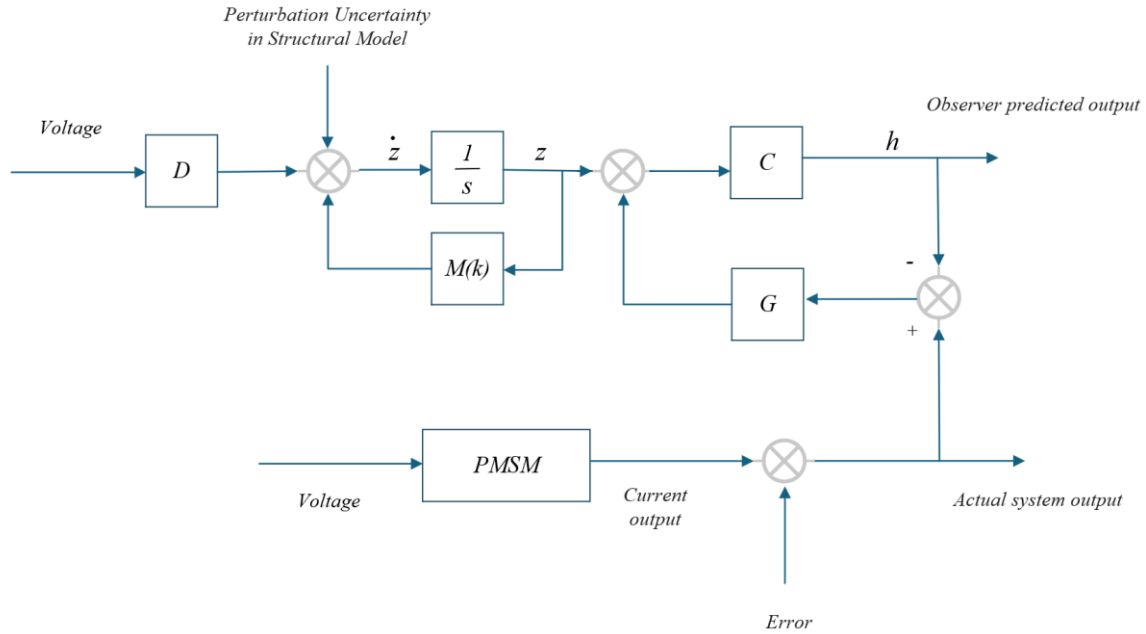


Figure 4.1.1 EKF Block diagram

Adapted from [4-1]

As illustrated in Figure 4.1.2 EKF algorithm for Permanent Magnet Synchronous Motor, during the application of the EKF observer, the variables supplied to the filter include the three-phase voltage signals v_a, v_b, v_c , and the three-phase current signals i_a, i_b, i_c derived from the PMSM. These signals are initially

sampled via dedicated hardware circuits and undergo analog-to-digital (AD) conversion before being passed to the EKF.

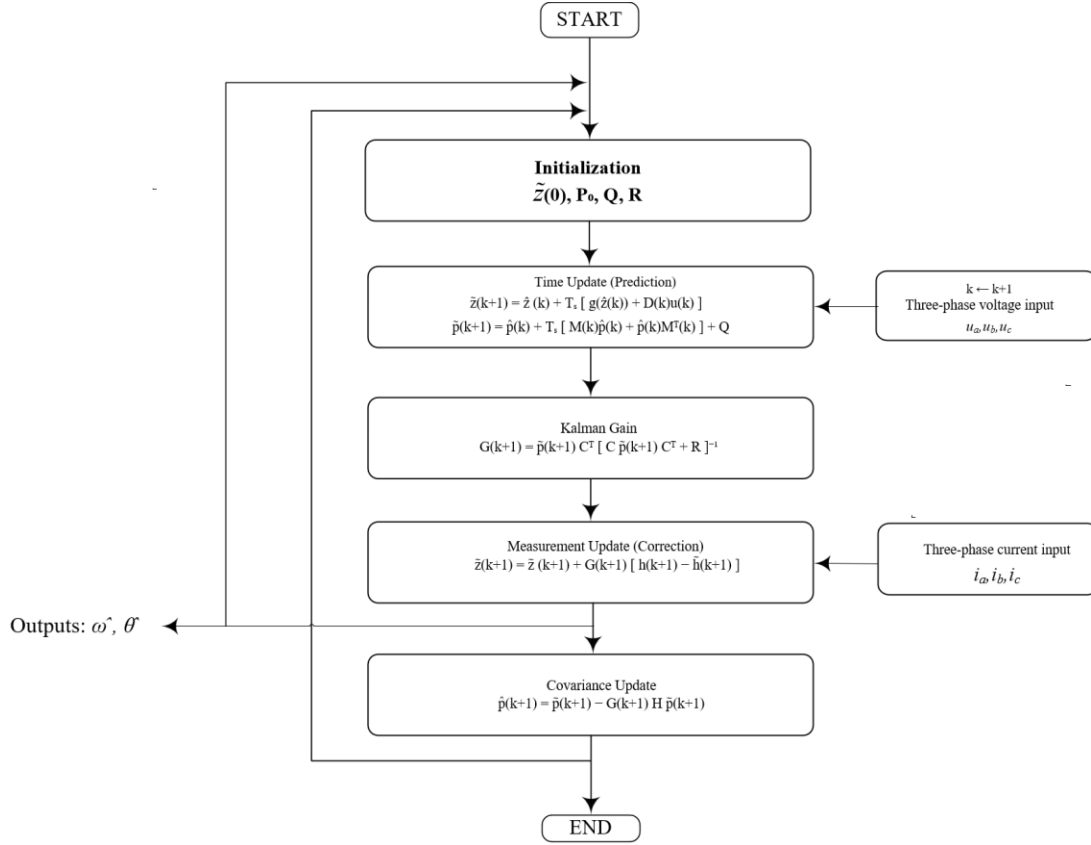


Figure 4.1.2 EKF algorithm for Permanent Magnet Synchronous Motor [4-71], [4-72], [4-73]

The resulting outputs from the EKF are the estimated motor angular velocity $\hat{\omega}$ and rotor angular position $\hat{\theta}$. Since the rotor position is thus inferred through the EKF process, there is no longer a necessity for a separate physical position sensor, thereby enabling sensorless PMSM operation. [4-71], [4-72], [4-73]

4.1.3 Sensorless Vector Control Scheme Leveraging the Extended Kalman Filter

The control approach adopts a zero direct-axis current strategy, $i_{d_ref} = 0$, in setting up a sensorless vector-controlled PMSM drive based on the EKF framework, with the associated control architecture shown in Figure 4.1.3 EKF Block diagram for PMSM sensorless vector control. [4-1], [4-7]

Initially, the machine's three-phase current signals i_a, i_b, i_c are acquired. Following Clarke and Park coordinate transformations, the control variables i_d, i_q are extracted. These feedback currents are then compared to their respective reference signals i_{d_ref} and i_{q_ref} .

Control voltages v_d, v_q within the d-q frame are computed by means of PI regulators. Subsequently, an inverse Park (Reverse Park) transformation converts v_d, v_q into voltages v_α and v_β within the stationary frame.

Next, a Space Vector Pulse Width Modulation (SVPWM) unit processes v_α and v_β to derive the appropriate switching signals for the six power semiconductor devices (MOSFETs), achieving precise voltage application to the motor terminals. In parallel, the EKF observer calculates the rotor's rotational speed and positional information during motor operation. [4-9], [4-8]

Specifically, the EKF input set comprises two-phase stator currents i_α and i_β and stator voltages v_α and v_β , while its output includes the rotor angle estimate $\hat{\theta}$ and the rotor speed estimate $\hat{\omega}$.

The described method establishes a fully closed-loop feedback system, contributing to enhanced system stability.

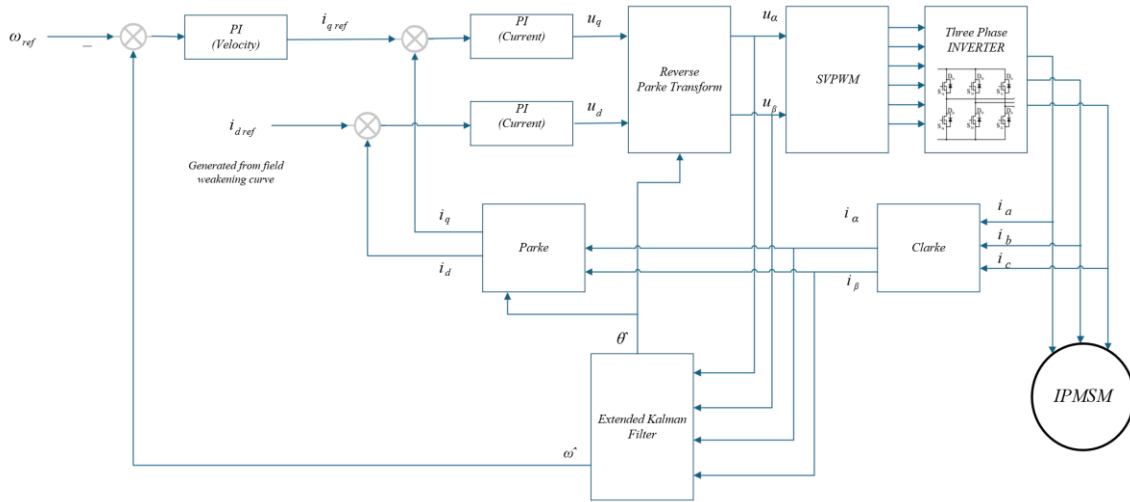


Figure 4.1.3 EKF Block diagram for PMSM sensorless vector control

Adapted from [4-1], [4-8]

Regarding the software system design, the proposed PMSM sensorless vector control framework divides responsibilities into two primary modules: the **command unit** and the **estimation unit**.

- The **command unit** is responsible for transmitting actuation commands to the inverter.
- The **estimation unit** monitors motor behavior, determining the rotor's position and feeding this information back to the command unit.

Referring again to Figure 4.1.3 EKF Block diagram for PMSM sensorless vector control, it is evident that the execution speed of the estimation process must surpass that of the controller to guarantee high-quality and timely corrections.

Employing a modular architecture — separating estimation and control — ensures minimal interdependency, enhances modular reusability, and streamlines maintenance and scalability efforts for the entire control system. [4-1]

4.2 SMO (Sliding Mode Observer)

4.2.1 Introduction

This section introduces a method for sensorless control of permanent magnet synchronous machines (PMSMs), emphasizing magnetic saliency detection. The approach employs high-frequency injection (HFI) within a modified PMSM model situated in the $\gamma\delta$ reference coordinate system. Beyond the sensorless strategy, the focus is on estimating magnetic saliency to offer a practical way to monitor PMSM inductance changes. Saliency evaluation is incorporated into the position and speed algorithms by leveraging current data. An important aspect of the revised PMSM model is its integration of both the saliency-induced and permanent magnet rotor flux. To implement HFI for sensorless operation, the flux model's structure is tailored appropriately. A new inductance matrix, formulated to align with the HFI approach, includes elements corresponding to average inductance and angular error variations. Additionally, a sliding mode observer (SMO) is implemented to compute rotor flux speed and angle using equivalent control, where a smoothed angular error function replaces the sign function to mitigate chattering. The effectiveness of the control relies on the suitability of both the modified model and the SMO architecture to precisely track rotor flux with sufficient robustness and accuracy. Areas of studies are permanent magnet synchronous machine (PMSM), sensorless control, sliding mode observer (SMO), magnetic saliency, high-frequency injection (HFI), rotor flux modification. [4-11], [4-15], [4-16].

Symbols:

- u_d, u_q : stator voltages in the dq reference frame
- i_d, i_q : stator currents in the dq frame
- λ_d, λ_q : dq-axis magnetic flux linkages
- λ_m : magnetic flux of the rotor
- L_d, L_q : inductances along the dq axes
- $\Sigma L = \frac{L_d + L_q}{2}$: mean inductance

- $\Delta L = \frac{L_d - L_q}{2}$: difference in inductance
- r_s : resistance of the stator
- u_γ, u_δ , voltages in $\gamma\delta$ coordinates
- i_γ, i_δ , currents in $\gamma\delta$ frame
- $\lambda_\gamma, \lambda_\delta$, magnetic fluxes in $\gamma\delta$ frame
- p : pair count of magnetic poles
- $\theta = \theta_e$, Electrical position angle
- $\omega = \omega_e$, angular velocity in electrical units

The application of permanent magnet synchronous machines (PMSMs) continues to rise across various industrial fields, driven by their advantageous traits. These include exceptional efficiency and dynamic behavior, high torque-to-inertia ratio, compact design, and extremely low torque ripple. Internal PMSMs (IPMSMs), specifically, allow for sensorless operation at low or even zero speed, and field weakening extends their operating range beyond nominal speeds. Operating these machines efficiently minimizes energy consumption, offsetting their higher upfront cost. To achieve optimal control of PMSMs, modern techniques such as field-oriented control (FOC) are widely adopted, offering full torque regulation and rapid dynamic response across all speeds. Yet, implementing such control requires accurate rotor position detection. In practical EPS systems, this is typically achieved using automotive-grade magnetic position sensors (e.g., resolvers or Hall-based sensors) that directly measure the rotor angle. However, these components add cost and introduce noise, particularly in environments with electromagnetic interference. Thus, interest has shifted toward sensorless techniques that replace hardware sensors with estimation-based methods. [4-19], [4-17], [4-21]

Accurately crafted observers, such as those estimating back electromotive force (back-EMF or BEMF) or rotor flux, are widely used to derive the rotor's angular speed and position. These back-EMF-based techniques are particularly adept at offering precise position estimation within a moderate to high-speed range. At lower speeds, however, the back-EMF becomes significantly weaker, causing a major decline in estimation precision, as its amplitude approaches the level of noise introduced by measurements. To address this limitation, an alternative group of methods employs spatial saliency and signal injection to identify rotor flux direction based on variations in inductance across the PMSM geometry. By injecting high-frequency voltage signals, as illustrated in Figure 4.2.1, rotor position can be estimated even at standstill. These sensorless strategies rely on the inherent magnetic anisotropy

or the existence of magnetic saliency, denoted by $(L_d - L_q) \neq 0$. [4-14], [4-16], [4-24]

Nonetheless, when observers operate in the $\alpha\beta$ stationary reference frame, they frequently exhibit phase lag between actual and estimated EMF, as the true extended EMF is inherently sinusoidal. This phase shift arises from the nature of the observer, which acts similarly to a filtering process. Despite $\alpha\beta$ -based modeling, PMSM variables like flux and extended EMF remain time-invariant in the synchronous dq frame—though this frame itself cannot be directly employed for estimation. A viable alternative is to adopt the $\gamma\delta$ reference frame, rotating at an estimated angular speed $\hat{\omega}$ and lagging behind the dq frame by an electrical angle $\bar{\theta}$. This angle represents the difference between dq and $\gamma\delta$ axes (see Figure 4.2.2). A principal advantage of switching to the $\gamma\delta$ frame lies in its applicability to sensorless estimation of dynamic variables in a rotating context, bypassing the limitations of stationary frames. The $\gamma\delta$ -transformed PMSM model offers particular benefits for handling magnetic saliency, as it is compatible with both salient-pole machines ($L_d \neq L_q$) and non-salient ones where $L_d = L_q = L_s$. Observers constructed within the $\gamma\delta$ frame estimate the deviation angle $\bar{\theta}$ between the dq and $\gamma\delta$ frames, instead of needing the rotor angle θ directly. This design resolves the phase delay issue since the $\gamma\delta$ frame aligns with rotor flux dynamics, ensuring minimal discrepancy between dq- and $\gamma\delta$ -based variables. [4-11], [4-19], [4-23]

4.2.1.2 PMSM Types and Saliency

Generally, PMSMs are categorized into two standard forms of permanent magnet machines: surface-mounted PMSMs (SPMSM or SPM) and internal permanent magnet variants (IPMSM or IPM). The term magnetic saliency refers to the inductance difference between d- and q-axis components, that is, $(L_d - L_q)$. In SPMSM configurations, magnets are installed on the rotor's exterior surface, while in IPMSMs, they are embedded within the rotor body. Since the magnetic permeability of the permanent magnets is quite low, it's typically approximated as being equal to that of air. Consequently, for SPMSMs, the effective air gap along both flux paths (d and q axes) remains consistent. This means the measured inductance stays constant regardless of rotor position, leading to $L_d = L_q$ and near-zero magnetic saliency, i.e., $(L_d - L_q) \approx 0$. Conversely, in IPMSMs, the effective air gap varies with rotor orientation because the magnetic permeability of the iron structure is higher than that of the magnets. This variation causes fluctuating inductance values, resulting in non-zero magnetic saliency, i.e., $(L_d - L_q) \neq 0$. [4-20], [4-16], [4-11]

4.2.1.3 Magnetic Saliency Characterization and Inductance Breakdown

Because internal magnets exhibit significantly lower magnetic permeability compared to iron, the overall machine inductance undergoes variations, resulting in non-zero magnetic saliency—namely, $(L_d - L_q) \neq 0$. These fluctuations can be exploited to deduce the rotor's angular position. Thus, position estimation through inductance saliency measurements becomes feasible, even enabling sensorless or open-loop control. However, accurately quantifying magnetic saliency in PMSMs poses substantial challenges, necessitating rigorous analysis via experimental testing or finite element methods (FEM). Variations in inductance are largely

attributed to two main causes: magnetic saturation and geometrical saliency of the rotor—both of which are considered primary sources. Additionally, eccentricities in the rotor, the generation of eddy currents, and the slotting effects present in the stator and rotor structure can contribute to saliency phenomena. [4-16], [4-18], [4-11] Within the dq coordinate system, synchronous inductances are typically modeled as a summation of leakage and magnetizing components: specifically, $L_d = L_{ld} + L_{md}$, and $L_q = L_{lq} + L_{mq}$,

where L_{ld} , L_{lq} represent leakage inductances, and L_{md} , L_{mq} account for magnetization-related contributions. Leakage terms are associated with magnetic field losses through slots, teeth, or core faces, while magnetizing inductances reflect flux through the air gap. Since both saturation and rotor geometry affect these inductive elements, they play a role in shaping the overall magnetic profile. Signal injection techniques are capable of detecting both types of saliencies by applying excitation within a frequency band, usually ranging from 0.5 kHz to 2 kHz. Nevertheless, injecting high-frequency signals might reduce estimation precision due to bandwidth constraints of the current regulator in closed-loop settings. Hence, introducing such harmonic voltages directly within the inverter stage is often preferred.

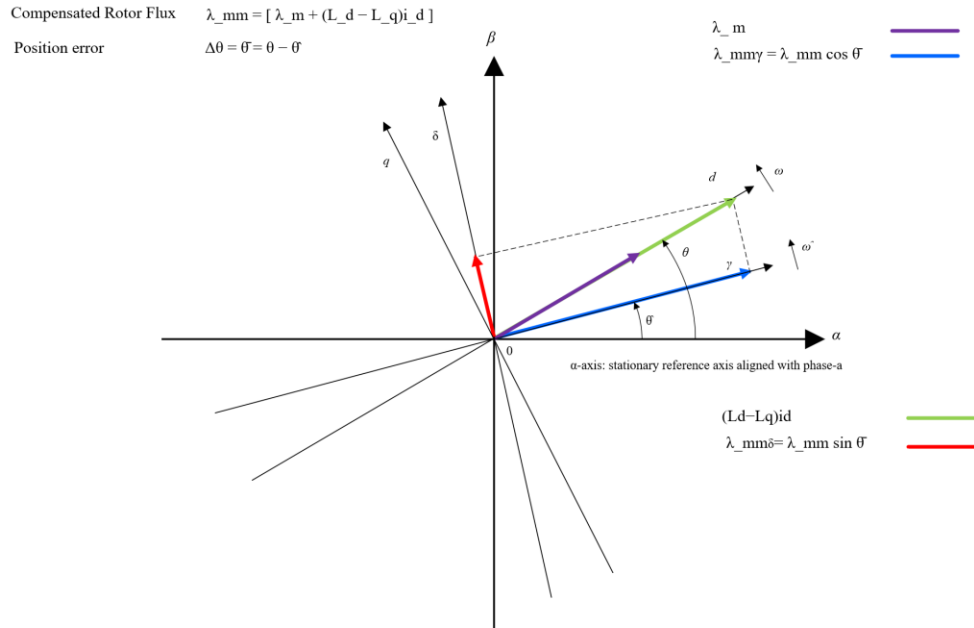


Figure 4.2.2 Phasor diagram illustrating the modified rotor flux λ_{mm} within both the dq and $\gamma\delta$ rotating reference frames, shown in relation to the $\alpha\beta$ stationary frame, where the α -axis aligns with the magnetic axis of phase a.

4.2.1.4 Proposal of New $\gamma\delta$ -Frame PMSM Model

This section introduces a revised PMSM mathematical model designed to express rotor flux equations within the $\gamma\delta$ rotating reference frame, which is aligned with an estimated angular velocity. The $\gamma\delta$ -based model proves especially suitable for sensorless control, as it integrates magnetic saliency and supports signal injection techniques. The model's development includes reformulating the rotor flux and stator flux relationships to derive a novel inductance matrix, $L_{\gamma\delta}$, which serves as a key theoretical contribution. Saliency-dependent expressions such as $(L_d - L_q)i_d$ are embedded into the $\gamma\delta$ flux formulation, enhancing model precision by reducing approximation errors. In contrast to methods that use extended EMF or saliency-augmented back-EMF, this $\gamma\delta$ formulation has the benefit that its rotor flux is unaffected by operating speed. Analytical derivations confirm that the $\gamma\delta$ inductance matrix $L_{\gamma\delta}$ is obtained from the standard dq inductance L_{dq} , transformed according to the angular difference between the dq and $\gamma\delta$ coordinate systems. Moreover, each element of $L_{\gamma\delta}$ is a function of the system's average inductance, differential inductance, and the angular deviation—demonstrating the critical importance of this transformation for ensuring accurate control dynamics. [4-11], [4-12], [4-19]

4.2.1.5 Observer Architecture and Control Structure

Introducing high-frequency voltage excitation induces HF stator currents composed of both positively and negatively rotating components, essentially generating two counter-rotating current vectors. By applying appropriate signal processing techniques—including filtering—essential information regarding the rotor's position is extracted using a sliding mode observer (SMO), which operates under an equivalent control framework. To mitigate the well-known chattering issue in SMO applications, the sign function $\text{sgn}(\cdot)$ is replaced with a smoother alternative, specifically the hyperbolic tangent function $\tanh(\cdot)$. This replacement substantially reduces chattering while preserving observer responsiveness. Another notable merit of the method lies in its ability to determine the dq inductances and magnetic saliency term $(L_d - L_q) = 2\Delta L$ by analyzing the amplitudes of the high-frequency current components. As depicted in Figure (4.2.1), the full control system constructed around the $\gamma\delta$ -modified PMSM model aims to estimate rotor speed, angle, and inductance-based saliency values. The sensorless control scheme comprises three phases: measurement, estimation, and actuation. During the measurement stage, only two stator currents $-i_a$ and i_b are required. The estimation of rotor angle and velocity is carried out by the observer, which processes the measured currents through steps like modulation, low-pass filtering (LPF), and band-pass filtering (BPF). Control is applied using three PI controllers: two managing the current loop and one handling the speed loop. The injection stage involves superimposing HF voltage onto the $\gamma\delta$ voltage commands generated by the current controllers. Once reference voltages u_a^* and u_b^* are generated, a voltage source inverter (VSI) drives the PMSM using space vector PWM (SVPWM) modulation. This improved algorithm enables accurate position/speed estimation and

real-time tracking of saliency and impedance, supporting diagnostic and fault detection applications. [4-11], [4-16], [4-25]

4.2.2 Investigation of the Revised Model for Permanent Magnet Synchronous Machine (PMSM)

4.2.2.1 Reformulated PMSM Voltage and Flux/Current Equations within the dq Rotating Reference Frame

The behavior of the PMSM mathematical representation is chiefly influenced by three aspects: whether the machine is an IPMSM or SPMSM, its inherent structural geometry, and the selected coordinate frame. For the following analysis, we focus on a salient-pole type permanent magnet synchronous machine, characterized by spatially varying air gap reluctance resulting from magnetic saliency, denoted by $L_d \neq L_q$. Moreover, the resulting back electromotive force (BEMF) waveform follows a sinusoidal profile. The voltage and flux relationships for this machine, expressed within the synchronous dq reference frame, are captured by the equations presented below [4-20], [4-21], [4-19]:

$$u_{dq} = r_s i_{dq} + \omega J_s \lambda_{dq} + \dot{\lambda}_{dq} \quad (4.2.1)$$

$$\lambda_{dq} = L_{dq} i_{dq} + \lambda_{mdq} \quad (4.2.2)$$

Where:

$$i_{dq} = \begin{bmatrix} i_d \\ i_q \end{bmatrix}, \quad \lambda_{dq} = \begin{bmatrix} \lambda_d \\ \lambda_q \end{bmatrix}, \quad J_s = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}, \quad L_{dq} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix}, \quad \lambda_{mdq} = \begin{bmatrix} \lambda_m \\ 0 \end{bmatrix} \quad (4.2.3)$$

Accounting for the saliency-induced asymmetry ($L_d - L_q$), the PMSM's flux relationships in the dq frame can be reformulated into a balanced structure. Specifically, the stator flux vector expressed in the dq reference frame can be symmetrically restated using the matrix format λ_{dq} shown in Equation (4.2.2), which provides an equivalent and mathematically consistent representation.

$$\lambda_{dq} = \begin{bmatrix} L_q & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \lambda_m + (L_d - L_q)i_d \\ 0 \end{bmatrix} = \mathbf{L}_{qq} \mathbf{i}_{dq} + \lambda_{mmdq} \quad (4.2.4)$$

In this formulation, λ_{mm} is defined as the modified rotor flux of the PMSM in the dq frame, which is influenced by the permanent magnet flux λ_m , the saliency component ($L_d - L_q$), and the stator current component along the d-axis i_d (refer to Figure 4.2.2). In practical applications, when the current i_d is small, the modified rotor flux λ_{mm} is predominantly governed by the permanent magnet flux λ_m . [4-11], [4-12], [4-13]

4.2.2.2 Reformulated Voltage and Flux/Current Expressions of PMSM in the $\gamma\delta$ Rotating Reference Frame

The vector form of the modified rotor flux is depicted schematically in Figure 4.2.2 across the $\alpha\beta$, dq , and $\gamma\delta$ coordinate systems. By definition, the $\gamma\delta$ reference frame represents a generalized, non-fixed coordinate system that rotates at an estimated angular speed $\hat{\omega}$, trailing behind the synchronous dq frame by an electrical angle $\Delta\theta$, which is calculated by $\Delta\theta = \bar{\theta} = \theta - \hat{\theta}$. The PMSM equations within the $\gamma\delta$ frame are derived by applying a rotational transformation to the original dq -based equations, utilizing the transformation matrix $K_{\Delta\theta}$ as defined in Equation (4.2.5), which is a function of the angular displacement between the dq and $\gamma\delta$ rotating coordinate systems.

4.4.2.3 Transformed PMSM Voltage and Flux/Current Relationships in $\gamma\delta$ Frame

As illustrated in Figure 4.2.2, the modified rotor flux vector is visualized within the $\alpha\beta$, dq , and $\gamma\delta$ reference systems. The $\gamma\delta$ reference frame is, by nature, an abstract rotational frame operating at an estimated angular velocity $\hat{\omega}$, and it lags behind the dq frame by an electrical angular separation $\Delta\theta$, formally defined as $\Delta\theta = \bar{\theta} = \theta - \hat{\theta}$. The PMSM model equations are obtained by applying a rotational transformation from the dq frame using the matrix $K_{\Delta\theta}$, which is defined in Equation (4.2.5) and captures the relative angular offset between the dq and $\gamma\delta$ coordinate systems. By pre-multiplying both sides of Equations (4.2.1) and (4.2.4) with the transformation matrix $K_{\Delta\theta}$, the stator voltage and flux relations in the $\gamma\delta$ coordinate system for the PMSM are formulated as follows [4-11], [4-19], [4-22]:

$$\begin{aligned} K_{\Delta\theta}\mathbf{u}_{dq} &= K_{\Delta\theta}r_s\mathbf{i}_{dq} + K_{\Delta\theta}\omega\mathbf{J}_s\lambda_{dq} + K_{\Delta\theta}\dot{\lambda}_{dq} \\ &\Rightarrow \mathbf{u}_{\gamma\delta} = r_s\mathbf{i}_{\gamma\delta} + \omega\mathbf{J}_s\lambda_{\gamma\delta} + (-\dot{\bar{\theta}}\mathbf{J}_s\lambda_{\gamma\delta} + \dot{\lambda}_{\gamma\delta}) \\ &\Rightarrow \mathbf{u}_{\gamma\delta} = r_s\mathbf{i}_{\gamma\delta} + \hat{\omega}\mathbf{J}_s\lambda_{\gamma\delta} + \dot{\lambda}_{\gamma\delta} \end{aligned} \quad (4.2.5)$$

$$K_{\Delta\theta} = \begin{bmatrix} \cos \bar{\theta} & -\sin \bar{\theta} \\ \sin \bar{\theta} & \cos \bar{\theta} \end{bmatrix} \quad (4.2.6)$$

$$K_{\Delta\theta}\lambda_{dq} = K_{\Delta\theta}\mathbf{L}_{qq}K_{\Delta\theta}^{-1}K_{\Delta\theta}\mathbf{i}_{dq} + K_{\Delta\theta}\lambda_{mmdq} = \mathbf{L}_{qq}\mathbf{i}_{\gamma\delta} + \lambda_{mm\gamma\delta} \quad (4.2.7)$$

The expression below describes the modified rotor flux vector $\lambda_{mm\gamma\delta}$ of a Permanent Magnet Synchronous Motor (PMSM), after it has been transformed from the dq -frame into a rotated δ -frame using an estimated angular error $\bar{\theta}$. Frame Transformation [4-11], [4-19], [4-22]:

$$\lambda_{mm\gamma\delta} = K_{\Delta\theta}\lambda_{mmdq} \quad (4.2.8)$$

Where:

$K_{\Delta\theta}$ is a 2D rotation matrix based on the angle error $\bar{\theta}$, given by [4-11], [4-19], [4-22]:

$$K_{\Delta\theta} = \begin{bmatrix} \cos \bar{\theta} & -\sin \bar{\theta} \\ \sin \bar{\theta} & \cos \bar{\theta} \end{bmatrix} \quad (4.2.9)$$

Modified Rotor Flux Vector in dq -Frame [4-11], [4-19], [4-22]:

$$\lambda_{mmdq} = [\lambda_m + (L_d - L_q)i_d] \begin{bmatrix} 1 \\ 0 \end{bmatrix} \quad (4.2.10)$$

This shows that the vector in the dq -frame has only a d -axis component, consisting of:

- λ_m : the permanent magnet flux
- $(L_d - L_q)i_d$: the saliency-based flux from i_d , where L_d and L_q are d- and q-axis inductances

Apply Rotation to Convert to δ frame [4-11], [4-19], [4-22]:

$$\lambda_{mmy\delta} = K_{\Delta\theta}\lambda_{mmdq} = \begin{bmatrix} \cos \bar{\theta} \\ \sin \bar{\theta} \end{bmatrix} [\lambda_m + (L_d - L_q)i_d] \quad (4.2.11)$$

Which gives:

$$\lambda_{mmy\delta} = \begin{bmatrix} \lambda_m \cos \bar{\theta} \\ \lambda_m \sin \bar{\theta} \end{bmatrix} + (L_d - L_q) \begin{bmatrix} i_d \cos \bar{\theta} \\ i_d \sin \bar{\theta} \end{bmatrix} \quad (4.2.12)$$

Physical Interpretation [4-11], [4-19], [4-22]:

The vector $\lambda_{mmy\delta}$ represents the rotor flux as seen from a rotated reference frame. It contains two parts:

- The rotated permanent magnet flux vector
- The rotated saliency-induced flux vector caused by i_d

This decomposition is key to understanding and implementing sensorless control algorithms, especially those based on saliency (e.g., Sliding Mode Observer, High-Frequency Injection), where rotor position is estimated based on flux behavior under rotation. And, $\lambda_{mmy\delta}$ represents the modified rotor magnetic flux from $\gamma\delta$ defined by [4-11], [4-19], [4-22]:

$$\begin{aligned}
\lambda_{m\gamma\delta} &= K_{\Delta\theta}\lambda_{mmdq} = \begin{bmatrix} \cos \bar{\theta} & -\sin \bar{\theta} \\ \sin \bar{\theta} & \cos \bar{\theta} \end{bmatrix} \lambda_{mmdq} \\
&= \begin{bmatrix} \cos \bar{\theta} & -\sin \bar{\theta} \\ \sin \bar{\theta} & \cos \bar{\theta} \end{bmatrix} \begin{bmatrix} \lambda_m + (L_d - L_q)i_d \\ 0 \end{bmatrix} \\
&= [\lambda_m + (L_d - L_q)i_d] \begin{bmatrix} \cos \bar{\theta} \\ \sin \bar{\theta} \end{bmatrix} = \begin{bmatrix} \lambda_m \cos \bar{\theta} \\ \lambda_m \sin \bar{\theta} \end{bmatrix} + (L_d - L_q) \begin{bmatrix} i_d \cos \bar{\theta} \\ i_d \sin \bar{\theta} \end{bmatrix}
\end{aligned} \tag{4.2.13}$$

It should be highlighted that the right-hand component of Equation (4.2.13) incorporates a term influenced by magnetic saliency $(L_d - L_q)$, the current i_d , and the angular deviation $\Delta\theta = \bar{\theta}$. Within the scope of HFI-based techniques, it is more practical to substitute i_d with its corresponding form, expressed as a function of $\gamma\delta$ -axis currents and the angle offset $\Delta\theta$. [4-11], [4-19], [4-22]

4.2.2.4 PMSM Inductance Matrix $L_{\gamma\delta}$ and Saliency Effects

Now, recognizing that $i_d = i_\gamma \cos \Delta\bar{\theta} + i_\delta \sin \Delta\bar{\theta}$, and that $\Delta L = (L_d - L_q)/2$, the magnetic saliency term $(L_d - L_q)$ present in Equation (4.2.8) can equivalently be rewritten in the following form [4-11], [4-30]:

$$\begin{aligned}
(L_d - L_q) \begin{bmatrix} i_d \cos \bar{\theta} \\ i_d \sin \bar{\theta} \end{bmatrix} &= (L_d - L_q) \begin{bmatrix} i_\gamma (\cos \bar{\theta})^2 + i_\delta (\sin \bar{\theta} \cos \bar{\theta}) \\ i_\gamma (\sin \bar{\theta} \cos \bar{\theta}) + i_\delta (\sin \bar{\theta})^2 \end{bmatrix} \\
&= \frac{1}{2}(L_d - L_q) \begin{bmatrix} (1 + \cos 2\bar{\theta})/\sin 2\bar{\theta} \\ \sin 2\bar{\theta}/(1 - \cos 2\bar{\theta}) \end{bmatrix} \begin{bmatrix} i_\gamma \\ i_\delta \end{bmatrix} = \Delta L_{\gamma\delta} \begin{bmatrix} i_\gamma \\ i_\delta \end{bmatrix}
\end{aligned} \tag{4.2.14}$$

Here, the matrix $\Delta L_{\gamma\delta}$ represents the differential inductance of the PMSM in $\gamma\delta$ -axis, which explicitly depends on the saliency factor $(L_d - L_q)$ and the angular displacement $\Delta\theta$. This matrix is described as:

$$\Delta L_{\gamma\delta} = \begin{bmatrix} \Delta L(1 + \cos 2\bar{\theta}) & \Delta L \sin 2\bar{\theta} \\ \Delta L \sin 2\bar{\theta} & \Delta L(1 - \cos 2\bar{\theta}) \end{bmatrix} \tag{4.2.15}$$

By substituting Equation (4.2.14) into Equation (4.2.13), the modified rotor magnetic flux $\lambda_{m\gamma\delta}$

can be represented as the aggregate of its component flux terms [4-11], [4-30]:

$$\lambda_{m\gamma\delta} = \lambda_{m\gamma\delta} + \Delta L_{\gamma\delta} \mathbf{i}_{\gamma\delta} \tag{4.2.16}$$

Where:

$$\lambda_{m\gamma\delta} = \lambda_m \begin{bmatrix} \cos \bar{\theta} \\ \sin \bar{\theta} \end{bmatrix} \quad (4.2.17)$$

Similarly, the stator magnetic flux expression given in Equation (4.2.7) can be reformulated as follows:

$$\begin{aligned} \lambda_{\gamma\delta} &= \mathbf{L}_{qq} \mathbf{i}_{\gamma\delta} + \lambda_{mm\gamma\delta} \\ &= \mathbf{L}_{qq} \mathbf{i}_{\gamma\delta} + \lambda_{m\gamma\delta} + \Delta L_{\gamma\delta} \mathbf{i}_{\gamma\delta} \\ &= (\mathbf{L}_{qq} + \Delta L_{\gamma\delta}) \mathbf{i}_{\gamma\delta} + \lambda_{m\gamma\delta} \\ &= \mathbf{L}_{\gamma\delta} \mathbf{i}_{\gamma\delta} + \lambda_{m\gamma\delta} \end{aligned} \quad (4.2.18)$$

In this case, the inductance matrix $L_{\gamma\delta}$ is specified by the sum $L_{\gamma\delta} = L_{qq} + \Delta L_{\gamma\delta}$. Utilizing Equation (4.2.15) and recalling that $\Sigma L = L_q + \Delta L$, the inductance matrix in Equation (4.2.18) may alternatively be reformulated as:

$$\begin{aligned} \mathbf{L}_{\gamma\delta} &= \mathbf{L}_{qq} + \Delta L_{\gamma\delta} \\ &= \begin{bmatrix} L_q + \Delta L(1 + \cos 2\bar{\theta}) & \Delta L \sin 2\bar{\theta} \\ \Delta L \sin 2\bar{\theta} & L_q + \Delta L(1 - \cos 2\bar{\theta}) \end{bmatrix} \\ &= \begin{bmatrix} \Sigma L + \Delta L \cos 2\bar{\theta} & \Delta L \sin 2\bar{\theta} \\ \Delta L \sin 2\bar{\theta} & \Sigma L - \Delta L \cos 2\bar{\theta} \end{bmatrix} \\ &= \Sigma L \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} + \Delta L \begin{bmatrix} \cos 2\bar{\theta} & \sin 2\bar{\theta} \\ \sin 2\bar{\theta} & -\cos 2\bar{\theta} \end{bmatrix} \end{aligned} \quad (4.2.19)$$

From Equation (4.2.19), it follows that the matrix $L_{\gamma\delta}$ is a function of the mean inductance ΣL , the saliency-based difference ΔL , and twice the angle deviation $2\Delta\theta$. It is important to emphasize that this same outcome is derived through the inductance matrix $L_{\gamma\delta}$, through the explicit transformation of the inductance matrix L_{dq} into the target reference frame, the result is given by [4-11], [4-30]:

$$\mathbf{L}_{\gamma\delta} = K_{\Delta\theta} \mathbf{L}_{dq} K_{\Delta\theta}^{-1} \quad (4.2.20)$$

Upon performing integration on both sides of Equation (4.2.6) and $L_{\gamma\delta}$ is [4-11], [4-30]:

$$\lambda_{\gamma\delta} = \int_0^t (\mathbf{u}_{\gamma\delta} - r_s \mathbf{i}_{\gamma\delta} - \dot{\omega} \mathbf{J}_s \lambda_{\gamma\delta}) dt \quad (4.2.21)$$

By inserting Equation (4.2.21) into Equation (4.2.18) and isolating the current vector $i_{\gamma\delta}$, the resulting expression becomes:

$$\begin{aligned}
\mathbf{L}_{\gamma\delta}\mathbf{i}_{\gamma\delta} + \boldsymbol{\lambda}_{m\gamma\delta} &= \int_0^t (\mathbf{u}_{\gamma\delta} - r_s\mathbf{i}_{\gamma\delta} - \hat{\omega}\mathbf{J}_s\boldsymbol{\lambda}_{\gamma\delta}) dt \\
\Rightarrow \mathbf{i}_{\gamma\delta} &= \mathbf{L}_{\gamma\delta}^{-1} \left[\int_0^t (\mathbf{u}_{\gamma\delta} - r_s\mathbf{i}_{\gamma\delta} - \hat{\omega}\mathbf{J}_s\boldsymbol{\lambda}_{\gamma\delta}) dt - \boldsymbol{\lambda}_{m\gamma\delta} \right]
\end{aligned} \tag{4.2.22}$$

Moreover, the inverse of the inductance matrix $L_{\gamma\delta}$, is formulated as:

$$(\mathbf{L}_{\gamma\delta})^{-1} = \frac{1}{\det(\mathbf{L}_{\gamma\delta})} \cdot \text{adj}(\mathbf{L}_{\gamma\delta}) = \frac{1}{L_d L_q} \begin{bmatrix} \Sigma L - \Delta L \cos 2\bar{\theta} & -\Delta L \sin 2\bar{\theta} \\ -\Delta L \sin 2\bar{\theta} & \Sigma L + \Delta L \cos 2\bar{\theta} \end{bmatrix} \tag{4.2.23}$$

In this context, the determinant $\det(L_{\gamma\delta})$ and the corresponding adjugated matrix $\text{adj}(L_{\gamma\delta})$ are evaluated using the following expressions [4-11], [4-30]:

$$\begin{aligned}
\det(\mathbf{L}_{\gamma\delta}) &= L_q^2 + [L_q \Delta L (1 + \cos 2\bar{\theta} + 1 - \cos 2\bar{\theta})] + (\Delta L)^2 [1 - (\cos 2\bar{\theta})^2 - (\sin 2\bar{\theta})^2] \\
&= L_q^2 + 2L_q \Delta L + \Delta L(1 - 1) = L_q(L_q + 2\Delta L) = L_q[L_q + (L_d - L_q)] = L_q L_d \neq 0,
\end{aligned} \tag{4.2.24}$$

While:

$$\begin{aligned}
\text{adj}(\mathbf{L}_{\gamma\delta}) &= \begin{bmatrix} L_q + \Delta L(1 - \cos 2\bar{\theta}) & -\Delta L \sin 2\bar{\theta} \\ -\Delta L \sin 2\bar{\theta} & L_q + \Delta L(1 + \cos 2\bar{\theta}) \end{bmatrix} \\
&= \begin{bmatrix} \Sigma L - \Delta L \cos 2\bar{\theta} & -\Delta L \sin 2\bar{\theta} \\ -\Delta L \sin 2\bar{\theta} & \Sigma L + \Delta L \cos 2\bar{\theta} \end{bmatrix}'
\end{aligned} \tag{4.2.25}$$

By analyzing the expressions on the right-hand side of Equation (4.2.23), it becomes evident that the current terms $i_{\gamma\delta}$ in equation (4.2.22) are dependent on the parameters $L_d, L_q, \Sigma L, \Delta L, \Delta\theta$. This indicates that the presence of angular error information within the stator current enables rotor angle estimation when the current signal $i_{\gamma\delta}$, is properly processed [4-11],[4-30].

4.2.3 High-Frequency Injection (HFI) of Stator Voltage for Rotor Angle Determination

4.2.3.1 Examination of High-Frequency Stator Current in the $\gamma\delta$ Rotating Frame

The change in stator inductance resulting from variations in the rotor angle signifies the presence of magnetic saliency. Within injection-based strategies, this magnetic saliency characteristic plays a crucial role, as it facilitates the identification of the PMSM's rotor speed and angular position. By evaluating the PMSM's response to an imposed high-frequency current, precise rotor angle estimation along with magnetic saliency computation becomes achievable.

Three principal techniques for voltage injections are widely recognized: the first involves applying a sinusoidal voltage waveform in the $\alpha\beta$ stationary frame; the second introduces a sinusoidal excitation within the dq synchronous frame; and the third utilizes discrete voltage pulses, also injected in the dq frame. A notable advantage shared by these methods is their capability to estimate the rotor flux orientation even when the machine operates at extremely low or zero velocity. [4-35]

In the present approach, the injected voltage $u_{i\gamma\delta}$ is a continuous sine wave, layered on top of the base electrical supply and formulated within the $\gamma\delta$ -estimated frame. This high-frequency voltage vector $u_{i\gamma\delta}$ revolves with an amplitude u_{im} and an angular velocity $\omega_i = 2\pi f_i$ with respect to the $\gamma\delta$ reference frame. As such, the expression for the high-frequency voltage $u_{i\gamma\delta}$ is written as [4-11], [4-30]:

$$\mathbf{u}_{i\gamma\delta} = u_{im} \begin{bmatrix} -\sin \theta_i \\ \cos \theta_i \end{bmatrix} = u_{im} \begin{bmatrix} -\sin \omega_i t \\ \cos \omega_i t \end{bmatrix} \quad (4.2.26)$$

Where:

$$\theta_i = \int_0^t \omega_i dt \quad (4.2.27)$$

By considering Equations (4.2.21) and (4.2.22), the resulting components of the stator flux and current induced by the injected high-frequency voltage can then be determined accordingly using the following expressions [4-11], [4-30]:

$$\lambda_{i\gamma\delta} = \int_0^t (\mathbf{u}_{i\gamma\delta} - r_s \mathbf{i}_{i\gamma\delta} - \hat{\omega} \mathbf{J}_s \lambda_{i\gamma\delta}) dt \quad (4.2.28)$$

Also:

$$\mathbf{L}_{\gamma\delta} \mathbf{i}_{\gamma\delta} = \lambda_{i\gamma\delta} \Leftrightarrow \mathbf{i}_{\gamma\delta} = \mathbf{L}_{\gamma\delta}^{-1} \lambda_{i\gamma\delta} \quad (4.2.29)$$

By inserting Equation (4.2.28) into Equation (4.2.29), the resulting expression becomes:

$$\mathbf{i}_{\gamma\delta} = \mathbf{L}_{\gamma\delta}^{-1} \left[\int_0^t (\mathbf{u}_{i\gamma\delta} - r_s \mathbf{i}_{\gamma\delta} - \hat{\omega} \mathbf{J}_s \lambda_{i\gamma\delta}) dt \right] \quad (4.2.30)$$

Assuming that the frequency f_i is sufficiently high such that $\hat{\omega} \ll \omega_i$, $r_s \ll \omega_i L_d$, and $r_s \ll \omega_i L_q$, the terms corresponding to the high-frequency voltage drop across the stator resistance can be neglected. Furthermore, since the integral of the injected voltage $u_{i\gamma\delta}$ dominates Equation (4.2.30), the expression for the high-frequency current is simplified as follows [4-11], [4-30]:

$$\begin{aligned}
\mathbf{i}_{i\gamma\delta} &\approx \mathbf{L}_{\gamma\delta}^{-1} \left[\int_0^t \mathbf{u}_{i\gamma\delta} dt \right] \\
&= \frac{1}{L_d L_q} \begin{bmatrix} \Sigma L - \Delta L \cos 2\bar{\theta} & -\Delta L \sin 2\bar{\theta} \\ -\Delta L \sin 2\bar{\theta} & \Sigma L + \Delta L \cos 2\bar{\theta} \end{bmatrix} \cdot \left[\frac{u_{im}}{\omega_i} \begin{bmatrix} \cos \omega_i t \\ \sin \omega_i t \end{bmatrix} \right] \\
&= \frac{u_{im}}{\omega_i L_d L_q} \begin{bmatrix} \Sigma L \cos \omega_i t - \Delta L \cos(2\bar{\theta} - \omega_i t) \\ \Sigma L \sin \omega_i t - \Delta L \sin(2\bar{\theta} - \omega_i t) \end{bmatrix}
\end{aligned} \tag{4.2.31}$$

By making use of Euler's identity, the resulting expression becomes:

$$\begin{aligned}
\mathbf{i}_{i\gamma\delta} &= \frac{u_{im}}{\omega_i L_d L_q} \{ \Sigma L (\cos \omega_i t + j \sin \omega_i t) - \Delta L [\cos(2\bar{\theta} - \omega_i t) + j \sin(2\bar{\theta} - \omega_i t)] \} \\
&= \frac{u_{im}}{\omega_i L_d L_q} [\Sigma L e^{j\omega_i t} - \Delta L e^{j(2\bar{\theta} - \omega_i t)}] = i_{i\gamma\delta p} + i_{i\gamma\delta n}
\end{aligned} \tag{4.2.32}$$

Where $i_{i\gamma\delta p}$ and $i_{i\gamma\delta n}$ represent the high-frequency current components, which are defined as follows:

$$i_{i\gamma\delta p} = \frac{u_{im}}{\omega_i L_d L_q} \Sigma L e^{j\omega_i t} \tag{4.2.33}$$

Also:

$$i_{i\gamma\delta n} = -\frac{u_{im}}{\omega_i L_d L_q} \Delta L e^{j(2\bar{\theta} - \omega_i t)} \tag{4.2.34}$$

In this context, $i_{i\gamma\delta p}$ and $i_{i\gamma\delta n}$ denote the high-frequency stator current components corresponding to positive and negative rotational frequencies, respectively. As anticipated, the stator current vector resulting from superposition consists of two independent components: $i_{i\gamma\delta p}$, which rotates in the positive direction with angular velocity ω_i , and $i_{i\gamma\delta n}$, which rotates in the negative direction at angular velocity $-\omega_i + \frac{d(2\Delta\theta)}{dt}$. It is important to emphasize that the average inductance is incorporated into Equation (4.2.33), while Equation (4.2.34) includes the differential inductance ΔL along with the doubled rotor angle error $2\Delta\theta$. By applying appropriate signal processing and filtering techniques, key quantities such as rotor position, stator inductance, and magnetic saliency can be extracted [4-11], [4-30].

4.2.3.2 Rotor Angle Error Between dq and $\gamma\delta$ Reference Frames

According to Equation (4.2.32), the full procedure for estimating the rotor angle error is split into two distinct stages (refer to Figure 4.2.3). In the first stage, the sensorless algorithm concentrates on isolating the angle error component embedded in Equation (4.2.32). In the subsequent stage, it proceeds to estimate the rotor position by employing a sliding mode observer (SMO).

To decouple the error angle term $\Delta\theta$ from the components in Equation (4.2.32), both parts of the equation are multiplied by $e^{j\omega_i t}$, that is [4-30], [4-29]:

$$i_{i\gamma\delta} = i_{i\gamma\delta p} + i_{i\gamma\delta n} = \frac{u_{im}}{\omega_i L_d L_q} [\Sigma L e^{j\omega_i t} - \Delta L e^{j(2\bar{\theta} - \omega_i t)}] \quad (4.2.35)$$

$$i_{i\gamma\delta} e^{j\omega_i t} = \frac{u_{im}}{\omega_i L_d L_q} [\Sigma L e^{j2\omega_i t} - \Delta L e^{j2\bar{\theta}}] \quad (4.2.36)$$

The outcome presented in Equation (4.2.36) indicates that modulation, achieved by multiplying with the unit phasor $e^{j\omega_i t}$, produces a resulting vector comprising two components that rotate with angular frequencies $2\omega_i$ and $2\frac{d\bar{\theta}}{dt}$, respectively.

Figure 4.2.3 Flowchart illustrating the high-frequency (HF) current signal processing stages:

- (1) modulation of the stator current signal,
- (2) application of low-pass and band-pass filters (LPF and BPF),
- (3) estimation of rotor speed and position, and
- (4) extraction of inductance parameters and magnetic saliency information.

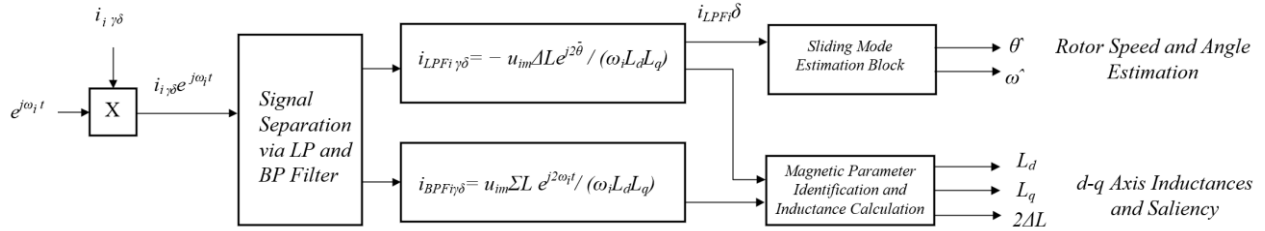


Figure 4.2.3 High-frequency (HF) current signal processing stages

Subsequently, when this high-frequency current—expressed in Equation (4.2.36)—is passed through a low-pass filter (LPF), the resulting current retains only the components proportional to $\sin 2\bar{\theta}$ and $\cos 2\bar{\theta}$, and takes the form given as [4-11], [4-29], [4-28]:

$$[i_{\gamma\delta} e^{j\omega_i t}]_{LPF} = i_{LPF\ i\gamma\delta} = -\frac{u_{im}\Delta L}{\omega_i L_d L_q} e^{j2\bar{\theta}} = k_i \begin{bmatrix} \cos 2\bar{\theta} \\ \sin 2\bar{\theta} \end{bmatrix} = \begin{bmatrix} k_i \cos 2\hat{\theta} \\ k_i \sin 2\hat{\theta} \end{bmatrix} \simeq k_i 2\bar{\theta}, \quad (4.2.37)$$

Where:

$$k_i = -\frac{u_{im}\Delta L}{\omega_i L_d L_q}, \quad (4.2.38)$$

This suggests that the angular information can be retrieved from the angle deviation that results from directly filtering the modulated high-frequency stator current signal as shown in Equation (4.2.36). Evidently, the low-pass filtered output from Equation (4.2.37) offers a suitable structure for use in both speed and position estimation tasks.

In the subsequent step, the signals produced by the LPF serve as the input to a sliding mode observer (SMO), which is employed to determine the rotor flux position and rotational speed of the PMSM (refer to Figure 4.2.3). Similarly, the initial segment of the high-frequency current signal contained in Equation (4.2.36) is also extracted by directing the modulated current vector $i_{\gamma\delta} e^{j\omega_i t}$ through a band-pass filter (BPF). Consequently, the BPF output retains only the high-frequency current component, as expressed by [4-11], [4-29], [4-28]:

$$[i_{\gamma\delta} e^{j\omega_i t}]_{BPF} = i_{BPF \gamma\delta} = \frac{u_{im} \Sigma L}{\omega_i L_d L_q} e^{j2\omega_i t} = k_j \begin{bmatrix} \cos 2\omega_i t \\ \sin 2\omega_i t \end{bmatrix}, \quad (4.2.39)$$

Where:

$$k_j = \frac{u_{im} \Sigma L}{\omega_i L_d L_q}, \quad (4.2.40)$$

Taking Equations (4.2.38) and (4.2.40) into account, it becomes evident that the parameter k_i is dependent on the values of L_d, L_q , and the differential inductance component, while k_j characterizes the amplitude of the BPF output and varies as a function of both L_d, L_q , and their mean value. The schematic representation of the current signal processing workflow is provided in Figure 4.2.3. As depicted, only the LPF-derived δ component is utilized in estimating speed and position, whereas both coefficients k_i and k_j , obtained from LPF and BPF respectively, are applied in determining the values of L_d, L_q , and the magnetic saliency defined by $(L_d - L_q)$. [4-11], [4-29], [4-28]

4.2.3.3 Rotor Flux Angle Error in the $\gamma\delta$ Reference Frame

Although Equation (4.2.36) yields two orthogonal components, namely $k_i \cos 2\bar{\theta}$ and $k_i \sin 2\bar{\theta}$, the estimation algorithm for rotor speed and position requires only the term proportional to $k_i \sin 2\bar{\theta}$. Figure 4.2.4 presents the internal configuration of the observer designed for precise tracking of rotor speed and flux position. Referring to the $i_{LPF i\delta}$ signal extracted from Equation (4.2.37), the estimated rotor flux angle is linked to the actual angle through the relation derived from the third Ptolemy identity, or equivalently, the sine subtraction formula, as also shown in Figure 4.2.4:

$$i_{LPF i\delta} = k_i \sin 2\bar{\theta} = k_i \sin(2\theta - 2\hat{\theta}) = k_i \sin 2\theta \cos 2\hat{\theta} - k_i \cos 2\theta \sin 2\hat{\theta}, \quad (4.2.41)$$

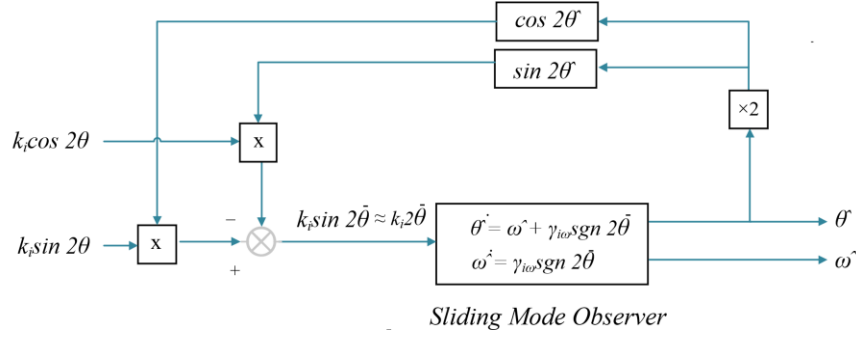


Figure 4.2.4. An alternative block diagram depicting the PMSM speed and position estimation scheme based on a sliding mode observer (SMO).

Equation (4.2.41) is depicted schematically in Figure 4.2.4, illustrating the way in which the angle error is correspondingly linked to both the actual and estimated rotor speeds through the signal $i_{LPFi\delta}$. When the deviation between the estimated and the true rotor flux position is sufficiently small, the current $i_{LPFi\delta}$ can be approximated by $k_i 2\bar{\theta}$ based on the relation in $2\bar{\theta} \approx 2\bar{\theta} = 2\Delta\theta = 2\theta - 2\hat{\theta}$

This indicates that $i_{LPFi\delta}$ is nearly linearly proportional to $\Delta\theta$ under small-angle conditions.

4.2.3.4 Sliding Mode Observer (SMO) Design for Rotor Flux Angle and Speed Estimation

Among the various observer techniques reported in the literature, sliding mode observers (SMOs) have gained wide attention and have been extensively utilized in a broad range of industrial applications. The key strength of sliding mode control lies in its inherent robustness and the ability to achieve finite-time convergence even in the presence of system disturbances and model inaccuracies.

In constructing the SMO, the design typically involves two sequential steps: defining the sliding manifold and formulating the control law. In this work, the sliding manifold is designed by selecting the rotor angle and speed estimation errors, $\bar{\theta}$ and $\bar{\omega}$, as the sliding surfaces. The control law utilizes sign functions $\text{sgn}(2\bar{\theta})$, while the observer gains—denoted $\gamma_{i\theta}$ and $\gamma_{i\omega}$ —are selected respectively for angle and speed tracking. Moreover, the equivalent control approach principles are incorporated to ensure the stability of the sliding mode observer during the post-reaching phase. Drawing from Equation (4.2.41), the proposed observer for rotor angle and speed estimation is formulated as follows [4-31], [4-32]:

$$\dot{\hat{\theta}} = \hat{\omega} + \gamma_{i\theta} \text{sgn}(2\bar{\theta}), \quad (4.2.42)$$

$$\dot{\hat{\omega}} = \gamma_{i\omega} \text{sgn}(2\bar{\theta}), \quad (4.2.43)$$

Here, $\text{sgn}(2\bar{\theta})$ refers to the second-order signum function. Figure 4.2.4 presents a schematic representation of the SMO's input and output signals via an equivalent block diagram.

4.2.3.5 Dynamic Behavior and Stability of the SMO

By applying the previously defined sliding manifold along with Equations (4.2.42) and (4.2.43), the resulting dynamics of the SMO can be formulated as follows [4-31], [4-32]:

$$\dot{\bar{\theta}} = \dot{\theta} - \dot{\hat{\theta}} = \omega - \hat{\omega} - \gamma_{i\theta} \operatorname{sgn}(2\bar{\theta}) = \bar{\omega} - \gamma_{i\theta} \operatorname{sgn}(2\bar{\theta}), \quad (4.2.44)$$

$$\dot{\bar{\omega}} = \dot{\omega} - \dot{\hat{\omega}} = -\gamma_{i\omega} \operatorname{sgn}(2\bar{\theta}), \quad (4.2.45)$$

Assume now that the value of $\gamma_{i\theta}$ is sufficiently high, satisfying the following condition:

$$\gamma_{i\theta} \gg |\bar{\omega}| \geq \bar{\omega}, \quad (4.2.46)$$

That is, i serves as an upper bound for the corresponding term, $\bar{\omega}$ and therefore [4-32], [4-33]:

$$(4.2.44) \quad \dot{\bar{\theta}} = 0 \Rightarrow \bar{\omega} = \gamma_{i\theta} (\operatorname{sgn}(2\bar{\theta}))_{eq} \Leftrightarrow (\operatorname{sgn}(2\bar{\theta}))_{eq} = \frac{\bar{\omega}}{\gamma_{i\theta}}, \quad (4.2.47)$$

In this context, the notation $(\cdot)_{eq}$ designates the equivalent control action. Ultimately, this equivalent control is utilized as the input in the speed estimation process. By substituting Equation (4.2.43) into Equation (4.2.45), we obtain the following result:

$$\dot{\bar{\omega}} = -\gamma_{i\omega} (\operatorname{sgn}(2\bar{\theta}))_{eq} = -\frac{\gamma_{i\omega}}{\gamma_{i\theta}} \bar{\omega}, \quad (4.2.48)$$

The stability of the proposed sliding mode observer can be further interpreted using a Lyapunov function that accounts for both position and speed estimation errors. By defining the Lyapunov candidate as

$V = \frac{1}{2} \bar{\theta}^2 + \frac{1}{2} \alpha \bar{\omega}^2$, where $\alpha > 0$ is a weighting coefficient, the coupled dynamics of the estimation errors can be analyzed. This formulation captures the interaction between the position error $\bar{\theta}$ and the speed error $\bar{\omega}$, ensuring that both states converge to zero under appropriate selection of the observer gains. In particular, the switching gains $\gamma_{i\theta}$ and $\gamma_{i\omega}$ govern the convergence rate of the position and speed estimation errors, respectively.

The conducted stability analysis confirms that the sliding mode observer defined by Equations (4.2.42) and (4.2.43) is asymptotically stable, with the estimation errors converging to zero. Referring to Equation (4.2.48), it becomes evident that the proposed SMO framework functions as a reduced-order asymptotic observer utilizing the principle of equivalent control. With respect to the reaching phase, the sliding surface condition $\bar{\theta} = 0$ is attained within a finite time interval t_r , which is determined by:

$$t_r \leq \frac{2V^{1/2}(0)}{\xi} = \frac{\sqrt{2}|\bar{\theta}(0)|}{(\gamma_{i\theta} - P_d)}, \quad (4.2.49)$$

Where:

$$\zeta = \sqrt{2}(\gamma_{i\theta} - P_d), \quad (4.2.50)$$

The term P_d denotes the upper threshold of the bounded disturbance associated with $\bar{\omega}$, the dynamics. Moreover, considering Equation (4.2.48), the general solution to the system converges asymptotically to zero, described by the following relation [4-32], [4-33]:

$$\frac{\dot{\bar{\omega}}}{\gamma_{i\theta}} = -\frac{\gamma_{i\omega}}{\gamma_{i\theta}} \bar{\omega} \Rightarrow \bar{\omega}(t) = \bar{\omega}(0)e^{-(\gamma_{i\omega}/\gamma_{i\theta})t}, \quad (4.2.51)$$

From the inspection of Equation (4.2.49), it is evident that the maximum time t_r is directly proportional to the initial angular deviation $|\bar{\theta}(0)|$, and inversely proportional to the gain parameter $\gamma_{i\theta}$, furthermore, the speed error diminishes toward zero at an accelerated rate as the ratio $\gamma_{i\omega}/\gamma_{i\theta}$ increases.

4.2.3.6 Approximation of the $\text{sgn}(\cdot)$ Function – Mitigation of Chattering

Despite the advantages and improved stability provided by sliding mode observers, the resulting control signal tends to induce high-frequency oscillations once the system states $\bar{\theta}$ and $\bar{\omega}$ arrive at the designated sliding surfaces. These unwanted oscillatory behaviors are commonly known as the "chattering phenomenon," which adversely affects the performance of sliding mode implementations. As a consequence, such chattering often excites unmodeled system dynamics, potentially leading to significant estimation errors or malfunction of the observer. A conventional strategy to address this issue is to replace the discontinuous sign function with a smooth and continuous approximation. This replacement technique has proven effective in mitigating the adverse effects of chattering. Among various alternatives, the hyperbolic tangent function $\tanh(\cdot)$ is widely regarded as a relatively simple and effective method for approximating the sign function. Specifically, it helps reduce chattering since [4-33], [4-34]:

$$\text{sgn}(2\bar{\theta}) \approx \tanh(k2\bar{\theta}) \quad \text{for } k \gg 1. \quad (4.2.52)$$

4.2.4 Estimating the Inductances and Magnetic Saliency of PMSM

4.2.4.1 Computation of Parameters k_i and k_j

Given that the $\gamma\delta$ rotating reference frame is orthogonal, the parameter k_i can be derived as the hypotenuse of a right triangle, where the legs correspond to the LPF output components (by applying the Pythagorean theorem). This results in the square root of the sum of squares of the current signals $i_{LPF\gamma}$ and $i_{LPF\delta}$ as described in Equation (4.2.37), that is [4-11], [4-29]:

$$k_i = - \left[\sqrt{(i_{LPF\gamma})^2 + (i_{LPF\delta})^2} \right] \text{sgn}(\Delta L), \quad (4.2.53)$$

Alternatively, the value of k_i may also be determined from the component $i_{LPFi\gamma}$ in Equation (4.2.37), assuming that the angular estimation error is negligible (i.e., $k_i \cos 2\bar{\theta} = k_i$). The sign of k_i is inversely related to the sign of ΔL ; specifically, if $\Delta L > 0$, then $k_i < 0$, and conversely, if $\Delta L < 0$, then $k_i > 0$. In typical motor operation of the PMSM, the condition $\Delta L < 0$ holds, which implies that $k_i > 0$. Similarly, the parameter k_j is strictly positive and is computed from the BPF output components within the $\gamma\delta$ coordinate frame as follows [4-11], [4-29]:

$$k_j = \left[\sqrt{(i_{BPF\gamma})^2 + (i_{BPF\delta})^2} \right], \quad (4.2.54)$$

From Equations (4.2.53) and (4.2.54), it follows that the magnitudes of both parameters k_i and k_j essentially represent the amplitudes of the low- and high-frequency current components described in Equation (4.2.36). Moreover, comparing the magnitudes and absolute values of k_i and k_j , we observe that $k_j > k_i$ and $|k_j| > |k_i|$, due to the relation $\Sigma L > -\Delta L$ and $\Sigma L > |\Delta L|$.

4.2.4.2 Representing dq-Axis Impedances via k_i and k_j Parameters

Examining Equations (4.2.38) and (4.2.40), it can be inferred that parameters k_i and k_j are respectively proportional ΔL and ΣL , with a ratio coefficient of $\pm \frac{u_m}{\omega_i L_d L_q}$. Additionally, the summation and subtraction of ΔL and ΣL yield L_d and L_q , respectively. That is:

$$(\Sigma L + \Delta L) = \left[\frac{L_d + L_q}{2} \right] + \left[\frac{L_d - L_q}{2} \right] = L_d \quad (4.2.55)$$

$$(\Sigma L - \Delta L) = \left(\frac{L_d + L_q}{2} \right) - \left(\frac{L_d - L_q}{2} \right) = L_q \quad (4.2.56)$$

This confirms that both d-axis and q-axis inductances of the PMSM can be directly calculated based on the values of k_i and k_j . By combining Equations (4.2.38) and (4.2.40) accordingly, the result follows [4-11], [4-28]:

$$k_j + k_i = \frac{u_{im}(\Sigma L - \Delta L)}{\omega_i L_d L_q} = \frac{u_{im} L_q}{\omega_i L_d L_q} = \frac{u_{im}}{\omega_i L_d}, \quad (4.2.57)$$

Rearranging Equation (4.2.57) to solve for L_d , the d-axis inductance is expressed as:

$$L_d = \frac{u_{im}}{\omega_i (k_j + k_i)}, \quad (4.2.58)$$

Furthermore, by subtracting Equation (4.2.38) from Equation (4.2.40) term-by-term, the resulting expression becomes:

$$k_j - k_i = \frac{u_{im}(\Sigma L + \Delta L)}{\omega_i L_d L_q} = \frac{u_{im} L_d}{\omega_i L_d L_q} = \frac{u_{im}}{\omega_i L_q}, \quad (4.2.59)$$

Then, solving Equation (4.2.59) with respect to L_d , the q-axis inductance can be determined as:

$$L_q = \frac{u_{im}}{\omega_i (k_j - k_i)} \quad (4.2.60)$$

4.2.4.3. Reformulation of PMSM Magnetic Saliency ($L_d - L_q$) in Terms of Parameters k_i and k_j

By substituting the expressions for L_d and L_q obtained from Equations (4.2.58) and (4.2.60), the magnetic saliency term can be reformulated as follows:

$$(L_d - L_q) = 2\Delta L = \frac{u_{im}}{\omega_i} \left[\frac{1}{k_j + k_i} + \frac{1}{k_j - k_i} \right] = \frac{u_{im}}{\omega_i} \left[\frac{2k_j}{k_j^2 - k_i^2} \right], \quad (4.2.61)$$

Equation (4.2.61) provides a direct formulation for calculating the PMSM's magnetic saliency through the parameters k_i and k_j . Upon completing the procedures outlined from Equation (4.2.53) through Equation (4.2.61), one can efficiently estimate both d- and q-axis inductances L_d and L_q as well as the magnetic saliency. The overall analysis indicates that changes in stator inductance, as well as magnetic fault conditions, can be effectively tracked using the proposed high-frequency injection (HFI) methodology.

In practical implementations, HFI-based position estimation is typically applied in the low-speed region, while back-EMF-based observers such as the sliding mode observer are used in the medium-to-high speed range. Therefore, a switching strategy between HFI and SMO is generally required in practical systems. The design of such switching logic is beyond the scope of this work and is considered future work. [4-11], [4-28]

4.3 MRAS

4.3.1 MRAS Fundamentals

Permanent magnet synchronous motors (PMSMs) are widely used in high-performance applications such as electric power steering (EPS) due to their high efficiency, compact size, and high torque density. In such applications, field-oriented control (FOC) is commonly employed to achieve precise torque and current regulation. [4-44]

In conventional PMSM drive systems, rotor position and speed information are obtained using mechanical sensors such as encoders. However, these sensors increase system cost and size, and may reduce reliability. Therefore, sensorless control techniques have become an important research topic. [4-46]

Among various sensorless estimation methods, such as back-EMF-based techniques, high-frequency signal injection, and model-based observers, the model reference adaptive system (MRAS) has attracted attention due to its relatively simple structure and fast adaptation capability. [4-39], [4-41]

In this section, an MRAS-based speed estimation method for PMSM under the FOC framework is investigated. The principle of MRAS is analyzed, and a corresponding speed estimation structure is derived.

Currently, various methods exist for estimating rotor position and speed, with the **model reference adaptive system (MRAS)** being a prominent and increasingly adopted approach due to its fast adaptability and ease of implementation. [4-47] MRAS is structured around three components: the **reference model**, the **compensated/adjustable model**, and the **adaptive law**.

In MRAS, a reference model with known parameters is used to represent the desired system behavior, while an equation with parameters—such as a speed-related equation—serves as the compensated/adjustable model. Both models operate under the same external input, and their output difference forms the basis for the adaptive law. This law adjusts the parameters of the compensated/adjustable model, driving its output to match that of the reference model.

As shown in Figure 4.3.1 MRAS operates with state vectors y (reference model) and $\hat{y}$ (adjustable model). Their difference, denoted as e , feeds into the adaptive law to generate a correction signal. This signal tunes to the adjustable model so it more closely matches the reference, achieving accurate output tracking and improved control system performance. [4-37], [4-38], [4-42]

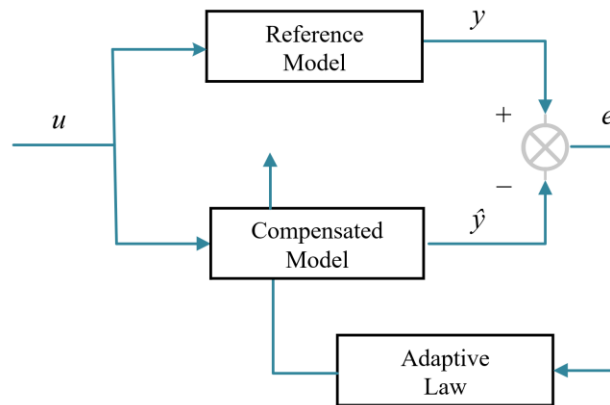


Figure 4.3.1 MRAS DIAGRAM [4-37]

4.3.2. MRAS PMSM SPEED TRACKING - Defining Reference and Adjustable Models

To construct a model reference adaptive system (MRAS) for speed estimation in a permanent magnet synchronous motor (PMSM), we begin with the mathematical representation of stator currents in the rotating d-q reference frame. [4-43]

The governing differential equation, expressed in terms of flux linkage and input voltages, is formulated as:

$$\frac{d}{dt} \begin{bmatrix} i_d + \frac{\psi_f}{L} \\ i_q \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{L} & \omega_e \\ -\omega_e & -\frac{R_s}{L} \end{bmatrix} \begin{bmatrix} i_d + \frac{\psi_f}{L} \\ i_q \end{bmatrix} + \frac{1}{L} \begin{bmatrix} U_d + \frac{R_s \psi_f}{L} \\ U_q \end{bmatrix} \quad (4.3.1)$$

As observed in the formulation above, the stator current model exhibits explicit dependence on the rotor speed ω_e . Consequently, this model is selected to serve as the **compensated/adjustable model** in the MRAS structure. Conversely, the actual motor behavior, free from explicit speed dependency, forms the **reference model**. This allows the parallel structure of MRAS to utilize discrepancies between the two models to estimate speed [4-43], [4-46].

To simplify notation, the following variable substitutions are made [4-43]:

$$i_d^* = i_d + \frac{\psi_f}{L}, \quad i_q^* = i_q, \quad U_d^* = U_d + \frac{R_s \psi_f}{L}, \quad U_q^* = U_q \quad (4.3.2)$$

With these substitutions, the governing equation becomes:

$$\frac{d}{dt} \begin{bmatrix} i_d^* \\ i_q^* \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{L} & \omega_e \\ -\omega_e & -\frac{R_s}{L} \end{bmatrix} \begin{bmatrix} i_d^* \\ i_q^* \end{bmatrix} + \frac{1}{L} \begin{bmatrix} U_d^* \\ U_q^* \end{bmatrix} \quad (4.3.3)$$

This can be represented in compact matrix form as:

$$\dot{i}^* = A i^* + B U^* \quad (4.3.4)$$

The estimated model, employing the estimated speed $\hat{\omega}_e$, takes the following form:

$$\frac{d}{dt} \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{L} & \hat{\omega}_e \\ -\hat{\omega}_e & -\frac{R_s}{L} \end{bmatrix} \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} + \frac{1}{L} \begin{bmatrix} U_d^* \\ U_q^* \end{bmatrix} \quad (4.3.5)$$

which can be written as [4-43]:

$$\dot{\hat{i}} = A \hat{i} + B U \quad (4.3.6)$$

Since the stator current vector itself is used as the measurable output, the output equation can be written as $y = i^*$.

4.3.3 Constructing the Adaptive Law

The error between the adjustable model and the reference model is quantified as [4-37]:

$$e = i^* - \hat{i} \quad (4.3.7)$$

Taking the time derivative of (4.3.7) and substituting (4.3.3) and (4.3.5), the error dynamics can be expressed as (4.3.8):

$$\dot{e} = Ae + W_{ref} - W \quad (4.3.8)$$

Here, W and W_{ref} represent the speed-mismatch terms associated with the adjustable model and the reference model, respectively. Through term W , defined as:

$$W = (\omega_e - \hat{\omega}_e) J \hat{i}, \quad W_{ref} = (\hat{\omega}_e - \omega_e) J i^* \quad (4.3.9)$$

where the skew-symmetric matrix J is [4-37]:

$$J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \quad (4.3.10)$$

This leads to an interpretation of the MRAS system as a nonlinear feedback structure, in which the difference between the two models drives the update of $\hat{\omega}_e$. [4-37] [4-46]

4.3.4 Feedback System Representation

The MRAS system can be reformulated as a nonlinear time-varying feedback structure. The block diagram is as follows:

- $W_{ref} - W$: Speed-dependent disturbance signal
- Summed with model output error
- Passed through an integrator
- Fed to a nonlinear adaptive mechanism
- Used to correct $\hat{\omega}_e$

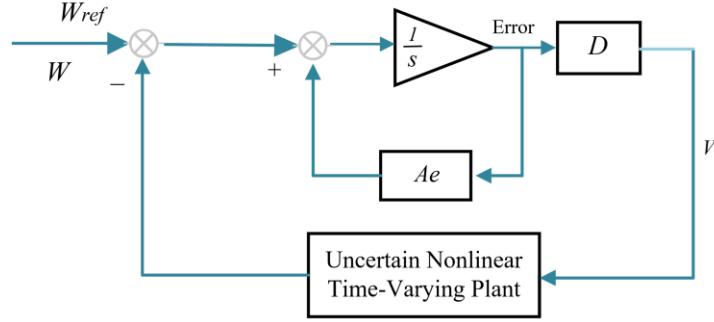


Figure 4.3.2 Block Diagram of Nonlinear Adaptive Feedback in MRAS Framework [4-37]

4.3.5. Stability Constraint via Popov Criterion

To ensure convergence and robustness, the Popov stability condition must hold [4-38], [4-40]:

$$\eta(0, t_0) = \int_0^{t_0} V^T W dt \geq -\gamma_0^2 \quad (4.3.11)$$

4.3.6 Proportional-Integral Speed Estimation

The adaptive law for updating $\hat{\omega}_e$ uses a proportional-integral format. Equation (4.3.12) is not directly derived but introduced as a general form of the adaptation law. It follows from the Popov stability criterion, which suggests that a PI-type structure ensures stability of the adaptive system. [4-38], [4-40]:

$$\hat{\omega} = \int_0^t Y_1(v, t, \tau) d\tau + Y_2(v, t) + \hat{\omega}_e(0) \quad (4.3.12)$$

Substituting into the error propagation equation gives:

$$\eta(0, t_0) = \int_0^{t_0} e^T \left[\int_0^t Y_1(v, t, \tau) d\tau + \hat{\omega}_e(0) - \omega_e \right] J \hat{i} dt + \int_0^{t_0} e^T Y_2(v, t) J \hat{i} dt = \eta_1(0, t_0) + \eta_2(0, t_0) \quad (4.3.13)$$

Thus, to satisfy (4.3.11), it must also hold that:

$$\eta_1(0, t_0) \geq -\gamma_1^2, \quad \eta_2(0, t_0) \geq -\gamma_2^2 \quad (4.3.14)$$

where γ_1, γ_2 are positive constants. [4-41]

4.3.7 Cross Product Error Formulation

The error signal for rotor speed estimation is defined as [4-48]:

$$\hat{\omega}_r = \int_0^t K_i (i_d^* \hat{i}_q - i_q^* \hat{i}_d) d\tau + K_p (i_d^* \hat{i}_q - i_q^* \hat{i}_d) + \hat{\omega}_r(0) \quad (4.3.15)$$

This can be rewritten in Laplace domain as [4-48]:

$$\hat{\omega} = \left(K_p + \frac{K_i}{s} \right) (i_d^* \hat{i}_q - i_q^* \hat{i}_d) = \left(K_p + \frac{K_i}{s} \right) \varepsilon_s \quad (4.3.16)$$

Where:

$$\varepsilon_s = i_d^* \hat{i}_q - i_q^* \hat{i}_d = \mathbf{i}^* \times \hat{\mathbf{i}} = \|\mathbf{i}^*\| \|\hat{\mathbf{i}}\| \sin \alpha_s \quad (4.3.17)$$

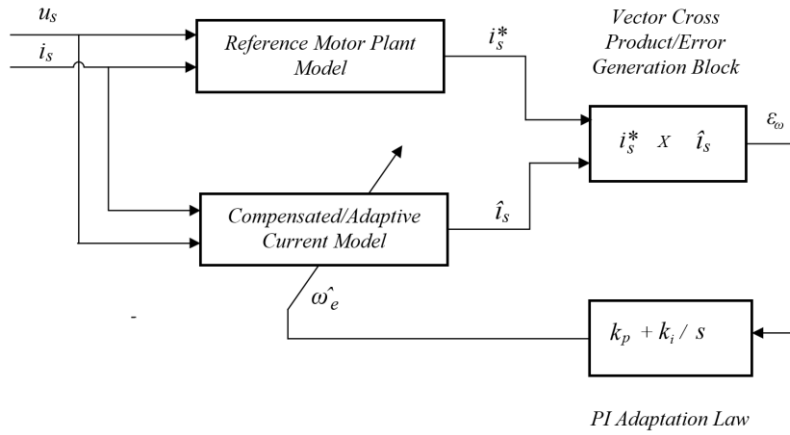


Figure 4.3.3. Structural Overview of the Stator Current Vector-Based Speed Estimation System.

[4-41], [4-42], [4-48]

4.3.8. Final Expression for Rotor Speed Estimation

The complete MRAS-based estimator, including compensation for ψ_f/L is:

$$\hat{\omega}_e = k_p (i_d \hat{i}_q - i_q \hat{i}_d) - \frac{\psi_f}{L} (i_q - \hat{i}_q) + k_i \int_0^t \left(i_d \hat{i}_q - i_q \hat{i}_d - \frac{\psi_f}{L} (i_q - \hat{i}_q) \right) dt + \hat{\omega}_e(0) \quad (4.3.18)$$

4.3.9 Restructured Control Framework and Simulation Considerations

Although a PMSM sensorless vector control system based on the MRAS algorithm can be implemented and evaluated in a MATLAB/Simulink environment, detailed modeling and quantitative validation of MRAS-based control are beyond the scope of this thesis. Conceptually, such a system would include standard functional modules such as coordinate transformation blocks (Clarke and Park), PI-based current regulators, an SVPWM generation module, and a speed estimation mechanism derived from the MRAS structure [4-39], [4-47]. In the present work, MRAS is therefore discussed primarily from a structural and comparative standpoint, serving as one of the reference estimation strategies in the EPS-oriented evaluation framework, rather than as a fully implemented and validated control architecture.

Figure 4.3.4 is planned to illustrate the structure of the complete control system. The system will process stator current and voltage inputs, and it will include current PI controllers regulating I_d and I_q , which will generate the U_d and U_q voltage references. These will be passed through the SVPWM module to produce the control signals for the inverter. Park and Clarke transformations will be applied to measure currents and convert them into the $d-q$ reference frame. The MRAS module will be used to estimate rotor speed based on the comparison between the reference and estimated stator current vectors. [4-44]

The objective of this simulation is to validate the capability of the MRAS algorithm in estimating rotor speed without the use of mechanical sensors. Specifically, it aims to evaluate how accurately the estimated speed tracks the actual speed under load disturbance conditions and to observe the dynamic behavior of the system. It is expected that the MRAS estimator will enable rapid convergence of the estimated speed to the reference speed, and that the torque and stator current waveforms will reflect prompt and stable responses when subjected to load changes. These observations will support the feasibility of implementing sensorless vector control in practical PMSM drive systems.

In developing a simulation model for a sensorless vector control system based on PMSM (Permanent Magnet Synchronous Motor), employing a model reference adaptive system (MRAS), the configuration integrates a number of fundamental control and estimation components. These essential building blocks include a coordinate transformation subsystem, a pair of proportional-integral (PI) control loops, an SVPWM (Space Vector Pulse Width Modulation) modulator, and a dynamic rotor speed estimation module structured around the MRAS concept.

A complete illustration of the logical interaction and signal flow among these components is presented in Figure 4.3.4, which outlines the core algorithmic structure used in the sensorless vector control strategy of the PMSM.

Block Descriptions:

- The reference torque current I_{qref} and flux current I_{dref} are subtracted from actual feedback values i_q and i_d respectively.
- The difference is regulated through PI controllers to generate voltage commands U_q and U_d in the synchronous reference frame.
- These voltages are converted into three-phase output signals u_a, u_b, u_c via SVPWM for feeding the motor.
- The motor's phase current measurements i_a, i_b are fed back and translated via Clark and Park transformations into i_q, i_d .
- These are used by the MRAS estimator to determine the rotor's electrical angular velocity $\widehat{\omega_e}$ which is subsequently looped back into the Park block for coordinate synchronization.

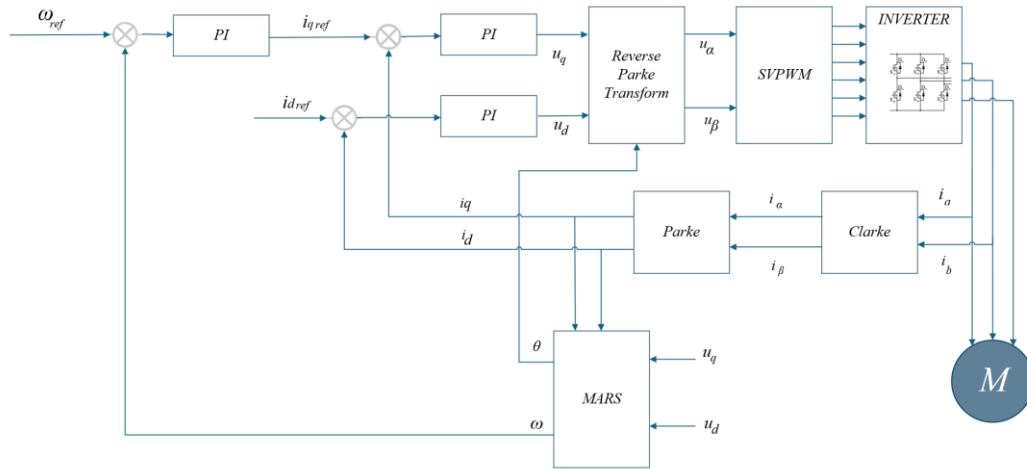


Figure 4.3.4 Block Diagram of MRAS-Based Sensorless Vector Control for PMSM [4-38]

4.4 Phase Lock Loop (PLL)

4.4.1 Position of PLL-Based Hybrid Estimation in PMSM Sensorless Control

Several sensorless control techniques for Permanent Magnet Synchronous Machines (PMSMs) have been investigated to eliminate mechanical position sensors, which add cost, complexity, and potential failure points—particularly undesirable in automotive systems such as Electric Power Steering (EPS). Among the estimation approaches discussed earlier in this thesis, Extended Kalman Filters (EKF), Sliding Mode Observers (SMO), and Model Reference Adaptive Systems (MRAS) represent widely studied model-based

strategies. Each method provides specific advantages: EKF offers statistical optimality under accurate modeling assumptions, SMO provides robustness against parameter variations, and MRAS features relatively simple implementation. However, it is well recognized that no single observer achieves uniformly satisfactory performance across the entire speed range, especially at low or zero speed where back-EMF-based observability deteriorates. [4-55], [4-56], [4-57], [4-58]

In industrial practice, hybrid estimation strategies are often adopted to address this limitation. A commonly reported approach combines Phase-Locked Loop (PLL)-based estimation with High-Frequency Injection (HFI). In such architectures, HFI enhances observability at very low speeds, while PLL-based estimation provides stable phase tracking once sufficient back-EMF is available. [4-49], [4-50], [4-54] This dual-structure approach reflects a practical engineering compromise for wide-range sensorless operation.

It should be emphasized, however, that the present study adopts the Sliding Mode Observer (SMO) as the primary estimation framework for medium-to-high speed operation. The discussion of PLL+HFI in this section is therefore intended to position the broader landscape of sensorless strategies and to clarify the motivation for hybrid estimation concepts, rather than to propose PLL as the principal estimator of this work.

Fundamentals of the PLL + HFI Concept

In hybrid implementations, the operating speed range is divided into low-speed and medium-to-high speed regions. At low speeds, where back-EMF amplitude is insufficient for reliable model-based estimation, HFI is employed. A high-frequency signal is injected into the stator windings, and the resulting current response—modulated by magnetic saliency—is demodulated to extract rotor position information. The extracted signal often appears in a double-angle form and requires polarity correction. [4-49], [4-52], [4-55]

When the motor speed increases and back-EMF becomes sufficiently pronounced, estimation responsibility can transition to a PLL-based structure. The PLL tracks the phase of the back-EMF vector through a closed-loop feedback mechanism that typically includes a proportional–integral (PI) controller. The phase error between measured back-EMF and estimated rotor position is driven toward zero, enabling continuous estimation of rotor speed and electrical angle. [4-49], [4-52], [4-63]

In well-designed hybrid systems, supervisory logic ensures smooth transition between HFI-based and PLL-based estimation. Hysteresis bands and polarity detection mechanisms are typically implemented to prevent abrupt estimator switching or discontinuities. [4-49], [4-54]

Relevance to EPS Applications and Relation to the Present Work

Hybrid PLL+HFI strategies are frequently discussed in the context of EPS because steering systems require position observability even at zero speed, such as during parking or startup.

However, the evaluation results presented in the later Chapter 5 demonstrate that SMO-based sensorless control maintains stable tracking and smooth torque behavior under varying load torque conditions within this operating range. Therefore, while PLL+HFI architectures represent a practical full-range solution in industrial systems as SMO+HFI solution, this study concentrates on establishing the quantitative robustness of SMO-based estimation in its valid operating region. The extension toward hybrid low-speed enhancement remains an identified direction for future work. [4-49], [4-51], [4-55]

Implementation Considerations

While the PLL+HFI concept is frequently discussed as a practical route toward wide-range sensorless operation, it also introduces implementation challenges. High-frequency signal injection may lead to audible noise or vibration and can impose additional filtering and calibration requirements to avoid degradation of driver comfort. Furthermore, the dual-estimator structure requires supervisory logic for region switching and polarity handling; inappropriate threshold selection or transition tuning can introduce discontinuities or instability in the estimated angle. [4-64], [4-69]

From an automotive engineering perspective, hybrid estimators such as PLL+HFI are often considered in the context of achieving low-speed observability and supporting robustness objectives. Some studies report that hybrid approaches can achieve estimation performance comparable to sensor-based feedback under specific test conditions, and such architectures may also be used as part of redundancy strategies in safety-oriented

4.4.2 PMSM Modeling for Phase Lock Loop (PLL)

To develop a robust sensorless control architecture for Permanent Magnet Synchronous Machines (PMSMs), it is essential to begin with a precise mathematical description of the machine's electrical behavior. In this study, the PMSM is modeled under sinusoidal excitation while neglecting magnetic saturation for analytical simplicity.

4.4.2.1 Stator Voltage Equations ($\alpha\beta$ Frame):

Under the $\alpha\beta$ stationary reference frame, the stator voltage dynamics of a surface-mounted PMSM can be described by the following differential vector equation [4-49], [4-61], [4-62]:

$$\mathbf{v}_{\alpha\beta} = R_s \mathbf{i}_{\alpha\beta} + L_s \frac{d\mathbf{i}_{\alpha\beta}}{dt} + \mathbf{e}_{\alpha\beta} \quad (4.4.1)$$

Where:

$\mathbf{v}_{\alpha\beta} = [v_\alpha \ v_\beta]^T$ is the stator voltage vector,

$\mathbf{i}_{\alpha\beta} = [i_\alpha \ i_\beta]^T$ is the stator current vector,

R_s denotes the stator winding resistance,

L_s is the phase stator inductance (assumed equal in both axes for a surface-mounted machine),

$e_{\alpha\beta}$ represents the back-electromotive force (Back-EMF) vector.

The individual components of the back-EMF can be expressed as functions of rotor speed ω_e

rotor flux linkage Ψ_{pm} , and rotor electrical position θ_e :

$$e_\alpha = -\omega_e \Psi_{pm} \sin(\theta_e), \quad (4.4.2)$$

$$e_\beta = \omega_e \Psi_{pm} \cos(\theta_e) \quad (4.4.3)$$

These expressions directly reflect the time-varying process of the rotor flux in the stationary $\alpha\beta$ reference frame and provide a foundation for observer-based angle estimation. The back-EMF vector rotates at the same frequency as the rotor's electrical position, which becomes central to PLL-based estimation, where Ψ_α and Ψ_β represent the instantaneous magnetic flux linkage components in the $\alpha\beta$ coordinate frame, and P denotes the number of pole pairs in the machine. Although the target machine is an IPMSM, the electromagnetic torque equation is expressed in the simplified surface-mounted PMSM form for clarity, since the torque model is not directly involved in the observer derivation. [4-49], [4-61], [4-62]

4.4.2.2 Electromagnetic Torque Production:

In the $\alpha\beta$ frame, the torque developed by the PMSM can be calculated from the cross-product of the flux and current vectors. Additionally, to better describe the estimator's behavior, the torque-generation principle is re-expressed using both $\alpha\beta$ and dq reference frames, for the surface-mounted topology (with no reluctance torque), the expression simplifies to [4-49], [4-61], [4-62]:

$$T_e = \frac{3}{2}P(\Psi_\alpha i_\beta - \Psi_\beta i_\alpha) \quad (4.4.4)$$

Where:

T_e is the instantaneous electromagnetic torque,

P is the number of pole pairs,

$\Psi_{\alpha,\beta}$ are the components of the flux linkage vector.

Since surface-mounted PMSMs exhibit a constant rotor flux magnitude aligned with the d-axis in the rotating frame, the torque equation may also be expressed in the dq frame for completeness, becomes valid under surface-mounted PMSM assumptions with aligned flux linkage [4-62], [4-61]:

$$T_e = \frac{3}{2} P \Psi_{pm} i_q \quad (4.4.5)$$

This alternative formulation is particularly useful within the context of Field Oriented Control (FOC), which aligns the stator current vector such that torque control can be decoupled from flux regulation. This hybrid framework ensures stable and accurate position tracking under various operating conditions and is therefore well-suited for real-time implementation in EPS sensorless application.

4.4.2.3 Supplementary Insight: Relation to Energy Conversion

The torque production mechanism described above reflects the electromechanical energy conversion in PMSMs. The cross-product formulation signifies the tendency of the stator-generated magnetic field to align with the rotor flux - a dynamic interaction that is maximized when the two are orthogonal. Additionally, under steady-state conditions, the power balance in the machine is represented by [4-61]:

$$P_{em} = T_e \cdot \omega_m = v_{\alpha\beta}^T i_{\alpha\beta} \quad (4.4.6)$$

Where:

P_{em} : Electromechanical power (output mechanical power generated by the machine)

T_e : Electromagnetic torque produced by the motor

ω_m : Mechanical angular velocity of the rotor (in rad/s)

$v_{\alpha\beta}^T$: Transpose of Stator current vector in the $\alpha\beta$ reference frame

$v_{\alpha\beta}^T i_{\alpha\beta}$: Inner product of voltage and current vectors $\rightarrow$ represents instantaneous electrical power input

4.4.3 Proposed Hybrid Sensorless Structure

4.4.3.1 Overview of the Hybrid Estimation Scheme

Achieving precise rotor position and speed estimation across the entire speed range of a PMSM — from zero to maximum velocity — remains a major challenge in sensorless control. This section introduces a dual-mode (hybrid) estimation framework that combines two complementary techniques:

High-Frequency Signal Injection (HFI): Effective at standstill and low-speed scenarios.

Phase-Locked Loop (PLL): Reliable in medium to high-speed ranges where back-EMF is significant.

To ensure seamless operation across all speeds, a switching mechanism is implemented between the two modes, along with logic to correct polarity and maintain estimation continuity during reversals or transients.

4.4.3.2 High-Frequency Injection (HFI)

High-frequency signal injection methods leverage either magnetic anisotropy or the saturation-induced saliency of the motor to obtain rotor position data. This is achieved because the position-dependent variation in stator inductance leads to a modulated high-frequency current response, which inherently carries angular information. Since these techniques do not rely on back-EMF, they are particularly effective during standstill and low-speed conditions. To retrieve the relevant position signal, the high-frequency rotating voltage described in equation (4.4.7) is superimposed onto the stator voltage reference generated by the torque control algorithm. [4-49], [4-52]

The injected voltage is expressed as:

$$\mathbf{v}_{HF} = V_{HF} \begin{bmatrix} -\sin(\omega_{HF}t) \\ \cos(\omega_{HF}t) \end{bmatrix} \quad (4.4.7)$$

The applied high-frequency excitation is expressed as a two-dimensional voltage vector in the stationary $\alpha\beta$ frame, defined as:

$$\mathbf{v}_{HF} = \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} \quad (4.4.8)$$

This injected voltage forms the basis of the signal perturbation used to probe rotor-dependent inductance variations. The resulting current response is then processed to extract the embedded rotor position information.

The resulting high-frequency current in the stator contains both positive and negative sequence components [4-49], [4-52]:

$$\mathbf{i}_{HF} = \begin{bmatrix} I_0 \cos(\omega_{HF}t) + I_1 \cos(2\theta_r - \omega_{HF}t) \\ I_0 \sin(\omega_{HF}t) + I_1 \sin(2\theta_r - \omega_{HF}t) \end{bmatrix} \quad (4.4.9)$$

As shown in equation (4.4.9), it is evident that only the negative-sequence component encodes the position information. Therefore, the measured stator currents must undergo specific processing to accurately recover the rotor position (refer to Fig. 4.4.1). The procedure consists of the following steps:

a) The measured three-phase currents are first transformed into the stationary $\alpha\beta$ frame by the Clarke transformation and expressed as the complex current signal $i_{\alpha\beta}(t) = i_\alpha(t) + ji_\beta(t)$.

- b) A band-pass filter centered at ω_{HF} is applied to isolate the injected high-frequency current component and remove the fundamental-frequency content.
- c) The filtered signal is multiplied by $e^{-j\omega_{HF}t}$, (Mixer1) which shifts the positive-sequence component to DC while the negative-sequence component is shifted to twice the injection frequency.
- d) A high-pass filter is then used to reject the DC-level positive-sequence component from the demodulated signal.
- e) Finally, the remaining signal is multiplied by $e^{-j2\omega_{HF}t}$, (Mixer2) and its phase is extracted by the argument operation, implemented as $\text{atan2}(\Im, \Re)$, to obtain a signal proportional to $\cos(2\theta_e)$, $\sin(2\theta_e)$, from which the doubled rotor angle $2\theta_e$ is recovered. [4-49], [4-52], [4-60]

$$\mathbf{i}_{\alpha\beta} = I_{HF} \begin{bmatrix} \cos(2\theta_e) \\ \sin(2\theta_e) \end{bmatrix} \quad (4.4.10)$$

The true rotor angle θ_e is then derived from:

$$\theta_{HFI} = \frac{1}{2} \tan^{-1} \left(\frac{i_\beta}{i_\alpha} \right) \quad (4.4.11)$$

The angle 2θ must be converted back to θ , ensuring that the estimated position is verified for proper polarity. This correction is essential because signal injection methods are inherently unable to differentiate between the magnetic north and south poles of the rotor. To address this constraint, a solution is introduced in this work and described in detail in Section 4.4.3.4

However, this estimation is ambiguous with respect to polarity: both $+\theta$ and $-\theta$ produce the same 2θ .

Therefore, a mechanism is required to resolve the initial angle sign, as discussed later.

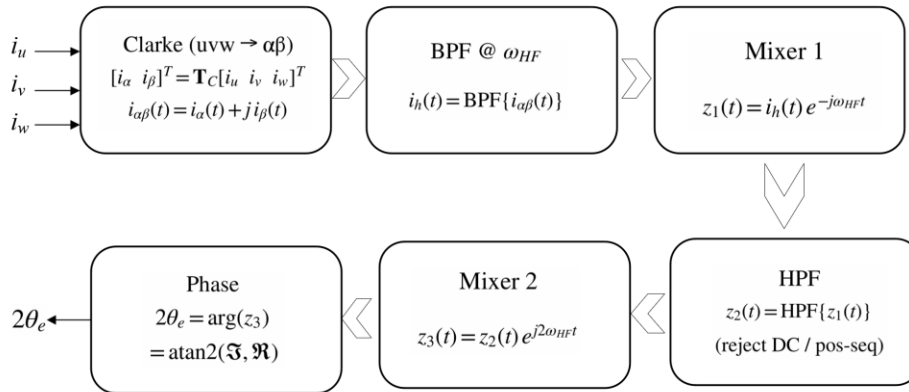


Figure 4.4.1 Heterodyne Demodulation Structure for HFI-Based Rotor Position Estimation

The appearance of the doubled angle term $2\theta_e$ is a direct consequence of the magnetic saliency of the IPMSM. Because the inductance distribution is symmetric every 180° , the high-frequency current response becomes periodic in $2\theta_e$, which leads to the well-known π ambiguity in HFI-based rotor position estimation.

4.4.3.3 Phase-Locked Loop (PLL) Estimator

For speeds where back-EMF becomes appreciable, a PLL-based estimator offers a computationally efficient method for tracking rotor position. Phase-Locked Loop (PLL) techniques have long been employed for synchronizing power converters with electrical grids. Nonetheless, PLL frameworks can also serve as effective tools for estimating rotor speed and position in a sensorless manner for electric machines. In the hybrid sensorless control scheme proposed here, the PLL configuration shown in Figure 4.4.2 is implemented to process the back-EMF signals, which are provided as input variables. Based on equations (4.4.2) and (4.4.3), the corresponding estimated back-EMF can be formulated as the α and β components of the estimated back-EMF are given by [4-49], [4-55]:

$$\hat{e}_\alpha = -\Psi_{pm}\hat{\omega}_e \sin(\hat{\theta}_e), \quad (4.4.12)$$

$$\hat{e}_\beta = \Psi_{pm}\hat{\omega}_e \cos(\hat{\theta}_e) \quad (4.4.13)$$

By comparing the actual estimated back-EMF components with their projections along the estimated angle, an angular error signal can be formed:

$$\hat{\theta}_e = \hat{\theta}_r + \theta_{error}. \quad (4.4.14)$$

$$\Delta_e = -\hat{e}_\alpha \cos(\hat{\theta}_r) - \hat{e}_\beta \sin(\hat{\theta}_r) = \Psi_{pm}\hat{\omega}_e \theta_{error} \quad (4.4.15)$$

This error is regulated by a Proportional-Integral (PI) controller:

$$\hat{\omega}_e = K_p \Delta_e + K_i \int \Delta_e dt \quad (4.4.16)$$

which feeds an integrator to reconstruct the estimated angle $\hat{\theta}_e$. In traditional approaches, this mechanism is sensitive to speed sign reversals, as the PI gains are often tuned for a single direction. To address this, we introduce speed-aware gain scheduling. The controller dynamically adjusts K_p and K_i depending on the sign of the estimated speed, ensuring consistent PLL convergence regardless of direction.

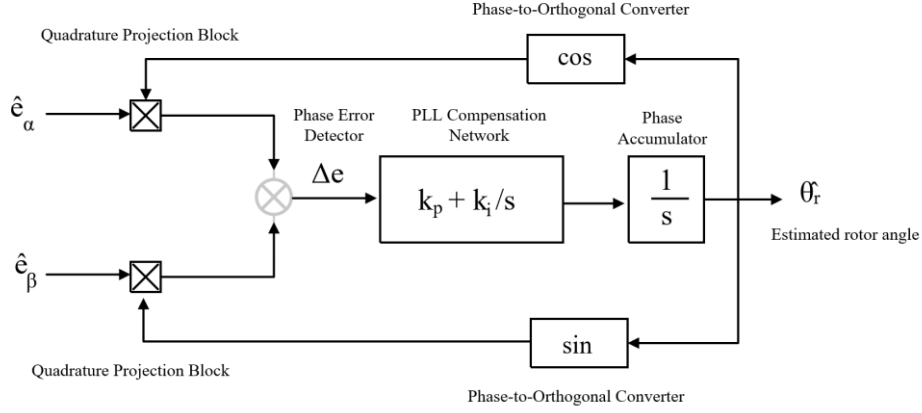


Figure 4.4.2 HFI-Based Synchronous Reference Frame PLL

where $\hat{\theta}_e$ and $\hat{\omega}_e$ denote the electrical rotor position and angular speed, respectively (both of which are incorporated into the estimated back-EMF), $\hat{\theta}_r$ refers to the inferred rotor angle, and θ_{error} represents the angular deviation between reference and estimate. A Proportional-Integral (PI) controller is employed to minimize this deviation, effectively driving the estimated angle toward convergence with the true value. Sensorless strategies based on conventional PLL configurations typically exhibit performance degradation during direction reversal, due to the following known challenges: The PI controller parameters K_p and K_i initially tuned for positive-speed operation, do not remain valid when the motor rotates in reverse. As a result, gain adjustments must be made for each directional mode, a process that can become highly intricate. [4-49], [4-52], [4-63]

Moreover, when the speed is negative, the angle estimate suffers from a 180-degree offset. To counteract these limitations, a modified PLL architecture has been developed, in which the PI controller gains are automatically adjusted depending on the sign of the estimated speed. When the rotor speed changes sign, the polarity of the back-EMF vector reverses. Since the PLL tracks the phase of this vector, the estimated angle may shift by π radians, which motivates the polarity correction mechanism introduced in the hybrid estimator.

This allows the system to remain effective across the entire speed spectrum. When a reversal in direction occurs, the polarity of the back-EMF changes, prompting the PLL to lock to a new angle offset of π radians. Therefore, an explicit angle offset correction mechanism has been incorporated into the proposed hybrid estimation system to enhance its reliability under reversal conditions. [4-53], [4-49]

This PLL reconfiguration is executed in a way that introduces no discontinuity into the ongoing angle or speed estimation process. This seamless behavior is ensured because the rotor position estimation is

temporarily managed by the HFI subsystem during the reconfiguration phase. A detailed explanation of the transition logic is provided in the following subsection.

4.4.3.4 Transition Logic Between PLL and HFI

Figure 4.4.3 outlines the architecture of the proposed hybrid rotor angle and speed estimation mechanism. Each time high-frequency signal injection is initiated, it becomes critical to verify the initial polarity of the rotor position. This check ensures that the double-angle estimate from HFI is accurately converted back to the fundamental rotor angle. If the polarity is not correctly resolved, the estimated position may exhibit a phase shift of π radians from the actual rotor location.

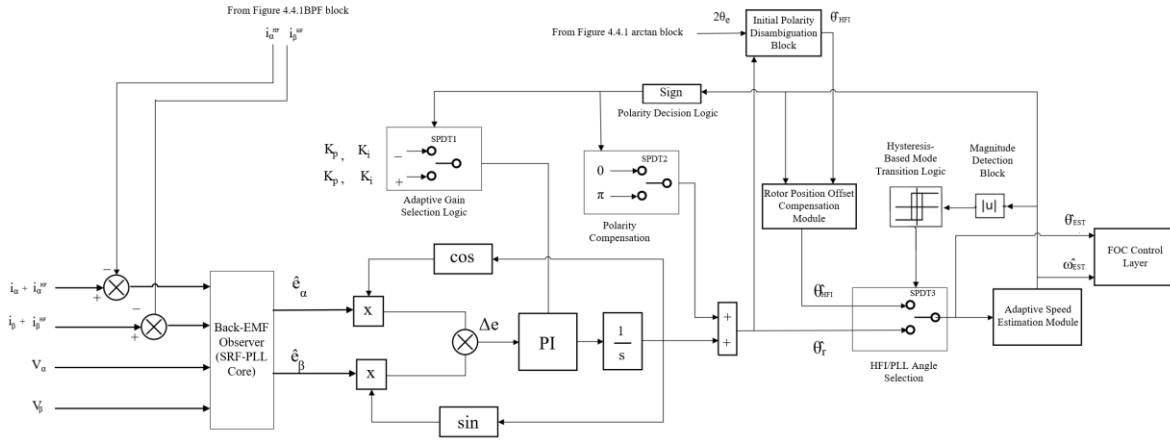


Figure 4.4.3 Integrated Hybrid Rotor Angle and Speed Estimation Architecture Based on HFI and PLL

To resolve the inherent π -radian ambiguity of HFI-based angle estimation, the proposed hybrid estimator introduces a polarity decision logic that determines whether an additional π offset should be applied to the HFI-derived angle. As expressed in (4.4.17), the polarity decision is obtained by referencing the PLL-based angle estimate in the speed range where back-EMF is sufficiently observable. In this overlap region, the PLL estimate provides a reliable absolute phase reference, allowing the controller to decide whether the HFI estimate must be shifted by π so that both estimators share a consistent electrical angle convention.

$$\text{Offset} = \begin{cases} 0, & \text{if } \theta_{PLL} \in \frac{\pi}{2} \pm \text{HystBand} \\ \pi, & \text{if } \theta_{PLL} \in \frac{3\pi}{2} \pm \text{HystBand} \end{cases} \quad (4.4.17)$$

To avoid incorrect polarity assignment caused by transient mismatch, noise, or electromagnetic interference, the polarity decision is not made using a single threshold. Instead, a hysteresis band is applied so that the decision changes state only when the PLL–HFI discrepancy (or the selected decision variable) crosses a pair

of separated thresholds and remains beyond them for a defined persistence interval. This prevents rapid toggling near the boundary and ensures stable estimator behavior during switching and reversals.

In addition, because the HFI path typically includes demodulation and filtering, it introduces a phase delay that appears as an effective angle lag in the FOC transformations. To compensate for this delay, a speed-dependent correction map (lookup table) can be applied to the HFI-derived angle, calibrated offline to account for the filter-induced phase shift at each operating point.

The overall estimator handover is therefore performed with an intentional overlap region: HFI-based estimation remains active until the PLL-based estimator is demonstrably stable, and only then is injection disabled. This prevents loss of observability at low speed and avoids discontinuities in the control angle used for Park and inverse Park transformations. In practice, the switching speed is implemented with hysteresis to avoid mode “chatter,” and the transition is designed to maintain continuous θ_e and bounded torque transients. [4-49], [4-50], [4-54]

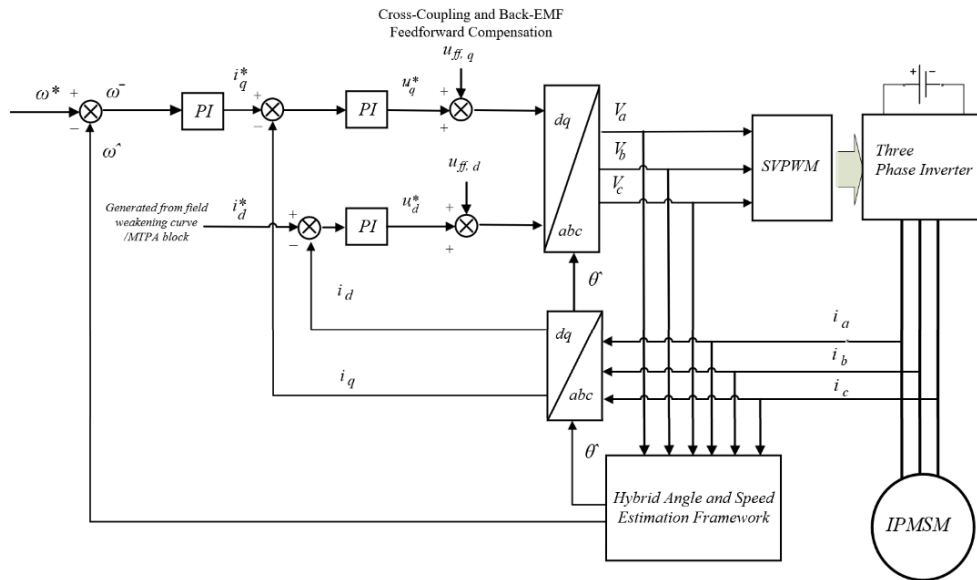


Figure 4.4.4 Implementation-Oriented Block Diagram of Sensorless IPMSM FOC with Hybrid HFI-PLL Estimation

This technique ensures that the estimator switch occurs smoothly across the entire range of operating speeds and acceleration rates. With this structure, consistency and continuity are preserved during estimator handovers.

Once both $\widehat{\theta}_{HFI}$ and $\widehat{\theta}_{PLL}$ have been determined, the system proceeds with a transition between estimators, as shown in figure 4.4.3. Several methods may be employed for this purpose; among them, applying weighting coefficients is one of the most prevalent approaches.

In this work, a speed-sensitive transition switch equipped with a hysteresis band is proposed to mitigate potential signal degradation near the crossover point. This configuration ensures reliable estimator operation and preserves estimation fidelity during the transition.

The final estimated rotor angle $\widehat{\theta}_{EST}$ is then used to compute $\widehat{\omega}_{EST}$, the estimated rotor speed. Notably, during PLL operation, this speed value is not obtained directly from the PLL itself.

This is due to the fact that when both HFI and PLL estimators are active, high-frequency signal components may be present in the estimated speed, contaminating the output. To address this, the PLL integrator inherently filters out the high-frequency portion of the signal, taking advantage of the fact that these components have zero average value.

Consequently, the need for additional external filtering is reduced, lowering the computational burden of the overall sensorless estimation process. A further refinement used to minimize reliance on external filters involves directly subtracting the injected high-frequency current components from the inputs to the PLL observer (see figure 4.4.3). [4-49], [4-60]

This approach effectively diminishes the adverse influence of injected signals on the back-EMF estimation, leading to a more accurate and stable estimator output during hybrid operation.

Chapter 5 Discussion and Conclusion – Ideal and Optimal Strategy for EPS Applications

This chapter presents the final integrated evaluation of rotor position estimation strategies for Electric Power Steering (EPS) systems in sedan-class vehicles. By comparatively examining representative estimation methods under EPS-specific operational constraints, the study identifies a hybrid SMO + HFI architecture as the most practical solution for sensorless EPS implementation.

5.1 Importance of Sensorless Control in Sedan-Class EPS

Electric Power Steering (EPS) systems have become the dominant steering technology in modern automotive platforms due to their superior energy efficiency, controllability, and compatibility with advanced driver-assistance systems. In cost-sensitive sedan platforms, reduction of bill of materials and simplification of hardware architecture are primary design objectives [5-16].

The elimination of rotor position sensors such as resolvers or Hall-effect sensors reduces wiring complexity, component cost, and potential failure points. Sensorless control strategies estimate rotor position using electrical signals, enabling cost reduction while maintaining torque precision.

However, EPS differs fundamentally from traction drives [5-1]. Steering systems must operate with high precision at very low speed and even at standstill. Because back-EMF approaches zero at low speed, pure back-EMF-based estimators are insufficient, necessitating hybrid solutions [5-7], [5-10], [5-22], [5-24], [5-25].

5.2 Technical Comparison of EKF, SMO, MRAS, and PLL

The following comparative evaluation assesses each observer with respect to: (1) full-speed capability, (2) robustness, (3) computational load, (4) implementation complexity, and (5) suitability for mass production.

Criterion	SMO	PLL	EKF	MRAS
Speed Range Suitability	Medium–High (requires HFI for low)	Medium–High (requires HFI for low)	Medium–High	Medium–High
Low-Speed Robustness	Moderate (with filtering)	Low	Moderate	Low
Estimation Accuracy	High	High (high speed)	Very High	High (steady state)

Computational Load	Low	Low	Very High	Medium
Implementation Complexity	Low	Medium	Very High	High
Adaptability ^{*1}	Fixed gain (robust)	Low	High (model dependent)	Adaptive
Determinism (MCU)	High	High	Lower	Moderate
Mass-Production Suitability	Best Tradeoff	Secondary Option	Limited	Rarely Adopted

^{*1}Adaptability refers to the ability of the estimator to maintain accurate position estimation under parameter variations such as resistance, inductance, temperature, and load changes.

Table 5.1 Comparative Overview of Sensorless Estimation Methods for PMSM [5-3], [5-5], [5-11], [5-13], [5-26]

5.2.1 Core Reasons for Selecting SMO + HFI

Technical Superiority: HFI ensures zero-speed observability, while SMO provides robust and computationally efficient estimation in the medium-to-high speed region.

Cost Efficiency: Eliminating rotor sensors reduces hardware cost, harness complexity, and long-term reliability risks.

Industrial Validation: Hybrid observer architectures combining saliency-based low-speed estimation with back-EMF-based observers have been demonstrated in academic literature and industrial disclosures [5-2].

5.3 Technical Comparison of Rotor Position Estimation Methods (Expanded)

This section expands the technical comparison of EKF, SMO, MRAS, and PLL under EPS-specific constraints: real-time determinism, robustness to parameter drift, steering feel sensitivity, and low-speed/standstill operation where back-EMF observability is limited.

5.3.1 Extended Kalman Filter (EKF) – Expanded Assessment

Principle. The Extended Kalman Filter (EKF) is a recursive Bayesian estimator for nonlinear systems. It linearizes the system dynamics and measurement model around the current estimate (Jacobian-based linearization) and updates state estimates using predicted covariance and measurement innovation.

Advantages. EKF can achieve very high estimation accuracy when (i) the plant model is sufficiently accurate, (ii) measurement noise statistics are representative, and (iii) initialization is robust. It can simultaneously estimate rotor position, speed, and selected parameters (e.g., R_s or flux linkage) in a unified framework.

Limitations. EKF typically requires Jacobian evaluation and matrix operations each control cycle. This creates high computational demand and may degrade determinism on cost-constrained EPS ECUs. Performance is strongly model-dependent; under rapid transients, modeling mismatch, or poor initialization, the filter may diverge.

Suitability for EPS. In sedan-class EPS, EKF is often considered technically capable but economically and operationally expensive. The combined requirement of strict timing (hard real-time), broad operating temperature, and high robustness under vehicle disturbances makes EKF less attractive for mass production [5-11], [5-12].

5.3.2 Sliding Mode Observer (SMO) – Expanded Assessment

Principle. A Sliding Mode Observer (SMO) is a nonlinear observer using discontinuous injection to drive estimation error to a sliding surface. In PMSM/IPMSM sensorless control, SMO is commonly used to reconstruct back-EMF in the stationary α - β frame, from which rotor position is obtained via phase extraction [5-3], [5-4].

Advantages. (1) Robustness: SMO provides strong robustness against bounded parameter variations (e.g., R_s drift), disturbances, and modeling error. (2) Computational efficiency: SMO implementations avoid matrix inversion and are well suited to mid-tier automotive MCUs. (3) Fast convergence: discontinuous correction yields rapid error suppression. These properties map well to EPS requirements (robust steering feel, deterministic execution) [5-21].

Limitations. The primary limitation is chattering introduced by discontinuous switching. Practical implementations use smoothing (boundary layer/saturation functions), low-pass filtering, or higher-order sliding-mode methods to reduce chattering without sacrificing convergence. Noise sensitivity and transient behavior require careful gain tuning and filter design.

Suitability for EPS. For the medium-to-high speed region where back-EMF is observable, SMO provides a favorable robustness-to-complexity tradeoff, and it integrates naturally into standard FOC architectures used in EPS. This makes SMO a strong primary estimator candidate in mass-production sedan EPS programs.

5.3.3 Model Reference Adaptive System (MRAS) – Expanded Assessment

Principle. MRAS uses two parallel models: a reference model that does not depend on the quantity to be estimated and an adjustable model that does. An adaptive law updates the estimated parameter (speed/position) to minimize the error between model outputs.

Advantages. MRAS offers adaptive behavior and can achieve high steady-state accuracy when convergence is reached, particularly under relatively stable operating conditions. It can compensate for some parameter drift via the adaptive mechanism.

Limitations. Convergence speed is the central concern. EPS frequently experiences rapid torque reversals and highly dynamic disturbances, which can interrupt MRAS convergence or require conservative gains that slow adaptation. MRAS design and gain tuning can become complex, and low-speed performance still suffers without an injection-based complement.

Suitability for EPS. MRAS is valuable academically and can be effective in controlled regimes, but it is less commonly selected for mass-produced sedan EPS due to tuning complexity and convergence sensitivity under dynamic steering conditions [5-12], [5-26].

5.3.4 Phase-Locked Loop (PLL) – Expanded Assessment

Principle. PLL-based estimators track rotor electrical angle by locking to the phase of a reconstructed signal (often back-EMF or estimated flux). In sensorless PMSM drives, the PLL minimizes phase error in a feedback structure, yielding smooth position and speed estimates at sufficiently high signal-to-noise ratio.

Advantages. PLL is computationally light, provides smooth steady-state estimation, and is well understood in control practice. For medium-to-high speed operation where back-EMF amplitude is high, PLL can achieve high accuracy and stable tracking.

Limitations. PLL performance is strongly speed dependent. At low speed, back-EMF is weak and phase extraction becomes unreliable. Filters used to suppress noise introduce delay and can degrade steering feel. Therefore, PLL alone cannot satisfy full-speed EPS requirements and is typically paired with a low-speed method such as HFI.

Suitability for EPS. PLL is a viable secondary estimator in hybrid sensorless EPS architectures, especially for higher-speed steering scenarios. However, it is rarely an optimal single estimator across the full EPS operating envelope [5-5], [5-13].

5.3.5 Summary Table – Comparative Analysis (Enhanced). Table 5.2 provides an EPS-oriented summary. The central conclusion is that, among back-EMF-based methods, SMO provides the best robustness-to-complexity tradeoff for mass-produced sedan EPS, while low-speed operation requires HFI regardless of the chosen back-EMF estimator.

Criterion	SMO	PLL	EKF	MRAS
Full-speed feasibility (standstill) [5-8]	Needs HFI at low/0 speed	Needs HFI at low/0 speed	Needs HFI at low/0 speed	Needs HFI at low/0 speed
Robustness to parameter drift	High (bounded mismatch tolerant)	Medium (signal-quality dependent)	High but model/noise dependent	Medium–High after convergence
Computational load (MCU)	Low	Low	Very High	Medium
Tuning burden	Moderate (gains + filtering)	Moderate (loop gains + filtering)	High (covariance + init)	High (adaptive law tuning)
Sedan EPS suitability	Best tradeoff in mid–high speed	Secondary estimator	Overhead too high for many	Rarely adopted in mass production

Table 5.2 Detailed Comparative Analysis of Sensorless Estimation Methods under EPS Constraints

provides an EPS-oriented summary. The central conclusion is that, among back-EMF-based methods, SMO provides the best robustness-to-complexity tradeoff for mass-produced sedan EPS, while low-speed operation requires HFI regardless of the chosen back-EMF estimator. [5-3], [5-8], [5-11], [5-12], [5-13], [5-23], [5-25]

5.3.6 Conclusion of Comparative Evaluation (Reinforced)

From a technical standpoint, SMO offers the best tradeoff between estimation performance, implementation cost, and robustness for sensorless EPS systems in sedan-class vehicles, particularly in the medium-to-high speed region where back-EMF is observable.

PLL remains a solid candidate for high-speed scenarios with smooth steady-state tracking, but its low-speed limitations necessitate hybridization.

EKF provides high theoretical estimation performance but faces practical constraints in computational burden, model dependence, and real-time determinism.

MRAS is academically valuable and can perform well after convergence, yet may be disadvantaged in EPS due to convergence sensitivity and tuning complexity.

In practice, many industrial sensorless EPS concepts adopt a hybrid approach: a robust back-EMF observer (often SMO or PLL) for medium–high speed, combined with HFI-based saliency tracking for low-speed and standstill operation [5-1]. This hybrid logic forms the foundation for the final selection in Section 5.5.3.

5.4 Cost Comparison: Sensor vs. Sensorless Architectures

This section evaluates the cost implications of sensor elimination in EPS. It preserves the original cost drivers and extends the discussion with explicit assumptions, sensitivity notes, and an ROI framing appropriate for a thesis discussion [5-28].

5.4.1 Overview of Cost Drivers in EPS Design

EPS system cost is influenced by both hardware and engineering activities. Primary cost drivers include rotor position sensing components (resolver, Hall, or TMR), wiring harness and connector count, PCB area and analog front-end components, development and calibration activities, manufacturing complexity, and warranty exposure from failure modes [5-29].

5.4.2 Cost Breakdown: Sensor-Based vs. Sensorless (SMO + HFI)

Table 5.3 summarizes per-unit hardware contributors and engineering activities. Values represent typical order-of-magnitude ranges used for program-level comparisons; actual values depend on supplier and platform. The key structural point is that sensorless architectures convert part cost into software development effort [5-28], [5-29].

Cost Element	Resolver-Based	Hall/TMR-Based	Sensorless (SMO + HFI)
Rotor position sensor module	~\$6–10 per unit (typical)	~\$2–3 per unit (typical)	\$0 (removed)
Harness & connector adders	+\$2–4 (shielding/sealing)	+\$1–2 (simpler)	\$0–\$1 (reduced)
Sensor interface / PCB area	Higher (AFE + routing)	Moderate	Not required (reuse existing ADC)
Development & calibration effort	Sensor calibration + offset maps	Sensor calibration	Algorithm tuning + validation

Table 5.3 Cost Comparison of Sensor-Based and Sensorless EPS Architectures

5.4.3 Economic Impact of Removing Rotor Sensors

Removing rotor position sensors reduces direct BoM and associated manufacturing complexity. In addition to the sensor itself, harness routing and sealing requirements are reduced, and the set of production-line sensor calibration steps can be simplified.

A reasonable conservative estimate for combined per-unit savings from sensor removal (sensor + harness + assembly simplification) is approximately \$8–14 per vehicle for high-volume sedan programs. [5-19]

At scale, the savings are material. For example, a program volume of 1 million units over lifecycle implies an incremental savings of roughly \$8–14 million (excluding secondary benefits such as improved reliability and reduced warranty cost).

5.4.4 Real-World Case Studies (Illustrative)

JTEKT (single-shunt / current sensing simplification). Supplier technical disclosures describe architectures that reduce sensing and hardware footprint, consistent with the broader industry trend toward cost-optimized motor control in EPS.

DENSO (EPS-MCU integration and sensor reduction roadmap). Public communications describe EPS motor control units that reduce size and cost and indicate continued integration and sensor reduction [5-17].

Academic and open literature note that sensor elimination can be economically justified when platform reuse of software is feasible and when reliability improvement offsets the incremental calibration/validation effort.

5.4.5 Cost Recovery and ROI Perspective

Sensorless SMO + HFI requires additional engineering and validation effort (e.g., estimator tuning, injection parameter tuning, low-speed steering feel evaluation, and fault diagnostics). This is front-loaded cost. However, it amortizes across platforms as the algorithm and calibration workflows mature.

In practice, the ROI depends on: (i) platform reuse rate, (ii) target volume, (iii) sensor cost baseline, and (iv) additional validation scope required by the OEM safety process. In high-volume sedans, even conservative BoM savings typically dominate the incremental software effort over program lifecycle

5.5 Market Adoption Evidence: Series Production Installations, Supplier & OEM Trends

This section organizes market adoption evidence into three categories: It preserves the original claims, but clarifies the nature of evidence: (i) patents and technical disclosures, (ii) production-facing product communications, and (iii) academic validations that are consistent with industrial architectures.

5.5.1 Global EPS Penetration and Context

EPS is broadly adopted across passenger vehicles, and increasing integration with ADAS amplifies demand for precise and robust motor control. In this context, sensor reduction is driven by cost, packaging, and reliability pressures in mass-produced vehicles [5-16], [5-19]

5.5.2 Tier-1 Supplier and Patent Indicators

Table 5.4 classifies industrial evidence without overstating claims. Patents and supplier communications are used as indicators of architectural direction (sensor reduction and hybrid estimation), while academic papers provide detailed algorithmic validation consistent with these directions.

Supplier	Evidence Type	Observed Direction	Citation
JTEKT (Japan)	Patent / technical disclosure	Resolver reduction, hybrid observer concepts	[5-20]
DENSO (Japan)	Product communication	Integrated EPS-MCU, sensor reduction roadmap	[5-17]
NSK (Japan)	Technical journal / disclosure	Sensor-reduced EPS approaches	[5-30], [5-31]
Bosch / ZF (EU)	Patent activity (general)	Hybrid sensorless methods reported in filings	[5-32], [5-33]
Nexteer (NA)	Architecture trends (public)	Modular EPS control; observer fallback patterns	[5-34], [5-35]

Table 5.4 Industry Evidence from Supplier Disclosures and Patent Trends

5.5.3 OEM Case Studies: Production Sedan Platforms (Illustrative)

Table 5.5 is intentionally labeled illustrative: it describes directional adoption rather than claiming confidential production details. The academically defensible core is that hybrid sensorless concepts are

consistent with Tier-1 disclosure patterns and are supported by peer-reviewed validations of the same class of algorithms.

OEM / Region	Representative Model(s)	Likely EPS Supplier	Sensor Strategy Trend	Notes (evidence type)
Japan	Compact / C-segment sedans	JTEKT / NSK / DENSO	Sensor reduction; hybrid sensorless direction	Patents + supplier disclosures
Korea	High-volume sedans	DENSO / local Tier-1	Integration and sensor simplification	Public communications
North America	Sedans / crossovers	Nexteer	Observer-based EPS control and modular system architectures	Industry product communication
Europe	Sedans / hatchbacks	Bosch / ZF	Patent activity suggests hybrid sensorless interest	Patent indicator
China	High-volume NEV platforms	Domestic Tier-1	Aggressive cost reduction; sensorless interest	Market trend + publications

Table 5.5 Illustrative Market Trends of Sensorless EPS Architectures across Regions

5.5.4 Patent and Publication Trends

A recurring theme in the literature is full-speed sensorless operation achieved by combining low-speed saliency methods (HFI) with a medium–high-speed back-EMF observer (often SMO or PLL). Hybrid observers validated across wide speed ranges demonstrate that this architecture is feasible and competitive.

This pattern aligns with industrial objectives: removing the rotor position sensor without sacrificing standstill performance. In mass-produced EPS, this translates directly into BoM reduction while maintaining steering feel.

5.5.5 Market Readiness Indicators (Toolchain and MCU Support)

Industrial readiness is also reflected in the availability of practical implementation guidance and toolchain support for SMO-based sensorless control and HFI-based saliency tracking on automotive-relevant microcontroller platforms. This reduces implementation risk and accelerates platform reuse [5-18], [5-36].

5.5.6 Conclusion of Market Adoption Evidence

Market evidence indicates that hybrid sensorless strategies (SMO/PLL for medium–high speed combined with HFI for low speed) are aligned with Tier-1 cost and packaging objectives. Academic literature demonstrates the feasibility of wide-speed-range hybrid operation, while the present work structurally validates the medium-to-high-speed SMO region and establishes architectural readiness for low-speed HFI integration. Therefore, the industrial adoption risk of such hybrid architectures is reduced relative to less mature alternatives.

5.6 Development Complexity and Cost: Implementation Challenges, Parameter Tuning, and Safety Compliance

The superiority of SMO + HFI in EPS must be supported not only by estimation performance, but also by a realistic account of implementation challenges, tuning workload, and safety compliance. This section expands these aspects without reducing the original content.

5.6.1 Implementation Challenges in SMO + HFI Structures

Motor model precision and parameter sensitivity. In practice, SMO performance depends on reasonable fidelity of key motor parameters such as stator resistance R_s and inductances L_d/L_q . These parameters vary with temperature, manufacturing tolerance, and aging. For EPS applications, robustness against these drifts is essential because ambient and winding temperatures can vary widely across vehicle conditions [5-6].

Nevertheless, sliding-mode structures are well known to provide disturbance rejection and bounded-mismatch robustness, and practical SMO designs commonly introduce gain scheduling and filtering to maintain performance across the operating envelope.

Chattering and filtering requirements. Discontinuous injection in SMO can cause chattering, which may manifest as current ripple, torque ripple, and acoustic noise. In EPS, these artifacts can degrade steering feel. Mitigation approaches include boundary-layer approximations (saturation instead of pure sign), low-pass filtering, and higher-order sliding-mode observers such as super-twisting designs [5-6], [5-21].

HFI design and NVH constraints. HFI introduces high-frequency excitation that can couple into acoustic and structural modes. Consequently, frequency, amplitude, and demodulation design must be selected to meet NVH targets, particularly for premium sedan platforms. Practical literature demonstrates that careful injection design and, where applicable, randomization or multi-frequency approaches can reduce audible artifacts while preserving saliency tracking [5-9], [5-14], [5-15].

Challenge	EPS Impact	Typical Mitigations
Rs / Ld / Lq drift	Angle bias; torque ripple; degraded convergence	Gain scheduling; adaptive elements; temperature compensation
SMO chattering	Current ripple; steering feel degradation	Saturation boundary layer; LPF; super-twisting SMO
Measurement noise / EMI	Phase noise; estimator instability	Signal conditioning; robust filtering; observer tuning
HFI audible noise (NVH)	Customer-perceived acoustic noise	Random/multi-frequency injection; amplitude optimization
Mode switching (HFI↔SMO)	Torque discontinuity; transient artifacts	Hysteresis; blending; phase continuity checks

Table 5.6. Implementation Challenges and Mitigation Methods (EPS-Oriented) [5-6], [5-14], [5-15], [5-21]

5.6.2 Parameter Tuning and Development Workload

SMO gain scheduling. A single set of SMO gains rarely achieves optimal performance across the entire speed and load range. At high speed, excessive gains can amplify noise; at low speed, insufficient gains can slow convergence. Therefore, practical implementations often apply gain scheduling as a function of electrical speed, current magnitude, or estimated back-EMF amplitude.

HFI parameterization. HFI requires selection of injection waveform, frequency, amplitude, and demodulation filtering. These choices must balance observability and NVH constraints. Furthermore, the saliency index varies with operating point, so robust demodulation design is necessary to avoid false angle estimates [5-10].

Transition logic. The transition from HFI-based low-speed estimation to SMO-based medium–high-speed estimation typically uses a threshold and hysteresis window (e.g., 200–400 RPM, motor dependent). Overlap and blending reduce mode-hopping and torque transients. Hybrid observer literature demonstrates practical approaches for whole-speed operation [5-2].

Subsystem	Key Parameters	Primary Objective	Risk if Mis-tuned
SMO	Injection gain, saturation width, LPF cutoff	Fast convergence with low ripple	Chattering or sluggish response

Back-EMF extraction	Filter bandwidth, phase compensation	Stable phase estimate	Phase delay; torque ripple
HFI injection	Frequency, amplitude, waveform	Standstill observability with NVH control	Audible noise; weak saliency response
HFI demodulation	Band-pass/LPF params, synchronous detection	Noise-robust angle extraction	Angle jitter; false lock
Switching logic	Threshold, hysteresis, blending ratio	Smooth transition	Mode hopping; discontinuity
Diagnostics	Angle mismatch threshold, plausibility timers	Fault detection	False DTC or missed fault
Thermal adaptation	Rs update law, compensation curves	Robustness across temperature	Temperature-dependent bias

Table 5.7. Key Tuning Parameters for SMO + HFI-Based Sensorless Control

The tuning list above makes explicit that sensorless SMO + HFI transfers part of the system complexity from hardware to software. This is acceptable in high-volume EPS because tuning processes can be standardized and reused across platforms.

5.6.3 Functional Safety (ISO 26262) Compliance Considerations

Redundancy and fault detection. A hybrid SMO + HFI architecture naturally supports two independent estimation pathways: (i) back-EMF-based estimation and (ii) saliency-based estimation. Their cross-check provides a basis for fault detection without requiring a physical rotor sensor, which can support ASIL-B or ASIL-C strategies when integrated with proper diagnostics [5-14], [5-27].

Diagnostic concepts. Practical diagnostics include: (a) angle mismatch thresholds between SMO and HFI outputs, (b) rate-of-change limits, (c) torque/current plausibility checks, and (d) injection integrity checks. Fault handling commonly transitions to a torque-limited fallback mode while maintaining driver assist within safe bounds.

Real-time determinism. Safety architectures require bounded response times. SMO implementations are computationally predictable, which supports deterministic execution time analysis compared to matrix-heavy methods.

Safety Mechanism	Observable Signal	Typical Criterion	EPS Benefit
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SMO vs HFI angle cross-check	$\theta_{\text{SMO}}, \theta_{\text{HFI}}$	$ \theta_{\text{SMO}} - \theta_{\text{HFI}} > \text{threshold}$	Detect estimator drift/failure
Rate-of-change plausibility	$d\theta/dt, \omega$	Exceeds physical limits	Detect false-lock or noise
Torque/current plausibility	i_q , commanded torque	Mismatch beyond limit	Detect actuator/estimator inconsistency
Injection integrity check	HF current response	Loss of saliency response	Detect HFI path failure
Mode-switch continuity check	θ continuity	$\Delta\theta$ jump beyond limit	Prevent torque shock

Table 5.8. Functional Safety Mechanisms Enabled by Hybrid Sensorless Estimation [5-14], [5-27]

5.6.4 Sensorless vs Sensor-Based Development Burden

Table 5.9 Overall, SMO + HFI increases early-stage software design and calibration burden but provides long-term benefits in hardware simplification, platform reuse, and the ability to implement software-based redundancy and diagnostics. This aligns well with high-volume sedan EPS programs.

Development Aspect	Sensor-Based (Resolver/Hall/TMR)	Sensorless (SMO + HFI)
Hardware BOM	Higher (sensor + harness)	Lower (sensor removed)
Production calibration	Sensor alignment/offset calibration	Algorithm calibration (software)
ECU compute demand	Lower	Moderate (still feasible)
Validation scope	Sensor failure modes + control	Observer/HFI failure modes + control
NVH considerations	Sensorless not applicable	HFI audible noise must be controlled
Reusability across platforms	Limited by sensor package differences	High (software reuse)
Field reliability	Sensor can be failure point	Fewer mechanical parts; software diagnostics
Safety strategy	Hardware redundancy often used	Software redundancy via dual estimates

Table 5.9 Comparison between Sensor-Based and Sensorless EPS Development Approaches

5.7 Conclusions and Future Outlook: Overall Superiority of SMO–HFI Hybrid Sensorless EPS

5.7.1 Synthesis of Key Findings

This chapter compared four observer families—EKF, SMO, MRAS, and PLL—under EPS-specific constraints (standstill capability, robustness to parameter drift, real-time determinism on automotive ECUs, steering feel/NVH, mass-production scalability, and functional safety). The synthesis yields a consistent outcome: a hybrid architecture combining High-Frequency Injection (HFI) at zero/low speed with a back-EMF observer in the medium–high speed region is required for full-envelope EPS operation. Among back-EMF observers, SMO provides the most favorable robustness-to-complexity tradeoff for sedan-class production programs [5-7], [5-22].

The core technical reason is observability. Back-EMF magnitude scales with electrical speed, so observability degrades at very low speed and at standstill; therefore, any back-EMF-based estimator (SMO/PLL/EKF/MRAS) requires a saliency-based low-speed method to maintain position estimation performance. HFI provides this capability by exploiting IPMSM saliency and/or saturation effects to generate a position-dependent high-frequency response independent of speed [5-25].

Within the back-EMF-observable region, SMO is strongly aligned with embedded EPS constraints: it is computationally light, robust to bounded parameter mismatch, and supports deterministic execution timing. Practical SMO designs mitigate chattering through smoothing and higher-order sliding-mode methods such as super-twisting observers, maintaining steering feel while preserving convergence.

EKF remains an academically powerful option due to probabilistic optimality and the ability to estimate multiple states and parameters simultaneously. However, its matrix-heavy computations, model/noise tuning burden, and sensitivity to initialization reduce its practical attractiveness for cost-constrained sedan EPS ECUs. MRAS offers adaptive capability but may be disadvantaged by convergence behavior and tuning sensitivity under highly dynamic steering conditions. PLL provides smooth tracking at higher speed but is limited by low-speed signal-to-noise and filtering delays; therefore, it is best used as a secondary estimator in hybrid designs. [5-22]

In this thesis, EKF and MRAS were not discussed with HFI method, while EKF and MRAS can theoretically be combined with HFI, the integration significantly increases computational complexity. In contrast, SMO and PLL structures allow a more practical hybrid implementation with HFI, which is why they were selected for detailed discussion in this study. In many practical motor control systems, especially in automotive applications, hybrid approaches are commonly used. High-frequency injection is typically applied in the low-speed region, while model-based observers such as SMO or PLL are used at medium and high speeds. The structures discussed in this work reflect these commonly adopted industrial practices.

5.7.2 Strategic Recommendation for Sedan-Class EPS

Based on the integrated evidence, the recommended baseline architecture for mass-produced sedan-class sensorless EPS is [5-8]:

- Motor type: IPMSM with sufficient saliency for HFI-based low-speed estimation.
- Low-speed/standstill (approx. 0–300 RPM, motor dependent): HFI-based saliency tracking with NVH-aware injection design.
- Medium–high speed: SMO-based back-EMF reconstruction (often paired with PLL phase extraction) for robust, low-cost estimation.
- Transition logic: threshold + hysteresis + optional blending to guarantee phase continuity and avoid torque shock.
- Safety concept: dual-estimate cross-check (θ_{SMO} vs θ_{HFI}) plus plausibility diagnostics to support ASIL-oriented design.

5.7.3 Future Technology Directions

Future development directions for sensorless EPS include:

- Improved HFI schemes to reduce audible noise and EMI (e.g., randomized pulse or low-frequency pulse injection) while maintaining observability [5-9].
- Advanced sliding-mode observers (super-twisting / flux sliding-mode observers) to reduce chattering and improve robustness under parameter drift.
- Enhanced mode-transition strategies and sensorless safety monitors, especially for steer-by-wire architectures requiring higher redundancy and fault coverage.
- Broader integration with ADAS torque overlay and software-defined vehicle architectures, where stable, deterministic EPS control becomes increasingly critical.

Chapter 6 The Impact Analysis: Strategies to Reflect the Transition from Sensor-Based to Sensorless EPS in a Systems Engineering Perspective

Introduction

The transition from sensor-based to sensorless control in Electric Power Steering (EPS) systems is not merely a substitution of a rotor position detection method. Rather, it represents a fundamental architectural reallocation of sensing, redundancy, diagnostic logic, and safety responsibility within the steering system.

In conventional EPS architectures, the rotor electrical position required for field-oriented control (FOC) is obtained through physical sensing devices such as resolvers, Hall-effect sensors, or magneto-resistive encoders. These sensors provide direct measurement of angular position, enabling deterministic Park and inverse Park transformations and stable torque control across the entire operating range. From a systems engineering perspective, the rotor sensor functions as an explicit hardware boundary between mechanical reality and electronic control.

When transitioning to a sensorless architecture, this physical boundary disappears. The electrical angle is no longer measured; it is estimated. The observer algorithm—whether based on back-electromotive force (back-EMF), magnetic saliency tracking, or hybrid methods—becomes a virtual sensor embedded within the control software. Consequently, responsibilities that were previously allocated to hardware components must now be redistributed across control algorithms, software structure, hardware computation resources, and functional safety mechanisms.

This redistribution has implications far beyond motor control theory. Removing a rotor position sensor affects:

- Mechanical and dynamic behavior at low speed,
- Electrical hardware design and signal integrity,
- Control loop architecture and estimator stability,
- Embedded software partitioning and execution timing,
- Functional safety diagnostics and fault detection coverage.

Therefore, sensorless EPS cannot be evaluated solely at the level of observer accuracy. It must be examined as a system architecture transformation.

From a design-change analysis standpoint—conceptually similar to “**Design Review Based on Failure Mode**” (DRBFM) [6-68] thinking—the elimination of a physical sensor constitutes a primary change point. The key engineering question becomes:

What must be reconsidered, redesigned, or revalidated across each subsystem when rotor position sensing is converted from hardware measurement to algorithmic estimation?

This chapter addresses that question through a structured impact analysis across five EPS subsystems:

1. Mechanics and dynamics,
2. Electrical hardware (ECU),
3. Control architecture (SMO + HFI hybrid),
4. Software architecture (AUTOSAR-based implementation),
5. Functional safety and diagnostic strategy.

The objective is not to perform a detailed DRBFM activity within this thesis. Instead, the purpose is to identify the structural implications of sensor removal and to clarify how system-level design must evolve in response.

In particular, this chapter provides the missing logical bridge between the algorithmic comparison presented in Chapter 5 and the final architectural conclusion of this thesis. Chapter 5 established that, among back-EMF-based methods, a Sliding Mode Observer (SMO) combined with High-Frequency Injection (HFI) provides the most favorable tradeoff for sedan-class EPS systems. However, such a conclusion is incomplete without demonstrating that this hybrid estimator can be coherently integrated into the broader EPS system architecture.

Accordingly, this chapter argues that the selection of an SMO + HFI hybrid structure is not only justified by estimation performance, but also by its compatibility with:

- Embedded real-time constraints,
- Software modularity under AUTOSAR,
- Redundant estimation cross-check strategies,
- Functional safety reallocation from hardware redundancy to algorithmic redundancy.

By analyzing the system-level impacts of sensorless conversion, this chapter reframes the observer not as an isolated control block, but as a structural component of the EPS architecture. This perspective enables a

coherent transition from conventional sensor-based EPS to a sensorless architecture that preserves safety, robustness, and production feasibility.

The following sections therefore examine, subsystem by subsystem, what changes when rotor position sensing is virtualized—and how those changes must be addressed to ensure that sensorless EPS remains technically sound, economically viable, and functionally safe.

6.1. Mechanics and dynamics,

In an Electric Power Steering (EPS) system, the “mechanics and dynamics” subsystem is the full torque-transmission chain that ultimately converts an electric motor’s electromagnetic torque (including any disturbances) into driver-perceived steering feel and road-wheel angle. Practically, this chain includes the steering column and intermediate shaft, torsion bar and torque sensing element, reduction mechanism (e.g., worm gear), rack-and-pinion or equivalent actuation interface, structural mounts, and the coupled vibration paths to the vehicle body. Because the engine is no longer the dominant masking noise source in electrified platforms, steering-system noise and vibration can become more perceptible to the driver; published NVH investigations of EPS assemblies explicitly frame modern vehicle contexts (hybrid/EV trends) as a reason steering NVH needs deeper attention, and demonstrate that full-assembly dynamic modeling (e.g., FEM-based) is now standard practice to capture these interactions [6-1].

Under the Chapter 6 framing (“sensor removal as architecture transformation”), the mechanics/dynamics subsystem is not being redesigned from scratch; instead, its *role* changes because rotor electrical position becomes an estimate whose quality depends on operating conditions. In other words: even when the same mechanical hardware is retained, the system now relies on software estimation strategies that must remain stable and perceptually acceptable *through the mechanics*, including during low-speed, high-assist conditions where friction, compliance, and road disturbances are most influential. The core mechanical insight for sensorless EPS is therefore: what was previously “absorbed” by the deterministic sensor-based torque control loop can reappear as torque ripple, oscillation, or acoustic artifacts that are transmitted through the steering mechanics and perceived as steering torque variation or noise [6-1], [6-4].

Change points introduced by sensorless conversion in mechanics and dynamics

The first and most important change point is low-speed and near-zero-speed operation. In sensor-based EPS, the rotor electrical angle is directly measured, so the control system can maintain stable torque production even when mechanical motion is slow (parking, creeping). In sensorless EPS—especially when low-speed operation is covered by saliency-based High-Frequency Injection (HFI) approaches—an injected signal is intentionally introduced to make rotor position observable. However, published HFI sensorless studies

explicitly treat *audible noise* as a first-order concern when high-frequency signals are injected at low speed, and propose methods that actively reduce this noise by adjusting injection amplitude to the drive state [6-4]. From a mechanics-and-dynamics viewpoint, this is precisely the architectural “handover”: the estimator’s observability mechanism can produce an acoustic and vibratory footprint, and that footprint is not purely electrical—it becomes steering feel and NVH through the mechanical path [6-1], [6-4].

The second change point is friction-dominant dynamics and manufacturing dispersion. In low-speed steering, the steering system can operate in regimes where friction and stiction dominate (worm gear, bearings, rack seals), and where small torque disturbances translate directly into steering-wheel torque variation. An experimentally grounded EPS study (test-rig based) reports that steering-system friction can be *highly load-dependent* and asymmetric with respect to steering speed, and notes that friction levels vary across steering devices as manufactured and installed—motivating friction level adaptation as a practical requirement [6-2]. This matters for sensorless EPS because estimator stability and torque-quality margins must be assessed not against one ideal plant, but across friction dispersions that shift both mechanical response (feel) and disturbance transmission. In effect, mechanical friction is no longer only a “feel calibration” issue; it becomes an estimator robustness boundary condition [6-2].

The third change point is structural vibration and rattle sensitivity to road reverse inputs. Industrial EPS engineering reports describe that rough-road reverse inputs can excite rattle noise originating in steering gear elements, and point out a fundamental tradeoff: tightening backlash to reduce rattle can increase friction and risk degrading driving stability, so designers must balance quietness and steering performance [6-3]. This tradeoff becomes more sensitive in sensorless EPS when low-speed control includes injected signals (or observer switching) that can introduce additional torque/force fluctuations. The implication is not that sensorless control “creates” rattle, but that it can change excitation content and makes it harder to rely on mechanical damping/friction alone as a “free filter,” particularly in quieter vehicle contexts where these noises are more noticeable [6-1], [6-3].

Literature synthesis for mechanics-relevant phenomena

Across the literature used here, the mechanics-and-dynamics picture can be consolidated into three engineering statements that are directly relevant to Chapter 6 bridging objective (algorithm → architecture). First, EPS is a coupled electromechanical NVH system, and credible studies treat full-assembly dynamic modeling as necessary to explain and correct real phenomena (e.g., lubricant-dependent or structural resonance effects) rather than viewing the steering system as an ideal rigid load [6-1].

Second, the low-speed steering regime is friction-sensitive and production-variable, and even control/estimation functions that appear “software-only” in block diagrams can become limited by friction

dispersion and asymmetry. Test-rig based EPS research formalizes this with both friction identification and the explicit observation that device-to-device variation requires adaptation strategies—an argument that maps cleanly into “production feasibility” rather than pure theory [6-2].

Third, noise, rattle, and torque variation are not secondary luxuries: they are part of steering system acceptability and can require dedicated bench evaluation and predictive design approaches, especially for vehicles where quietness is a differentiator. Practical EPS documentation using bench excitation and measured vibration acceleration shows how rattle noise becomes an engineering target, and explicitly frames quiet vehicle demand as a market-level driver [6-3].

Counter-arguments and boundary conditions

A reasonable counter-argument is: *“Mechanics are unchanged—if we remove the rotor sensor, the mechanical subsystem is physically identical, so Chapter 6.1 should not claim a mechanical redesign.”* This counter-argument is directionally correct if interpreted narrowly (no new gears, no new rack). The mechanical subsystem is not automatically re-engineered by the mere absence of a rotor sensor.

However, the literature still supports a stronger system-engineering rebuttal: the mechanical subsystem is the *carrier* of driver acceptance metrics (torque variation, rattle, NVH) and therefore becomes a stricter constraint when sensorless schemes introduce additional excitation mechanisms. HFI sensorless control work explicitly treats audible noise as a defect that must be mitigated by design choices (e.g., injection amplitude scheduling), which is exactly the kind of “new excitation channel” that the mechanical chain will transmit [6-4].

A second counter-argument is: *“Low-speed events are comfort issues, not fundamental safety issues.”* This is not fully reliable for EPS, because safety processes for EPS architectures treat steering malfunctions as safety-relevant and derive explicit safety goals and requirements for EPS systems. A major U.S. government functional safety assessment applies ISO 26262 concept-phase thinking and derives vehicle-level safety goals and a large set of EPS safety requirements, explicitly framing EPS as subject to systematic hazard analysis and test scenario development [6-7]. While Chapter 6.5 is the correct home for detailed functional safety logic, the mechanics/dynamics implication is: low-speed mechanical disturbances and unintended torque behaviors can be part of the hazard space that must be evaluated, not dismissed as purely “NVH tuning.” [6-7].

Verification methodologies, safety implications, and production feasibility

For mechanics/dynamics verification, the key is to use objective test definitions that quantify driver-relevant steering effort/variation and that can detect changes caused by estimator-induced excitations. The International Organization for Standardization provides an explicit low-speed/standstill steering effort test

method standard (though scoped primarily to heavy commercial vehicles and buses), where steering effort and feel are quantified using measurements such as steering-wheel torque, steering-wheel work, return ability, and steering-wheel torque variation. The same standard also emphasizes measurement integrity: it recommends a steering machine with emergency override, and warns that friction and inertia added by test instrumentation (e.g., torque sensors) can influence measured steering-wheel torque and should be estimated/compensated [6-5]. For a sedan-class EPS applications, the most defensible use is methodological: adopting its “what to measure” logic (torque variation, returnability) and its measurement cautions as a basis for ensuring sensorless-induced fluctuations are not hidden by test artifacts [6-5].

At the vehicle-handling level, SAE International J266 defines steady-state directional control test procedures for passenger cars and light trucks, including measurement of steering-wheel torque gradient with respect to lateral acceleration and other response gradients; it also notes that characteristic/critical speed and steer compliances may be determined within its framework [6-6]. This supports a clean verification logic for Chapter 6: *if sensorless EPS introduces additional low-speed torque variation (HFI-related) or changes steering feel attributes, the architecture must still satisfy standardized handling metrics and not introduce unintended response gradients across conditions* [6-6].

Production feasibility follows directly from the friction and NVH findings. Test-rig EPS research reports that friction varies by steering device as manufactured and installed, motivating friction adaptation concepts as a practical necessity [6-2]. Industrial steering-gear reporting similarly shows that rattle noise evaluation cannot rely only on subjective in-vehicle judgement: bench excitation and quantitative vibration evaluation are used to establish design-stage predictions and thresholds, particularly because quietness demand has increased in “quiet vehicles” and rough-road globalization has expanded expected operating conditions [6-3]. Together, these sources justify a production-feasible conclusion for 6.1: sensorless EPS (especially with HFI at low speed) must be validated against friction dispersion and rattle/NVH sensitivity, not just against mean-case observer accuracy, because the mechanical subsystem converts small algorithmic differences into repeatability issues at scale [6-2], [6-3], [6-4].

6.2 Electrical Hardware Design and Signal Integrity for Sensorless EPS

Hardware boundary shift when rotor position sensing is virtualized

In a conventional EPS motor-inverter-ECU stack, the rotor electrical angle required by FOC is “hard-measured” at the motor by a dedicated position sensor (resolver/Hall/magneto-resistive encoder) and delivered to the ECU as a relatively low-noise, explicitly bounded interface. In a sensorless EPS, that boundary disappears: the “angle signal” becomes the output of an observer that is reconstructed from inverter-side electrical signals (phase currents, DC-link voltage, switching states, and—in low/zero speed—saliency

responses to injected high-frequency excitation). This architectural change means that the ECU's analog and power-electronic signal integrity is no longer just a motor-control implementation detail; it becomes part of the sensing chain that produces the virtual sensor itself. Therefore, hardware design choices that previously affected “torque ripple” or “EMC margin” can directly affect angle observability, estimator stability, and diagnostic coverage in sensorless operation.

In EPS specifically, the low-speed, high-torque region (parking maneuvers, curb loads, low-velocity steering reversals) is both (i) the region in which back-EMF is weak and saliency-based methods (e.g., HFI) are typically required, and (ii) a region in which steering feel and noise-vibration-harshness constraints are stringent. Literature on sensorless “sensor-reduced” EPS-related drives shows that when measurement hardware is reduced or virtualized, compensating inverter nonidealities (e.g., dead-time-induced voltage errors) and managing the quality of electrical signals becomes essential, because estimation error can be dominated by the difference between commanded and actual inverter output voltages rather than by the motor model itself. [6-8]

A practical way to frame “ECU hardware impact” in sensorless EPS is to treat the observer as a software sensor whose raw inputs are produced by the inverter hardware. From this viewpoint, the ECU must reliably provide: (a) phase current feedback with sufficiently low offset drift and switching-noise sensitivity, (b) DC-link voltage and/or phase-voltage reconstruction accurate enough for back-EMF and disturbance modeling, and (c) deterministic sampling timing aligned with PWM edges to avoid biasing the reconstructed signals. The EPS-oriented “current sensorless” study in [6-8] is instructive (even though it removes current sensing rather than position sensing): it explicitly motivates compensation for dead-time-related voltage error because the mismatch between real voltages and voltage commands drives estimation error. The sensorless EPS transition similarly shifts burden from dedicated sensing hardware to “estimation-grade” inverter measurement and compensation mechanisms. [6-8]

Current sensing topology and sampling feasibility are central to this burden shift. Automotive inverters often use shunt-based sensing for cost/package, but shunt sensing introduces constraints on measurable time windows and can create operating regions where direct phase-current measurement becomes difficult unless PWM patterns and sampling strategy are co-designed. In SAE's automotive-focused inverter current sensing study [6-9], the authors compare single-shunt vs. three-shunt approaches and discuss limitations tied to minimum dwell time of switching states and “immeasurable regions,” motivating hybrid schemes that mix sensing and reconstruction. Separately, Cho-Ha-Sul formalize the current-measurement boundary problem for three-shunt sensing under SVPWM/DPWM and propose voltage-injection/sampling-point strategies to expand feasible measurement regions, emphasizing that sensing feasibility depends on modulation pattern and sampling timing rather than on sensor accuracy alone. These results are directly relevant to sensorless EPS

because estimator robustness depends on whether the ECU can always supply valid, unbiased current samples under all operating PWM states, including low modulation depth and fast transients. [6-9], [6-11]

In the low/zero-speed region where an SMO is commonly paired with HFI, the measurement bandwidth and noise floor requirements tighten further. Lu et al. demonstrate an HFI-based sensorless scheme that is specifically designed around a restricted inverter/sensing configuration (three-phase four-switch inverter with a single current sensor), illustrating how the feasibility of HFI demodulation, signal extraction, and estimator performance is conditioned by inverter topology and sensing constraints rather than by “control theory alone.” [6-10] Complementarily, Crossref metadata for Alaei et al. shows a focused study on how HFI amplitude/frequency and waveform design trade off acoustic noise, HF losses, and THD—all of which map back to inverter switching behavior and measurement contamination pathways that the ECU must manage if HFI is to be production-feasible in noise-sensitive applications like EPS. [6-12]

Switching noise, coupling paths, and their impact on estimator stability

The dominant signal-integrity challenge in sensorless EPS hardware is that inverter switching (high dv/dt , high di/dt) produces strong common-mode voltage steps and associated displacement currents that can couple into sensor traces, ADC references, and harness interfaces. Akagi’s ISPSD work attributes EMI issues in motor drives primarily to step-like common-mode voltage changes at the PWM inverter output and presents a compact passive mitigation approach, reinforcing that EMI is an intrinsic consequence of fast switching rather than an incidental layout problem. In a sensorless EPS, those same switching phenomena can directly corrupt the current and voltage signals feeding the observer, effectively injecting “angle noise” into what would otherwise be a deterministic transformation pipeline. [6-14]

The literature on leakage-current modeling and mitigation provides a more detailed physical picture of why sensorless architectures are hardware-sensitive. Ogasawara and Akagi model high-frequency leakage currents in PWM inverter-fed motor systems and show that mitigation strategies depend on the resonant behavior formed by parasitic capacitances/inductances; importantly, “simple” series common-mode chokes may influence peak currents but not necessarily RMS/average components, motivating more targeted damping approaches. [6-15] In parallel, Swamy–Yamada–Kume provide techniques specifically aimed at attenuating common-mode currents in PWM drives, a core mechanism by which inverter switching noise propagates into the surrounding electrical environment. [6-16] For EPS packaging where motor leads, ECU grounding, and chassis return paths are tightly constrained, these results justify why the ECU hardware “must be co-designed” with common-mode and differential-mode current control, not simply verified at the end of the program. [6-15], [6-16]

Beyond mitigation circuits themselves, signal-integrity risk is amplified by the fact that automotive motor-drive sensing chains often sit on mixed-signal PCBs where current-sense conditioning, PWM timers, and digital communication coexist. The IEEE Industry Applications Magazine survey by Skibinski–Kerkman–Schlegel provides a system-level treatment of EMI in PWM drives, including practical coupling mechanisms into nearby signals and an explicit discussion of current-sensing circuitry behavior under fast switching and long-cable conditions. [6-17] This is directly transferable to sensorless EPS because the observer is “only as good as” the measured currents/voltages after they have passed through these coupling paths; thus, EMI that would previously manifest as a “compliance issue” can manifest as estimator instability, chattering-like behavior in nonlinear observers, or degraded low-speed angle observability. [6-17]

Automotive EMC and electrical-environment constraints relevant to sensorless EPS

Automotive production feasibility requires passing both emissions and immunity regimes at the component/module level. CISPR 25 defines limits and measurement procedures for radio disturbances over a broad frequency range to protect on-board receivers (i.e., it directly operationalizes the constraint that “the ECU must not pollute the vehicle’s RF environment”). [6-18] ISO 11452-2 defines radiated immunity testing in an absorber-lined shielded enclosure (ALSE), which is particularly relevant for sensorless EPS because an observer-based angle estimator must maintain acceptable functional behavior under electromagnetic field exposure that can couple into harnesses and analog inputs. The key architectural point is that once rotor angle is virtualized, EMC disturbances can become estimation disturbances, so EMC compliance and estimator robustness become partially coupled engineering problems. [6-18], [6-19]

In addition to RF emissions/immunity, the ECU must tolerate the automotive supply environment (conducted transients, load switching, and voltage excursions). ISO 7637-2 provides bench test methods/procedures to ensure compatibility to conducted electrical transients along supply lines for devices installed in 12 V/24 V systems. [6-20] ISO 16750-2 describes electrical loads and associated stresses for road-vehicle electrical/electronic equipment. [6-21] For sensorless EPS, these standards matter not only for “survival” and reset immunity: supply disturbances can perturb DC-link voltage measurement, modulate inverter output, and temporarily invalidate the relationship between commanded and applied voltage—conditions that can drive observer bias or loss of lock if the hardware/software partition is not designed to detect and manage them (e.g., estimator hold/freeze strategies, plausibility gating, and recovery procedures triggered by power-quality monitors). [6-20], [6-21]

Production-feasibility implications and observed market design patterns

From a cost and packaging perspective, sensorless EPS is sometimes motivated by the removal of motor-side sensors and their associated harness/connectors. However, Section 6.2 must explicitly acknowledge the “cost

redistribution”: savings at the motor may be partially offset by tighter requirements on ECU sensing fidelity and EMC countermeasures (better current-sense front ends, improved PCB partitioning, filtering/shielding, and potentially higher-performance MCUs/ADCs to support synchronized sampling and demodulation). SAE’s current-sensing work [6-9] shows that even within the current-sensing problem alone, automotive inverter implementations frequently require hybrid sensing/reconstruction strategies to meet performance across operating regions—suggesting that the measurement chain is already an area where “simple hardware” can become “complex system design.” [6-9]

Market-facing EPS hardware trends also indicate that industry pushes integration to control wiring length and associated EMI exposure. JTEKT reports volume production of a column-type EPS with a small, motor-integrated ECU (MCU), emphasizing integration and packaging as product features. [6-22] From a signal-integrity standpoint, bringing the ECU physically closer to the motor can reduce motor-lead length (and therefore reduce the severity of cable-reflection and radiated/conducted coupling issues described in classic long-lead filter literature), while simultaneously raising thermal and vibration constraints on the ECU hardware. The Jouanne–Enjeti output-filter paper provides an analytical foundation for why cable length and dv/dt are not “secondary details”: long leads can induce overvoltage reflections and stress insulation unless dv/dt is controlled or filtered. [6-13] Together, these sources support a production-feasibility argument: sensorless EPS hardware is most credible when the mechanical packaging strategy (integrated or semi-integrated ECU) is explicitly leveraged to improve the estimator’s input signal integrity margin, rather than treated as an independent mechanical-only decision. [6-22], [6-13]

Counterarguments and conclusion

A credible literature review must also acknowledge counterarguments that favor retaining a physical rotor sensor in EPS. The strongest hardware-centric counterargument is that a resolver/encoder provides a deterministic, EMI-resilient boundary condition for angle, whereas a sensorless observer “inherits” the inverter’s switching noise, supply disturbances, and measurement imperfections. The EMI-oriented literature shows that common-mode and leakage-current phenomena are intrinsic to PWM inverters and can be complex to suppress across all operating conditions and parasitic configurations. [6-14]–[6-17] This supports a conservative position: if the ECU/motor harness environment cannot be sufficiently controlled (packaging constraints, cost limits on filtering/shielding, mixed-signal PCB density), sensor-based EPS may remain preferable for safety and robustness reasons. [6-14], [6-15], [6-17]

The system-level conclusion for Chapter 6 is that selecting an SMO+HFI hybrid is only architecturally coherent if the ECU hardware is must be engineered with estimator-grade signal integrity requirements. The most important hardware implications are: (i) current sensing and sampling must be co-designed with PWM strategy to avoid infeasible measurement regions and biased samples [6-9], [6-11]; (ii) HFI requires additional

measurement bandwidth and careful management of its side effects (losses, acoustic noise, spectral content) [6-10], [6-12]; (iii) EMI prevention and suppression are not “afterthought compliance tasks” but foundational to estimator stability [6-14]–[6-17]; and (iv) compliance with CISPR/ISO EMC and electrical transient standards must be interpreted through the lens of estimator robustness, because disturbances translate into angle-estimation disturbances after sensor virtualization [6-18]–[6-21]. These points complete the Subsystem 6.2 argument in a way that bridges Chapter 5’s algorithm selection to a viable production architecture. This hardware-level burden shift explains why Chapter 5’s selection of SMO+HFI cannot be evaluated purely at the algorithmic level.

6.3 Controls

Inputs, outputs, and architectural role of the virtual position sensor

In a sensorless EPS motor drive, the rotor electrical angle is no longer a “given” boundary condition provided by a resolver/encoder; it becomes a computed state that must be produced with defined latency, bounded error, and explicit confidence logic. As multiple surveys emphasize, this turns rotor position into a *software-measured quantity* whose quality depends on the excitation condition (speed, load, saliency), inverter nonidealities, and implementation timing. This is why many mature sensorless architectures are explicitly multi-source (different information carriers across the operating range), rather than being “one observer everywhere.” [6-23]

For Chapter 6’s *systems-engineering* intent, the key architectural statement is: the estimator is not a small block inside FOC; it is part of the closed-loop plant definition. It directly influences (i) the dq transformation used by current regulation, (ii) the torque production accuracy under transient steering commands, and (iii) the stability margins of the cascaded loops whenever the angle estimate is delayed or biased. Hybrid sensorless papers explicitly treat the estimator loop as a first-class control loop with its own bandwidth and stability constraints, particularly highlighting the “transition region” where information sources are switched or blended. [6-24]

Hybrid estimator partitioning across the EPS operating envelope

The structural reason for a hybrid is well-established: back-EMF–driven information is weak at (near) zero speed, while saliency-tracking via injected high-frequency excitation is strongest there; the situation reverses at medium/high speed where back-EMF becomes informative and injection becomes an avoidable disturbance. A broad review of position-sensorless PMSM drives frames this as a speed-dependent “information availability” problem that is best addressed by combining low-speed saliency methods with medium/high-speed model-based estimators. [6-23]

A mature architectural partition for an IPMSM-based EPS actuator can therefore be expressed as two cooperative estimation channels: - Low/zero-speed channel (HFI): angle from saliency response under injected HF voltage/current (typically aligned with an estimated d-axis to minimize torque ripple coupling), with explicit mechanisms to manage inverter voltage errors and acoustic/torque side-effects. [6-29], [6-32], [6-34]

Medium/high-speed channel (SMO family): angle from observer-estimated EMF (often extended EMF variants for saliency), with PLL-type tracking (or equivalent) shaping the angle dynamics and providing an interface between noisy estimated EMF and the control loops that need a smooth θ_e . [6-25], [6-26]

Importantly, this is not a conceptual “either/or.” The transition literature shows that, architecturally, both estimators are often running (or are at least “warm-startable”) near the transition region so that switch-over does not introduce a discontinuity in θ_e and does not force a re-synchronization event in the current loops. The “transition region” is treated as a design object in itself, not a threshold hack. [6-24] For clarity, the estimator roles across the operating regions are summarized in Table 6.1.

Operating region (architectural view)	Primary information carrier	Primary estimator block	Why it is chosen (system- level)	Typical risks that must be engineered out
Standstill to very low speed	Saliency via HF excitation	HFI channel + demodulation/PLL	Provides angle observability where EMF is insufficient; enables controlled startup and parking-range torque	Audible noise & added losses; sensitivity to inverter voltage error; phase delay from demodulation/filters
Transition region	Mixed (saliency + EMF)	Arbitration/blending logic	Prevents dq-frame “jumps” and unstable torque transients	Discontinuity in θ_e , estimator phase lag, inconsistent angle polarity/frame definitions
Medium to high speed	EMF (incl. extended EMF forms)	SMO(+PLL) channel	Minimizes injection disturbance; leverages higher SNR EMF	Sensitivity to sampling ratio, model/inverter nonidealities; chattering/phase lag if poorly conditioned

Table 6.1 Hybrid Estimator Partition Across Operating Regions; as summarized, the hybrid estimator assigns the information source that is structurally strongest in each operating region.

Integration with FOC torque control loops and estimator timing

In the SMO channel, a critical architectural requirement is that estimated angle dynamics be compatible with the inner current-loop bandwidth. A representative SMO paper (for PMSM sensorless position control) highlights SMO design explicitly to achieve sensorless control without relying on mechanical sensors, positioning the observer as the core enabler for FOC in the absence of direct position measurement. [6-25]

However, as multiple sensorless implementations emphasize, the “theoretical observer” is not what is deployed: the deployed structure is a discrete-time observer + smoothing/tracking estimator that (i) filters switching noise and harmonics, (ii) shapes estimator bandwidth, and (iii) produces a differentiable θ_e used by both Park transforms and speed-dependent scheduling. The quasi-sliding-mode observer approach for IPMSMs is explicitly motivated by executing the algorithm on digital controllers under realistic sampling constraints, i.e., the architecture is chosen so it can run at production sampling rates while preserving stability and accuracy. [6-26]

Timing is not a secondary detail—timing is the estimator. A key recent paper on SMO-based sensorless PMSM drives under low sampling ratio is directly relevant to production ECUs because it treats sampling ratio as a first-order design constraint (error mechanisms and design rules) rather than an implementation nuisance. This supports Chapter 6’s architectural claim that computation resources and timing partitioning meaningfully reshape what “observer choice” means in practice. [6-27]

On the HFI side, an explicit architectural advantage of square-wave/PWM-synchronous injection lines is the ability to extract position error signals with reduced reliance on low-pass filtering that would otherwise introduce phase delay into the estimation loop. The PWM-switching-frequency injection method for IPMSMs is directly aimed at designing the injection and processing so the sensorless method remains viable in real-time drives. This supports the EPS-oriented concern: parking and near-zero-speed steering can be highly dynamic (frequent reversals, micro-motions), so the estimator cannot be “slow by construction.” [6-29]

The architectural “glue” for the hybrid in this is the SMO+HFI structure, which is not merely conceptual but has been demonstrated in practice. A recent open-access study proposes an SMO approach enhanced by high-frequency voltage injection and PLL to cover very low speed and standstill, and demonstrates real-time operation on an IPMSM test platform. This evidence supports the central claim of this thesis that the SMO+HFI hybrid is not only accurate but also integratable as a coherent control architecture. [6-28]

Transition logic, arbitration, and robustness measures

The best-established architectural lesson from the “transition region” literature is that switching is not only about speed thresholds. Instead, the hybrid estimator should be treated as an arbitrated state machine with continuity constraints, where the output is not simply “ θ from method A or B,” but a single, continuous θ_e stream that meets constraints required by the current controllers. The transition-region paper frames this as a sensorless estimator that uses signal injection at low speed and EMF at higher speed, with *particular focus on the transition region*, i.e., the switching regime itself is explicitly analyzed. [6-24]

1. A production-feasible arbitration logic for EPS can be expressed without re-deriving any mathematics, as a set of architectural constraints: 1. Continuity constraint: θ_e must not jump (bounded derivative); torque transients caused by dq-frame discontinuities are architecturally unacceptable in steering. [6-24]
2. Consistency constraint: both estimators must share the same polarity conventions, electrical/mechanical angle mapping, and reference-frame definitions (a common integration failure mode in hybrid prototypes). [6-24]
3. Confidence constraint: arbitration should prefer the estimator with better instantaneous “observability,” which can be implemented via estimator residuals, PLL error energy, or stability/lock indicators (rather than speed alone). Hybrid EV-oriented architectures explicitly implement full-speed-range hybrids (HFI at low speed, PLL/back-EMF at high speed) with *smooth transitions* as a named design objective. [6-33]
4. Fault containment constraint: if one estimator becomes inconsistent (e.g., injection path corrupted by voltage error; SMO corrupted by sampling/harmonics), the system must degrade gracefully to the other estimator while flagging diagnostics upstream (this connects to Chapter 6.5, but the control architecture must expose the required signals). [6-23]

Robustness is not only “observer robustness.” It includes robustness to inverter voltage errors and distortions. One HF square-wave injection paper explicitly targets high robustness to voltage errors in position sensorless control, which is especially relevant in automotive ECUs where dead-time effects, parameter drift, and voltage drops can be non-negligible across temperature and load. This reference belongs in the control-architecture subsection because it justifies architectural mitigations (compensation blocks, voltage error-tolerant injection algorithms) rather than re-derivation of control theory. [6-32]

Another non-negotiable robustness issue is estimation delay: injected-signal methods often require demodulation and filtering, which can introduce phase lag that directly acts like an angle delay inside FOC. The SLED 2012 paper on position estimation delays investigates causes and effects of such delays and discusses compensation approaches. Architecturally, this supports the point that the HFI estimator must be engineered as a latency-budgeted pipeline (sampling $\rightarrow$ demodulation $\rightarrow$ filtering $\rightarrow$ PLL/angle update), with compensation aligned to the closed-loop bandwidth targets of EPS torque control. [6-31]

Evidence, counter-arguments, and verification methods

The literature provides direct “bridge evidence” that a hybrid estimator is not only academically viable but aligned with automotive-grade needs. An EV-oriented hybrid sensorless strategy (PLL + HFI) makes “smooth transitions” a core goal, which is conceptually consistent with EPS requirements for repeatable torque delivery and stable operation through frequent reversals, even though the application domain differs. This is an appropriate “market-adjacent” case example because EV traction drives share harsh inverter environments and wide operating envelopes, and they face similar pressure for sensor elimination. [6-33]

A steering-specific case example exists in the steer-by-wire literature: a conference paper on sensorless position tracking for steer-by-wire applications reports that a sensorless observer can track steering transients and can be used as backup in case of encoder failure, using a high-frequency injection approach. Even though steer-by-wire is not identical to EPS, the steering-domain constraints (bandwidth for steering transients, safety-motivated redundancy thinking) map closely to the “system architecture transformation” argument that Chapter 6 is making. [6-35]

Counter-arguments must be acknowledged explicitly in Chapter 6.3 because they influence architecture choice: - HFI side-effects: acoustic noise and additional losses are a major practical objection; modern work explicitly proposes spread-spectrum / random-sequence injection to reduce discrete spectral peaks and audible artifacts while preserving saliency observability. This motivates architectural scheduling (injection amplitude/frequency shaping) rather than a purely fixed injection design. [6-34]

- SMO implementation risk: SMO robustness is attractive, but discrete-time realization under low sampling ratio and the need to manage phase lag/noise places hard constraints on controller scheduling and computation partitioning. This is why low-sampling-ratio error analysis and quasi-SMO structures are directly relevant to “architecture,” not merely “control design.” [6-26] [6-27]
- Transition-region fragility: unstable or discontinuous switching can create dq-frame discontinuity torque spikes; transition-region studies treat this as the primary pitfall and motivate blended or carefully parameterized switching functions. [6-24]
- Machine/inverter nonidealities: carrier-signal choice and processing must account for estimation errors introduced by both machine and inverter nonideal attributes; this is a fundamental design-pressure pushing the architecture toward robust carrier selection and explicit error budgeting. [6-30]

Verification methodology in a system level architecture argument should be framed as *what must be validated* rather than *how to derive it*. Across the cited works, the recurring validation pattern is: (i) demonstrate low/zero speed operation, (ii) demonstrate transition behavior without estimator loss-of-lock, and (iii) demonstrate robustness under practical distortions (voltage error, sampling constraints). The Ates et al. work

is particularly relevant because it reports real-time implementation spanning low-speed/standstill using injection-enhanced SMO + PLL, while the steer-by-wire work motivates validation against steering transients. Together, they justify a verification strategy for sensorless EPS that combines dynamometer testing (motor-level) with steering-transient profiles (system-level) and explicitly targets the transition region as a validation item. [6-28] [6-35]

In summary, the SMO + HFI hybrid is architecturally coherent for sensorless EPS because it assigns *observability responsibility* to the method that is structurally strongest in each region (saliency at low speed; EMF at higher speed), while explicitly treating switching, delay compensation, and nonidealities as first-class architecture elements. This matches the Chapter 6 objective stated: demonstrating that algorithm choice (from Chapter 5) remains coherent when expanded into a full control-system structure that respects embedded constraints and steering-domain dynamics. [6-23] [6-24] [6-28]

6.4 Software architecture impact analysis for sensorless EPS under AUTOSAR

Architectural change points introduced by sensor virtualization

This subsection focuses on what changes in the ECU embedded software architecture when rotor electrical angle sensing is “virtualized” (i.e., moved from a physical sensor into software estimation), consistent with the system-level intent, because the rotor position becomes a software-generated quantity, its generation, validation, and distribution must be explicitly managed within the ECU software architecture.

In a conventional sensor-based EPS, the software architecture can treat rotor electrical angle as a hard boundary input: the angle arrives through a dedicated sensing chain (sensor → signal conditioning → MCU interface), then becomes an input to the motor-control stack. In a sensorless EPS, the “angle signal” becomes an internal software product computed by observers and signal-processing routines that must execute with strict timing, defined numerical assumptions, and explicit confidence monitoring. This shift is architectural, not just algorithmic, because it introduces a new high-criticality data product (estimated position) whose creation, distribution, and validation must be integrated into the ECU’s scheduling model, communication semantics, diagnostics, and toolchain [6-48].

Two software-level change points are particularly relevant for an SMO + HFI hybrid estimator in an EPS ECU:

First, timing coupling becomes a first-class architectural requirement. HFI-based estimation inherently relies on intentional injection waveforms and tightly synchronized sampling/demodulation steps, which makes estimation quality sensitive to jitter and latency in ADC sampling, PWM update, runnable dispatch, and data handoff across software layers. Once the estimator is integrated into AUTOSAR scheduling, timing analysis

can no longer be “implicit”; it must be explicitly modeled, budgeted, and verified end-to-end across runnables, OS tasks, and relevant basic software services.

Second, the ECU must now treat “angle validity” as a managed state rather than an always-deterministic sensor channel. The software architecture must support start-up transitions (e.g., pre-estimation alignment, low-speed saliency tracking, transition to back-EMF observer dominance), and it must define what the rest of the control system is allowed to consume at each phase (e.g., raw estimate, filtered estimate, quality flags, fallback states). The practical architectural implication is that the estimator is not merely an internal function but a software “virtual sensor service” with a defined interface contract and diagnostic behavior. [6-48]

AUTOSAR layering and component allocation for sensorless EPS

A sensorless EPS ECU that must remain production-feasible typically benefits from explicit separation between (a) hard real-time motor-control execution and (b) supervisory, diagnostic, and state-management logic, while still preserving software modularity, configurability, and supplier/OEM integration workflows. AUTOSAR Classic Platform’s layered model (Application, RTE, Basic Software) is directly suited for this separation, provided that the time-critical parts are allocated to the appropriate architectural “depth.”

AUTOSAR’s Layered Software Architecture explanatory document explicitly defines the three major layers and positions “Complex Drivers” as the standardized integration point for non-standard or highly hardware-coupled software modules, which is especially relevant for motor-control loops and estimator pipelines requiring tight synchronization with peripherals. [6-36], [6-37], [6-39]

A practical AUTOSAR allocation (consistent with production patterns reported for power steering controllers) is to implement the control feature as a set of software components (SW-Cs) whose internal behavior is expressed as runnables, while treating the most timing-sensitive peripheral interaction and high-rate signal conditioning as Complex Driver / tightly coupled BSW integration. A directly relevant industrial-academic example is the motor-driven power steering (MDPS) AUTOSAR methodology paper by Kim, Lee, and Ahn, which focuses on a software component architecture for MDPS control systems using AUTOSAR concepts, demonstrating feasibility of AUTOSAR-based component decomposition and integration for steering applications. [6-36], [6-37]

Conceptually, the SMO + HFI hybrid can be placed as follows:

Application Layer (SW-Cs)

- Torque control / assist logic SW-C
- Sensorless Estimation Service SW-C (SMO state update, HFI mode/state mgmt)
- Mode/state coordination SW-C (start-up, transition logic, degrade logic)
- Diagnostic monitor SW-Cs (plausibility, quality flags, watchdog checkpoints)

RTE

- Data exchange between SW-Cs (signals, IRVs, sender/receiver)
- Mode communication and controlled access patterns

Basic Software (BSW)

- OS (tasks/alarms/ISRs; memory & timing protection mechanisms)
- Diagnostic stack (Dem, Dcm)
- E2E protection library for safety-relevant data exchange (where applied)
- MCAL drivers (PWM, ADC, ICU, SPI, etc.)
- Complex Drivers (when strict latency/jitter and peripheral coupling dominate)

This allocation is not an attempt to “hide” control complexity inside AUTOSAR abstractions; rather, it uses AUTOSAR to formalize interfaces and lifecycle management, while recognizing that high-frequency estimator loops may require deeper integration to meet timing integrity. AUTOSAR’s own RTE requirements explicitly allow conceptual communication via the RTE to be implemented as direct mechanisms in some cases (i.e., an implementation may bypass RTE for specific paths), which supports the common motor-control engineering practice of limiting abstraction overhead on ultra-fast paths. [6-39]

Real-time scheduling, jitter control, and resource budgeting

The key software-architecture question for SMO + HFI integration is not “can the ECU compute the estimate,” but “can the ECU compute it with bounded latency and jitter, while preserving system schedulability and diagnosability.” AUTOSAR’s timing analysis technical report is explicit that timing analysis within the AUTOSAR development process should progress from timing requirement definition to verification at component and system levels, including ECU implementation concerns. This is directly applicable when adding HFI-related runnable and filters that increase CPU load and introduce additional periodic activities. [6-38], [6-40], [6-41]

At the OS layer, AUTOSAR OS is described as statically configured and intended to be amenable to real-time performance reasoning, and it defines mechanisms relevant to controlling faults and protection behaviors (e.g., hook routines, protection reactions). This matters because the estimator tends to be both high-rate and safety-relevant-by-architecture: faults in its execution (deadline misses, overruns, invalid intermediate states) must be contained, detected, and managed. [6-38]

A well-documented and recurring counter-argument is that AUTOSAR layering (especially RTE and BSW interaction) can introduce non-trivial overhead and can impact schedulability margins. Chauhan’s University

of Waterloo thesis specifically studies AUTOSAR overhead effects on software timing and schedulability, reinforcing that overhead is not merely theoretical—it must be modeled and budgeted, especially when integrating additional high-rate functions. For sensorless EPS, this counter-argument becomes practical when HFI demodulation/filtering and SMO updates are added at rates comparable to (or harmonized with) current-loop execution. [6-41]

A software-architecture mitigation strategy is to apply determinism-preserving execution semantics for critical data paths. Biondi et al. discuss Logical Execution Time (LET) concepts in the context of AUTOSAR processes, describing LET as an approach to eliminate output jitter by delaying outputs to the end of a task period (trading latency for reduced jitter), and they explicitly tie analysis to typical AUTOSAR runnable/RTE structures. For sensorless estimation, LET-style thinking is valuable not as a “required standard feature,” but as an architectural lens: decide which estimator outputs must be jitter-minimized and structure runnable execution and data release points accordingly. [6-42]

In practice, this yields a coherent scheduling pattern: tightly synchronized ISR/peripheral-driven acquisition and actuation, periodic “fast control” tasks for core estimator and FOC calculations, and slower supervisory/diagnostic tasks that validate estimator quality and manage transitions. The essential point for Chapter 6 is that SMO + HFI is compatible with this pattern: the architecture can isolate time-critical loops while preserving SWC boundaries for state machines, monitoring logic, and integration with the vehicle software ecosystem. [6-42]

Diagnostics, safe data exchange, and serviceability

Sensorless EPS adds failure modes that are intrinsically software-mediated: estimator divergence, low-speed observability loss, injection/sampling misalignment, and mode-transition instability. Even without expanding ISO clauses, the software architecture must be demonstrably capable of (a) surfacing these faults, (b) storing them consistently, and (c) supporting service diagnostics and production testing.

Within AUTOSAR Classic Platform, these responsibilities map naturally into the standardized diagnostic stack. The Diagnostic Event Manager (Dem) specification defines Dem as the basic software module that provides a common approach to diagnostic fault memory, offering interfaces to the application layer and other BSW modules and supporting interactions with Dcm for stored DTC retrieval. That makes Dem the natural sink for sensorless-estimator diagnostic monitors (e.g., “angle plausibility,” “observer convergence timeout,” “HFI signal integrity”). [6-46]

The Diagnostic Communication Manager (Dcm) specification defines the Dcm module as providing a common API for diagnostic services in AUTOSAR, positioning it as the standard pathway for development/service tooling to interrogate ECU state and fault memory. This is the primary AUTOSAR

interface point for implementing UDS-based workflows (read/clear DTCs, read data by identifier, routine control, etc.) for sensorless EPS calibration and service. [6-47]

At the protocol level, ISO 14229-1:2020 specifies diagnostic services at the application layer allowing a tester (client) to control diagnostic functions in an ECU (server). The architectural implication for sensorless EPS is straightforward: because angle estimation becomes a software-generated product, service tooling must rely on diagnostic exposure (through Dcm/UDS mechanisms) to access estimator states, quality metrics, and event histories in a structured way—especially in production debugging and field return analysis. [6-49]

For safety-related data paths (for example, if the ECU partitions safety-relevant functions and requires protected communication), AUTOSAR provides an End-to-End Communication Protection Library specification. The E2E library is designed around runtime configuration passed through call parameters and is explicitly aligned with functional safety terminology and communication fault considerations. In an AUTOSAR-based sensorless EPS architecture, E2E protection is a coherent mechanism for protecting safety-relevant data exchange (e.g., critical state/quality signals crossing partitions or software clusters) without returning to hardware redundancy as the only solution. [6-45], [6-46]

Verification methodology and production feasibility under a systems-engineering lens

The production feasibility question for Chapter 6.4 is whether the SMO + HFI hybrid estimator can be integrated into an AUTOSAR pipeline without creating unverifiable timing behavior or unmanageable integration complexity. The answer depends on whether the software architecture supports evidence-based verification across the scheduling model and toolchain—not just algorithm correctness.

On the formal/timing verification side, Zhang et al. describe an approach that models systems defined by AUTOSAR specifications as timed automata and verifies timing properties (using UPPAAL), emphasizing that AUTOSAR timing behaviors can be extracted from the three-layer architecture (Application software, RTE, and BSW layers). While this type of tooling is not a mandatory industrial artifact, it strengthens the argument that AUTOSAR architectures can be subjected to formal timing verification methodologies—highly relevant when adding estimator computation paths and mode transitions. [6-43]

On toolchain and integration feasibility, Sandmann and Seibt’s SAE work on AUTOSAR-compliant development workflows highlights the practical need for interoperability across architecture tools, model-based design, code generation, and iterative round-trip engineering, using AUTOSAR standardized exchange formats (ARXML) and emphasizing the iterative nature of architecture changes. This directly supports the “systems engineering” point of Chapter 6: removing a hardware sensor is a design change point that triggers iterative adjustments across architecture, runnables/interfaces, verification artifacts, and integration steps. [6-44]

Finally, software production readiness in safety-relevant automotive ECUs is typically reinforced by disciplined coding practices and analysis. MISRA C:2012 is documented as guidance for safe use of the C language in critical systems and is widely used as a reference point for coding discipline in embedded contexts. For this chapter's argument, MISRA is not about re-stating coding rules; it is evidence that, once position sensing becomes software-derived, the development process must treat the estimator and its monitors as high-integrity embedded software requiring analyzable implementation practices. Once rotor position becomes a software-derived safety-relevant signal, coding discipline and static analyzability become architectural requirements rather than project preferences.

This software-architecture analysis demonstrates that the SMO+HFI hybrid selected in Chapter 5 remains compatible with AUTOSAR-based production ECUs when timing, diagnostic, and safety integration are explicitly engineered. [6-43], [6-44], [6-48], [6-50]

6.5 Functional Safety and Diagnostic Strategy

Safety responsibility shifts when the rotor sensor becomes a virtual sensor

Architectural meaning of “sensor removal.”

In a sensor-based EPS, rotor electrical angle is measured by dedicated hardware (resolver / Hall / encoder), and the field-oriented control loop treats that measurement as a bounded, safety-relevant interface between the mechanical rotor and software torque control. In a sensorless EPS, the rotor angle is estimated from electrical signals; therefore, the estimator becomes a safety-related element whose correctness depends on a chain of assumptions across sensing (current/voltage), computation timing, signal integrity, and algorithm stability. This is precisely the type of system-level change that functional safety processes are designed to address: hazards due to malfunctioning behavior of E/E systems are controlled by a lifecycle that begins with item definition, hazard analysis and risk assessment (HARA), and a functional safety concept (FSC), then flows into system-level technical safety requirements (TSRs) and architectural allocation. [6-51], [6-52].

Why EPS safety is evaluated at the “foundational system” level, not at the observer block level.

Steering is consistently treated as a foundational actuation function for vehicle safety (and, increasingly, for automated functions that demand predictable steering response). Public functional-safety studies conducted by the National Highway Traffic Safety Administration and U.S. DOT research partners have explicitly applied ISO 26262 concept-phase processes to generic EPS architectures, combining multiple hazard analysis techniques (e.g., HAZOP and STPA) and then deriving safety goals and functional safety requirements from the identified hazards. In the EPS-focused report, the study states that it follows ISO 26262 concept phase, applies HAZOP, functional FMEA and STPA, identifies four vehicle-level safety goals, and derives 125 EPS

system safety requirements, also proposing test scenarios and areas for diagnostic trouble-code coverage. [6-58].

A closely related ESV paper applies the same ISO 26262 concept-phase framing to both EPS and steer-by-wire, and explicitly discusses how “fail-safe” versus “fail-operational” needs can influence the functional safety concept and requirements derivation—an important point because “safe state” expectations differ depending on whether mechanical fallback exists. In a conventional EPS architecture with a mechanical steering fallback path, the minimum-risk condition typically corresponds to disabling or limiting assist torque while preserving manual steering capability and informing the driver through warning indicators. [6-59].

Load-bearing hazard example for EPS: unintended assist and time-to-mitigate.

In an EPS hazard analysis example presented for a column-drive architecture, Hyungju Kwon and coauthors focus on the failure mode “Unintended Steering Assist,” rate it as ASIL D, and connect hazard testing results to defining a “fault tolerant interval” (the time available to detect and mitigate the fault and revert to a safe state). This is directly relevant to sensorless EPS because estimator divergence or sign/phase errors can be a primary trigger for unintended assist torque. [6-60].

Change point analysis: sensorless estimation introduces new safety-relevant failure classes

Functional safety versus “algorithmic insufficiency.”

A sensorless estimator changes the safety argument in two ways:

- 1) Random hardware failures remain (ADC, current sensors, power stage, MCU), but the hazard pathway changes: the same hardware fault now corrupts both torque control and the virtual sensor, increasing the need for robust plausibility checks and safe-state reactions at the system level. Hardware-oriented requirements such as safety requirements allocation to hardware, architectural metrics evaluation, and analysis of safety goal violations due to random hardware failures are core ISO 26262 hardware-development topics. [6-53].
- 2) Systematic failures (software defects, model/parameter errors, timing violations) become more prominent because the angle estimate is a software-derived quantity. ISO 26262 treats systematic control via software safety requirements, software architecture/unit design, and verification at multiple integration levels. [6-54].
- 3) “Correctly functioning but still unsafe” behavior becomes more visible in sensorless EPS, especially near low speed/standstill, temperature extremes, saturation, or high noise. These are cases where the system may not be faulty in the ISO 26262 sense, but the intended functionality can be insufficient in certain operational scenarios. This is the domain of SOTIF (Safety of the Intended Functionality), formalized in ISO 21448, and it is relevant to sensorless EPS because estimation accuracy can degrade without a discrete fault event. [6-56].

Sensorless-specific “malfunctioning behavior” patterns that matter for safety.

For a hybrid estimator (SMO + HFI), the safety problem is not “mean angle error,” but which estimator failure patterns can create hazardous torque and how those are detected within the fault tolerant time interval.

- SMO-related risks (medium/high speed channel): Sliding-mode observers are valued for robustness but are known to require careful design for chattering reduction and stable/accurate back-EMF reconstruction. A representative IEEE Transactions paper proposes a high-speed SMO structure using a smooth nonlinearity (sigmoid/boundary-layer approach) and validates the concept with simulations and experiments, illustrating both the attractiveness and the sensitivity of SMO-based sensorless estimation to implementation details. [6-64].
- HFI-related risks (zero/low speed channel): Classic self-sensing work shows that injection-based methods can produce position/velocity estimates down to standstill by exploiting saliency/inductance variation, using the motor and inverter as part of the sensing mechanism. This is one of the foundational enablers for low-speed sensorless control in salient-pole PMSM/IPMSM contexts. [6-66].
- A critical counter-argument for HFI (and thus a key safety caveat): the “saliency position” tracked by HF injection can differ from the true rotor position and may include offset and rotor-position-varying components. This has two safety implications: (i) reliance on HFI alone can create systematic bias in estimated angle, and (ii) even a correctly operating HFI algorithm may deliver an estimate that is not sufficiently aligned to guarantee correct torque direction under all operating conditions—again linking to SOTIF-style reasoning. [6-65], [6-56].

A sensorless EPS safety argument must therefore treat “SMO + HFI” not as a single estimator, but as a two-channel virtual sensing subsystem whose cross-consistency properties are part of the safety mechanisms.

Diagnostics and safety mechanisms aligned with the SMO + HFI hybrid architecture

Safety goal framing and “what must be detected.”

Across EPS concept-phase studies, the recurring safety concern is preventing hazardous steering behavior (e.g., unintended assist, steering in the wrong direction, excessive assist, or loss of controllability). ISO 26262 concept and system phases require that identified hazards be converted into safety goals and then into requirements allocated to architectural elements with defined reactions and diagnostic expectations. [6-51], [6-52], [6-58].

For sensorless EPS, a practical decomposition of estimator-related hazardous causes (for DRBFM-like “change point” thinking) is:

- Angle plausibility loss (large angle error; wrong quadrant; discontinuities).
- Angle dynamic implausibility (estimated speed inconsistent with electrical/mechanical constraints).
- Estimator divergence or saturation (residuals grow; internal states clamp).
- Common-cause sensor-chain corruption (current/voltage sensing faults drive both torque and estimation wrong in the same direction).
- Mode-transition faults (switching SMO↔HFI introduces transient mismatch).

Diagnostic strategy: analytical redundancy plus hardware safety mechanisms.

ISO 26262 hardware development emphasizes achieving safety goals under random failures via architectural measures and quantification (e.g., diagnostic coverage for safety mechanisms). [6-53].

In sensorless EPS, the diagnosability challenge is that SMO and HFI share much of the same measurement chain (phase current sensing and DC-link/inverter voltage knowledge), so the main benefit is analytical redundancy rather than independent physical redundancy. This makes the following safety mechanisms structurally important:

- Estimator cross-check (SMO vs HFI) with gated activation: Use both estimates during overlap regions (or perform periodic consistency checks) and declare “angle invalid” when disagreement exceeds a calibrated threshold for longer than a calibrated persistence time. The key is persistence to avoid nuisance triggers, while still meeting “the derived fault tolerant time interval” (FTTI). For unintended assist hazards rated at ASIL D, the estimator fault detection and mitigation must occur within FTTI, consistent with ISO 26262 concept-phase allocation.
- Residual-based monitoring per estimator: Monitor observer innovation/residual signals, saturation flags, or back-EMF reconstruction quality indicators for SMO; monitor injected-signal response amplitude/phase and demodulation quality for HFI.
- Torque-direction and torque-magnitude plausibility: Cross-check commanded torque/current vector against measured current response, including limits on torque rate and bounds on dq current trajectories.
- Safe state reaction: For conventional EPS with mechanical fallback, “no/limited assist” is typically the minimum-risk state; the system must disable assist torque while preserving manual steering path (and notify the driver), consistent with foundational-system safety thinking. [6-59], [6-58].

A crucial supporting method is quantitative diagnostic coverage evaluation for difference-check mechanisms when signals are correlated. A SAE journal article explicitly analyzes how diagnostic coverage for correlation/difference checks can be derived statistically, discussing sensor systems where two paths measure

the same physical quantity and a difference-check decides whether to pass the data or flag an error. The paper also notes that this principle is used in applications such as steering-torque sensing and current sensing of actuators, and ties diagnostic coverage figures to FMEDA/FTA needs. [6-62].

ASIL decomposition and dependent failures: why it is non-trivial for SMO + HFI.

A common temptation is to treat “SMO” and “HFI” as two redundant channels enabling ASIL decomposition. However, ISO 26262 explicitly treats ASIL-oriented analyses, including requirements decomposition, coexistence criteria, and dependent failure analysis as part of ensuring that “redundancy” is not undermined by common-cause pathways. In sensorless EPS, SMO and HFI are algorithmically distinct, but they typically share MCU execution environment, ADC, current sensors, inverter, and often the same calibration data. Therefore, the appropriate safety claim is usually **increased fault detection coverage** (analytical redundancy) rather than full independence suitable for aggressive ASIL decomposition without additional measures.

[6-55], [6-62].

Market implementation cases: fail-safe EPS vs fail-operational steering.

The ESV steering paper highlights that fail-safe and fail-operational needs differ and that the functional safety concept must account for driver engagement and the availability of a minimum-risk condition. [6-59].

A concrete “fail-operational” implementation trend is dual-lane EPS architectures (dual-winding PMSM + dual ECUs), motivated by higher automation needs. An open-access IEEE Access paper explicitly states that, in EPS systems, redundancy can be achieved by applying dual-winding PMSMs and dual ECUs, and it proposes diagnostic strategies to detect disagreement between lanes and to distinguish which lane is faulty—illustrating how diagnostic logic becomes a first-class architectural element when continuous operation is required. [6-67].

From a sensorless-EPS system perspective, this is a valuable market case because removing the rotor sensor (hardware deletion) is often justified economically, but functional safety pressure can drive architecture in the opposite direction (more monitoring, more redundancy, more structured disagreement handling). The system-level implication is that “sensorless” does not automatically mean “simpler safety case”; it can instead mean “different safety argument,” with more emphasis on monitor quality and failure containment.

Production feasibility and verification methodologies in an AUTOSAR EPS ECU

Production feasibility: safety mechanisms must be compatible with AUTOSAR partitioning and safety evidence.

For an AUTOSAR-based ECU, functional safety demands are not only about “having diagnostics,” but about implementable separation, traceability, and verification evidence. The AUTOSAR Safety Extensions

specification explicitly states that it aims to enable ISO 26262 development in an AUTOSAR context, standardizing the exchange of safety information and supporting consistent management across vendors and tools. [6-57].

At the software safety level, a practical feasibility hotspot is freedom from interference (e.g., memory protection, execution separation, and preventing non-safety software from corrupting safety software). A recent SAE paper on EPS protection implementation explicitly proposes memory protection as a software safety mechanism and states that it is realized in AUTOSAR architecture by configuration—directly aligning with the type of safety partitioning an estimator-plus-monitor architecture needs (observer as safety-related SW element; monitor potentially as an independent safety mechanism). [6-63], [6-54].

Verification methodologies: demonstrating safety is about fault coverage and time, not estimator accuracy plots.

A functional safety analysis must demonstrate how safety evidence can be generated and verified. The strongest publicly available EPS safety studies emphasize concept-phase rigor and testability:

- The NHTSA EPS report documents that it follows ISO concept-phase processes and applies HAZOP, functional FMEA, and STPA; it derives a large set of safety requirements and connects results to potential test scenarios and diagnostic trouble-code coverage. This is a blueprint-style example of how hazard analysis outputs can be made verifiable. [6-58].
- The ESV steering paper similarly frames the development of a functional safety concept through multiple safety analysis methods and explicitly discusses fail-safe vs fail-operational needs that can change what “verification success” means. [6-59].
- The ISO 26262 process context for these activities is anchored in the concept phase (item definition, HARA, FSC) and then system-level product development (TSRs, TSC, integration and safety validation). [6-51], [6-52]. An industry-oriented supporting example for proof structure is an SAE technical paper describing EPAS development under ISO 26262, performing HARA, improving architecture to meet a safety goal, and using FTA/FMEA to support reliability/safety analysis. Even if the underlying details are proprietary, the paper demonstrates the expected methodological shape of evidence for an EPS safety case. [6-61].

Therefore, the removal of a physical rotor sensor does not reduce safety responsibility; instead, it redistributes safety assurance from hardware determinism to estimator integrity, diagnostic coverage, and controlled fault handling within the software-defined architecture.

Chapter 7 Methodology:

Proposed simulation model for Sensorless Control System by Sliding Mode Observer

7.1 Purpose and Scope

7.1.1 Role of Modeling in This Thesis

This thesis is primarily conducted from a system engineering perspective for automotive electric power steering (EPS) systems. Accordingly, the overall scope of the thesis spans a wide range of domains, including functional safety, mechanical structure, ECU hardware, and software development processes. Within this broad system-level context, the purpose of the present chapter is to focus on a well-defined and critical subsystem: the modeling, control, and sensorless position estimation of the EPS drive motor.

The modeling presented in this chapter is not intended to represent the entire EPS system or vehicle dynamics. Instead, it serves as a quantitative verification of the motor drive subsystem, which forms the foundation upon which higher-level EPS functionalities are built. In particular, this chapter evaluates whether a sensorless field-oriented control (FOC) scheme can provide sufficiently accurate and stable electrical angle information to support closed-loop motor operation.

By isolating the motor control and estimation problem, the modeling effort enables a clear assessment of the underlying control performance without conflating it with vehicle-level effects such as steering feel, tire–road interaction, or rack-and-pinion mechanics. This separation is consistent with a system engineering approach, in which complex systems are validated through the verification of well-defined subsystems before full integration. [7-1], [7-2], [7-3]

7.1.2 What Is Evaluated and What Is Not

Based on the above perspective, the scope of evaluation in this chapter is explicitly constrained.

Evaluated Items

The following aspects are quantitatively evaluated in this chapter:

- IPMSM-based motor model

The permanent magnet synchronous motor model is parameterized to represent an interior PMSM

(IPMSM), which is commonly used in EPS applications due to its efficiency and controllability characteristics.

- SMO-based sensorless FOC in the medium-to-high speed region

A sliding mode observer (SMO) is employed to estimate the rotor electrical position, which is required for Park and inverse Park transformations in FOC.

The evaluation focuses on the medium-to-high speed operating region, where back-EMF-based observers are theoretically observable and practically applicable.

- Control performance at the motor-drive level

Performance is assessed using quantitative metrics such as electrical angle estimation error, speed tracking accuracy, and stability of dq-axis current regulation. These metrics directly reflect the suitability of the sensorless scheme for stable EPS motor operation.

Items Outside the Scope

The following aspects are intentionally not evaluated in this chapter:

- **Low-speed steering feel and on-center behavior**

Subjective steering characteristics and vehicle-level dynamics require a full steering system and vehicle model, which are beyond the scope of the present motor-drive-focused analysis.

- **System identification of an actual EPS unit**

The modeling does not attempt to identify or replicate the detailed nonlinear characteristics of a specific production EPS system.

- **Closed-loop low-speed sensorless enhancement**

Although saliency-based high-frequency injection (HFI) is implemented at the signal-processing level, its stable closed-loop integration for low-speed operation is not finalized within this study.

Consequently, HFI is treated as a **future extension** rather than a primary evaluation target.

By clearly defining what is and is not evaluated, this chapter avoids overgeneralization while maintaining relevance to EPS applications. The results presented herein establish the feasibility and limitations of sensorless IPMSM motor control at the drive level, providing a solid technical foundation for subsequent system-level integration.

7.2 Model and Overall System Configuration

7.2.1 MathWorks Reference Model Overview

The modeling framework used in this study is based on the MathWorks reference model `mcb_pmsm_foc_sensorless_f28379d`, which provides a validated implementation of sensorless field-oriented control (FOC) for permanent magnet synchronous motors (PMSMs).

This reference model includes a complete control and plant structure, consisting of:

- an inner dq-axis current control loop,
- an outer speed control loop,
- a sensorless position and speed estimation block,
- a three-phase inverter and SVPWM model,
- a motor plant model.

The primary advantage of using this reference model is that it represents a well-tested and stable baseline, in which the fundamental FOC architecture and signal scaling are already verified. This allows the present study to focus on the sensorless estimation and modeling aspects, rather than on reconstructing a motor control framework from scratch.

Although the original reference model is configured for a surface-mounted PMSM (SPMSM) and speed-controlled operation, its modular structure enables systematic modification and extension. In particular, the separation between the control subsystem, the inverter–motor plant, and the estimation blocks makes it suitable for research-oriented customization while preserving numerical stability and reproducibility.

Figure 7.1 illustrates the top-level structure of the reference model, highlighting the major subsystems for speed control, current control, and the inverter–motor plant.

required for sensorless evaluation and high-frequency injection (HFI) experiments. Importantly, these extensions do not alter the fundamental dq current control structure.

2. Explicit Exposure of Plant Signals for Evaluation

Motor phase currents and the true electrical rotor position (MtrPosElec) are explicitly routed to the estimation and evaluation subsystems. The true rotor position is used **only for offline comparison and evaluation**, and is not employed in the control loop, thereby preserving fully sensorless operation.

3. Separation of Control and Evaluation Paths

The generation of the control angle used for Park and inverse Park transformations is strictly based on observer outputs. Evaluation-related signals, including those used for error calculation and analysis, are kept separate from the control path to avoid contamination of the closed-loop behavior.

These design choices ensure that the modified model remains faithful to the original reference architecture while enabling detailed analysis of sensorless estimation performance.

Figure 7.2 shows the current control level of the model, illustrating how the original control structure is preserved and where evaluation-related signal paths are introduced.

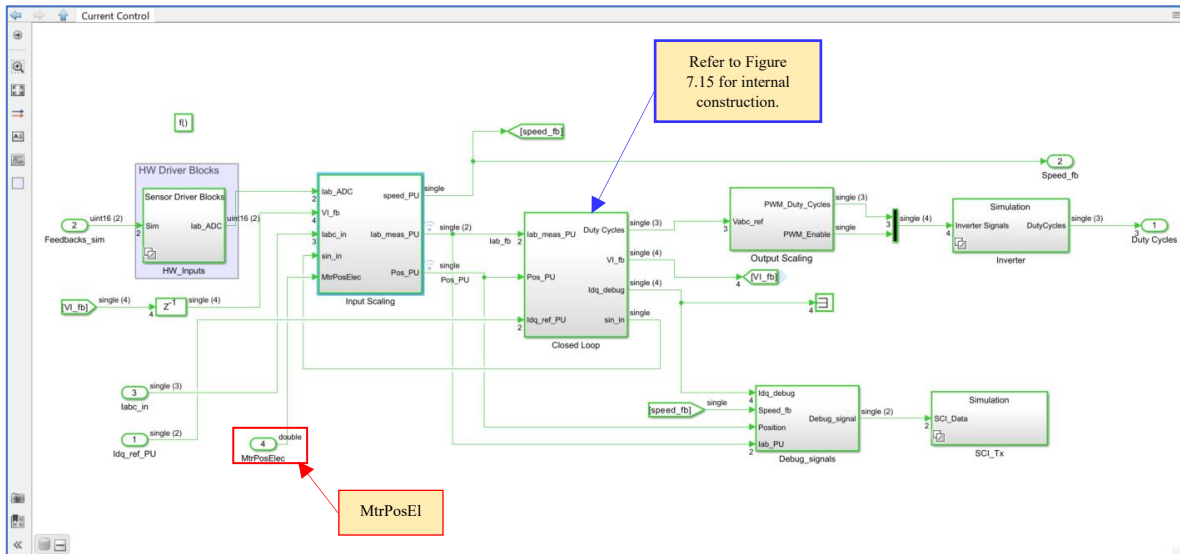


Figure 7.2 Current control level of the model

7.2.3 Position of the Modeling Framework within EPS Applications

Although the reference model employs a speed-controlled outer loop, this choice is consistent with the objectives of the present study. In EPS applications, the motor drive is ultimately required to deliver assist

torque; however, the inner FOC current control loop and the sensorless position estimation are common to both speed-controlled and torque-controlled architectures.

From a system engineering standpoint, the speed-controlled reference model serves as a testbed for validating the sensorless estimation and current control performance at the motor-drive level. Successful operation within this framework indicates that the essential building blocks required for EPS motor control—namely, accurate electrical angle estimation and stable dq current regulation—are in place.

The extension from speed control to assist torque control is therefore treated as a higher-level integration task, which is discussed later as future work. This layered approach allows the present chapter to focus on verifiable and quantifiable subsystem behavior, consistent with the overall system engineering methodology adopted in this thesis.

7.3 IPMSM Parameterization and PU System Consistency

7.3.1 Motivation for IPMSM Parameterization

Electric power steering (EPS) drive motors are commonly designed with salient rotor structures to achieve high efficiency and controllability over a wide operating range. To reflect this characteristic while retaining the validated structure of the reference model, the permanent magnet synchronous motor (PMSM) parameters were reconfigured to represent an interior PMSM (IPMSM), in which the d-axis and q-axis inductances differ ($L_d \neq L_q$).

The objective of this parameterization is twofold. First, it aligns the motor model with practical EPS motor characteristics. Second, it establishes a consistent modeling foundation for both back-EMF-based sensorless estimation (SMO) and potential saliency-based extensions such as high-frequency injection (HFI), which are relevant to low-speed EPS operation.

7.3.2 Datascript Design and Parameter Override Strategy

The motor, inverter, and control parameters are defined and initialized using a modified datascript, `mcb_pmsm_foc_sensorless_f28379d_datascript_corrected.m`, which is derived from the original MathWorks reference datascript.

In the reference implementation, motor parameters are loaded from a predefined SPMSM preset. In contrast, the present study introduces an explicit IPMSM override block, in which key electrical and mechanical parameters are reassigned prior to any derived calculations.

A critical design requirement is that the parameter override must occur before:

- inverter parameter updates,
- base speed computation,
- per-unit (PU) system definition, and
- controller and observer parameter tuning.

This ordering ensures that all subsequent calculations consistently reference the modified IPMSM parameter set.

7.3.3 Motor and Inverter Parameters

Table 7.1 summarizes the motor and inverter parameters used in the modeling and evaluation.

- Number of pole pairs: $p = 2$
- Stator resistance: $R_s = 0.08 \Omega$
- d-axis inductance: $L_d = 2.5 \text{ mH}$
- q-axis inductance: $L_q = 4.0 \text{ mH}$
- Permanent magnet flux linkage: $\psi_f = 0.06 \text{ Wb}$
- Load torque: $T_{\text{load}} = 0.136 \text{ N.m}$
- Rated current: $I_{\text{rated}} = 10 \text{ A}$
- Rotor inertia: $J = 1.0 \times 10^{-4} \text{ kg.m}$
- Viscous friction coefficient: $B = 2.5 \times 10^{-6} \text{ N.m.s}$

Table 7.1 Motor and inverter parameters

The inverter parameters are derived from the selected hardware configuration and updated using the modified motor parameters to maintain consistency between the electrical limits and the control design.

The parameter values listed in Table 7.1 were selected to represent a realistic EPS drive motor while maintaining consistency with the validated MathWorks reference model. In particular, the d- and q-axis inductances were chosen such that ($L_d \neq L_q$), enabling IPMSM characteristics relevant to both SMO-based estimation and potential saliency-based extensions.

The mechanical parameters J and B were selected to provide a reasonable dynamic response for evaluating load variation effects at the motor-drive level. The DC bus voltage was fixed at 24 V for the primary evaluation to ensure stable operation and reproducibility within the reference model framework.

7.3.4 Per-Unit System and Base Quantities

Following the parameter override, the per-unit system is established using the updated motor and inverter parameters. The base speed N_{base} is computed from the DC bus voltage and inverter limits, and all PU quantities are derived accordingly. By constructing the PU system after the IPMSM parameter override, the model guarantees that:

- current, voltage, speed, and torque quantities are scaled consistently,
- observer parameters are tuned against the correct base values, and
- numerical stability of the simulation is preserved.

This consistency is essential for reliable comparison of estimation accuracy and control performance across different operating conditions. Details would be discussed in the later section “7.5.1.1 Per-Unit System and Physical Interpretation”

7.3.5 Implications for Sensorless Estimation and Control

The corrected datascript design directly supports the validity of the subsequent sensorless evaluation. Because the IPMSM parameter override precedes all dependent computations:

- the sliding mode observer (SMO) parameters are derived from the same motor model used in the plant,
- the electrical angle conversion employs a single, well-defined transformation based on the correct number of pole pairs, and
- ambiguities related to parameter mismatch or unit inconsistency are eliminated.

As a result, any limitations observed in the sensorless estimation performance cannot be attributed to inconsistencies in the motor parameterization or PU system definition. This separation enables a clear interpretation of the evaluation results presented in later sections.

7.4 Sensorless Position Estimation Using Sliding Mode Observer

7.4.1 Role of Sensorless Position Estimation in EPS Motor Control

In field-oriented control (FOC) of permanent magnet synchronous motors, accurate knowledge of the rotor electrical position is essential for correct Park and inverse Park transformations. In EPS applications, eliminating mechanical position sensors is attractive from the viewpoints of cost, reliability, and robustness. However, this places strict requirements on the accuracy and stability of sensorless position estimation.

The objective of this section is to describe the sensorless position estimation framework adopted in this study, with particular emphasis on the sliding mode observer (SMO) used for medium-to-high speed operation. The focus is on how the observer is integrated into the reference model and how its outputs are utilized for control, while preserving fully sensorless operation.

7.4.2 Principle of Sliding Mode Observer in the Medium-to-High Speed Region

Sliding mode observers estimate the rotor position by exploiting the back-electromotive force (back-EMF) induced in the stator windings. In the medium-to-high speed region, the back-EMF magnitude is sufficiently large to ensure observability, making SMO-based estimation both theoretically sound and practically robust.

The key characteristics of the SMO approach are:

- robustness against parameter variations and disturbances,
- fast convergence due to the switching nature of the observer, and
- suitability for real-time implementation in embedded motor control systems.

Because back-EMF is proportional to electrical speed, SMO performance degrades at very low speeds. This limitation is well known and motivates alternative approaches, such as saliency-based methods, for low-speed operation. In the present study, this theoretical limitation is explicitly acknowledged, and the evaluation scope is restricted to the operating region where SMO is expected to be valid.

7.4.3 Implementation within the Model Framework

The sensorless estimation is implemented within the Figure 7.3 “Calculate position and speed” subsystem of the reference model. An estimator selection mechanism is provided, allowing different observer types (SMO, flux observer, extended EMF observer) to be selected. In this study, the SMO configuration is used for all evaluations.

The “SMO” subsystem of the reference model receives stator voltage and current information in the stationary reference frame and generates estimates of:

- the rotor electrical position, and
- the rotor electrical speed.

The estimated electrical position is used exclusively to generate the sine and cosine signals required for the Park and inverse Park transformations in the FOC scheme. Importantly, the true rotor electrical position (MtrPosElec) provided by the motor plant is not used for control purposes. It is routed only to evaluation and

analysis paths, ensuring that the control loop remains fully sensorless. This design choice preserves the integrity of the sensorless evaluation and prevents artificial performance improvement due to the use of ground-truth information.

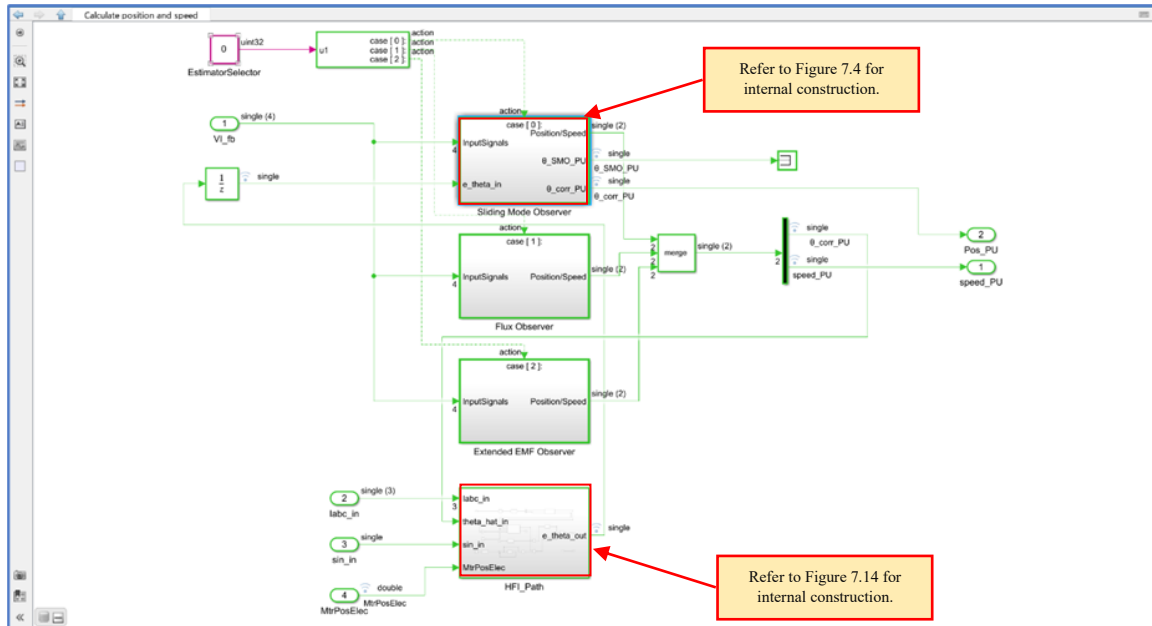


Figure 7.3 Calculate position and speed of the model

7.4.4 Separation of Control and Evaluation Paths

A key design aspect of the modeling framework is the strict separation between control paths and evaluation paths.

- The control path relies solely on observer-generated signals for position and speed.
- The evaluation path uses the true electrical position only for offline comparison and error analysis.

This separation ensures that any observed control performance reflects the intrinsic capability of the sensorless estimator rather than unintended signal leakage. As a result, the estimation accuracy assessed in later sections can be directly interpreted as the performance achievable in a practical sensorless EPS motor drive.

7.4.5 Position Correction Interface and Design Considerations

The reference model architecture allows for the introduction of additional correction signals to the estimated position. In this study, an interface is provided for position correction signals, enabling experimental investigation of enhancement methods such as high-frequency injection (HFI).

Within the sliding mode observer block, a correction signal can be added to the estimated electrical position through a simple proportional structure. While this interface facilitates research extensions, it does not alter the baseline SMO estimation used for the primary evaluation in this chapter.

For the purposes of the present study, the baseline SMO output is treated as the primary control angle. Correction-based extensions are evaluated separately and are discussed later as future work, consistent with the defined scope of this chapter.

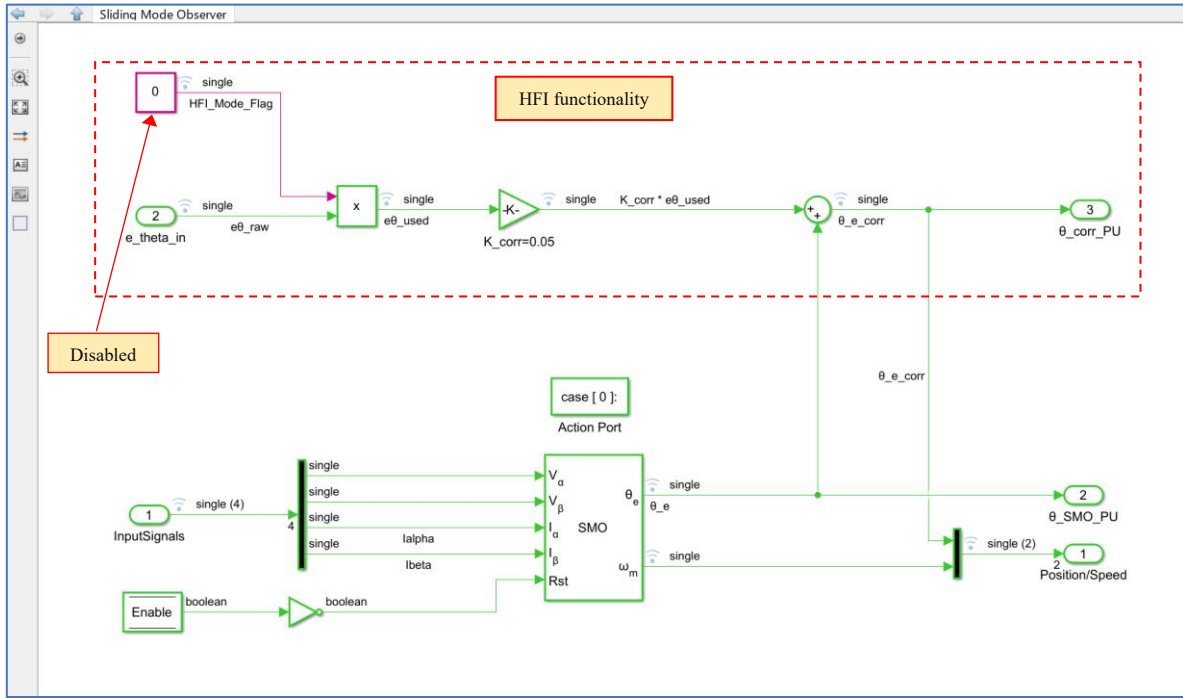


Figure 7.4 SMO block and its external signal interfaces (HFI)

7.4.6 Summary of Sensorless Estimation Framework

In summary, the sensorless position estimation framework employed in this study:

- uses a sliding mode observer for medium-to-high speed operation,
- is fully integrated within a validated reference model,
- preserves sensorless operation by isolating true position signals from the control loop, and

- provides a clean interface for future low-speed enhancement methods.

This framework forms the basis for the quantitative evaluation presented in the following section, where the estimation accuracy and control performance are assessed under representative operating conditions.

7.5 Evaluation Plan

7.5.1 Evaluation Scope and Alignment with EPS Applications

The evaluation methodology in this chapter is designed to assess the motor-drive-level performance of a sensorless IPMSM control system in a manner consistent with EPS application requirements, while respecting the limited scope and time constraints of this study.

From an EPS system perspective, actuator performance must satisfy multiple criteria, including dynamic response, smoothness, thermal robustness, and safety. In the present work, the evaluation scope is intentionally restricted to **two primary metrics** that directly reflect the effectiveness of sensorless position estimation and FOC operation:

1. Tracking and dynamic response, and
2. Smoothness of torque production, evaluated through a current-based proxy.

Other EPS-relevant aspects—such as thermal behavior, on-center steering feel, and functional safety—are acknowledged but treated as out of scope for quantitative evaluation and are discussed qualitatively.

7.5.1.1 Per-Unit System and Physical Interpretation

In this study, all simulation variables are expressed in per-unit (PU) values. The per-unit system normalizes physical quantities with respect to predefined base values, enabling consistent scaling across the model and facilitating comparison of control performance. The relationship between PU values and physical quantities is defined as:

$$x_{actual} = x_{PU} \times x_{base}$$

where x_{base} represents the corresponding base value of each variable.

For the present model, the base values as *conversion examples* are defined as follows:

Variable	PU Value	Physical Value	Base Value (PU=1.0)
Speed	1.0 PU	464 rpm	N_base = 464 rpm
Speed	0.6 PU	278.4 rpm	-

q-axis current	1.0 PU	10 A	$I_{\text{base}} = 10 \text{ A}$
q-axis current	0.2046 PU	2.05 A	-
Load torque	1.0 PU	0.272 N·m	$T_{\text{base}} \approx 0.272 \text{ N·m}$
Load torque	0.05–0.50 PU	0.0136–0.136 N·m	-
DC bus voltage *2	-	24 V	$V_{\text{dc}} = 24 \text{ V}$

Table 7.2 Examples for Per-unit to physical quantity conversion in the present model

*2 In this study, all electrical variables are expressed in per-unit (PU) values. The voltage is also normalized within the model; however, its base value is not directly equal to the DC bus voltage. Instead, the voltage base is implicitly determined by the inverter constraints and the DC bus voltage level ($V_{\text{dc}} = 24 \text{ V}$). Therefore, while all computations are performed in PU, the corresponding physical voltage is bounded by the inverter output capability derived from the DC bus.

7.5.2 Test Cases and Operating Conditions

Three simulation test cases are defined to evaluate the robustness and repeatability of the sensorless control scheme under representative operating conditions.

Test Case 1: Nominal Condition

- DC bus voltage: 24 V
- Motor and controller parameters: nominal values defined in Table 7.1
- Speed disturbance application time: $t = 4.0 \text{ s}$
- Objective: establish baseline performance of sensorless control in the medium-to-high speed region

Test Case 2: Load Variation Condition

- DC bus voltage: 24 V
- Speed disturbance application time: $t = 4.0 \text{ s}$
- Increased Motor Load Torque Condition ($T_L \times 3$)
- Objective: evaluate robustness of tracking performance and smoothness under increased load

All other parameters and control settings are kept identical between the two test cases to ensure a fair comparison.

Test Case 3: Strong Load Torque Disturbance Condition

To further assess actuator-level robustness under realistic EPS-related disturbance conditions, disturbance rejection test was conducted. In this case, a step-like external load torque disturbance was applied after the drive had already reached steady operation in the medium-to-high speed region.

The disturbance condition was defined as follows:

- Initial load torque: 0.05PU Load_Trq
- Final load torque: 3.0PU Load_Trq
- Torque disturbance application time: $t = 4.8$ s
- Load-torque low-pass filter: bypassed

This configuration was selected to represent a stronger actuator-side disturbance than the previously evaluated steady load variation case, while preserving the focus on motor-drive-level behavior. The test is intended to emulate abrupt external torque changes relevant to EPS actuator environments, such as road disturbance, rack friction variation, or steering torque spikes. [7-1], [7-2], [7-3]

7.5.3 Logged Signals

The following signals are logged during simulation using the Simulink Data Inspector (SDI):

- **Rotor electrical position**
 - True position: MtrPosElec (Not presented in the waveform)
 - Estimated position: θ_SMO_PU (Not presented in the waveform)
- **Speed**
 - Speed command: Spd_filt
 - Measured speed: Speed_Meas_PU
 - Estimated speed: speed_PU (Not presented in the waveform)
- **Currents**
 - d-axis current: Id_fb
 - q-axis current: Iq_fb
- **Torque**
 - Motor load torque: Load_Trq

These signals are sufficient to evaluate tracking performance, estimation accuracy, and torque smoothness at the motor-drive level.

7.5.4 Evaluation Window Definition

To exclude startup transients and operating regions where back-EMF-based sensorless estimation is theoretically unreliable, a dedicated evaluation window is defined. The evaluation window satisfies the following conditions:

- **Speed-based condition:**

$$\text{Speed_Meas_PU} \in [0.10, 0.60]$$

- **Time-based condition:**

$$t \geq 1.0 \text{ s}$$

All metrics are computed exclusively within this window. This definition ensures that the evaluation focuses on the **medium-to-high speed region**, where SMO-based estimation is expected to be valid.

7.5.5 Evaluation Metrics

7.5.5.1 Tracking and Dynamic Response

Tracking performance is evaluated by comparing the speed command to the measured speed within the evaluation window.

The primary metric is the root-mean-square (RMS) tracking error, defined as:

$$e_{\omega}(t) = \omega_{\text{meas}}(t) - \omega_{\text{cmd}}(t)$$

Additional peak error metrics may be reported to capture transient deviations. This metric reflects the ability of the sensorless control system to provide stable and accurate actuation in response to command changes, which is a fundamental requirement for EPS motor operation.

7.5.5.2 Smoothness Metric Based on Current Ripple

In EPS applications, smooth torque production is essential to avoid vibration and degraded steering feel.

Under field-oriented control, the electromagnetic torque of an IPMSM is predominantly determined by the q-axis current when the d-axis current is regulated near its target value.

Accordingly, q-axis current ripple is used as a proxy metric for torque ripple in this study.

The smoothness metric is defined as:

- RMS value of the q-axis current ripple within the evaluation window, and
- optionally, the peak-to-peak variation of I_q .

To support the validity of this proxy, the d-axis current I_d is monitored to confirm that it remains close to its reference value and does not exhibit significant oscillations.

7.5.5.3 Disturbance Rejection Metric

In addition to tracking performance and torque smoothness, disturbance rejection performance is evaluated as a third actuator-level metric as CASE3. This additional axis is motivated by EPS literature, in which the controller is commonly required not only to track the desired assist current derived from the characteristic curve, but also to maintain acceptable performance in the presence of external disturbances and plant uncertainty.

In prior EPS control studies, the control problem is explicitly framed as a tracking problem under disturbance and uncertainty, and control methods are compared in terms of performance, robustness, and disturbance rejection. From that perspective, disturbance rejection is not an auxiliary metric but a necessary core actuator-level requirement.

Accordingly, a standardized strong load torque disturbance test is introduced in this study. A step-like external load torque is applied after the speed loop has already reached steady operation in the medium-to-high speed region. The following scalar metrics are used:

- peak speed deviation after disturbance,
- RMS speed error in the disturbance window,
- settling behavior with respect to the predefined tolerance band,
- integral absolute error (IAE), and
- q-axis current response / ripple characteristics.

This evaluation is intended to represent realistic actuator-side external perturbations relevant to EPS operation, such as road disturbance, rack friction variation, and steering torque spikes.

7.5.6 Automation and Reproducibility

To ensure reproducibility and consistency across test cases, all evaluation metrics are computed using an automated post-processing script.

The evaluation procedure is as follows:

1. Run the simulation under the specified test condition.
2. Log the predefined signals using SDI.
3. Extract the evaluation window based on speed and time conditions.
4. Compute tracking and smoothness metrics within the window.
5. Repeat the procedure for each test case and compare results.

This automated workflow guarantees that identical evaluation criteria are applied to all simulations, enabling objective comparison between operating conditions.

7.5.7 Scope Limitations

The evaluation methodology deliberately excludes quantitative assessment of:

- thermal losses and temperature rise,
- on-center steering feels and vehicle-level dynamics, and
- functional safety and diagnostic coverage.

These aspects require additional modeling layers and are therefore addressed only qualitatively or reserved for future work.

7.6 Evaluation Results

This section presents the evaluation results of the proposed sensorless IPMSM drive system based on the methodology defined in Section 7.5. The evaluation focuses on actuator-level performance relevant to EPS applications, specifically tracking performance and torque smoothness, under both nominal and increased load torque conditions.

All results are computed within the predefined medium-to-high speed evaluation window, where sliding mode observer (SMO)–based estimation is theoretically reliable.

7.6.1 CASE1 Nominal Condition (24 V)

7.6.1.1 Tracking Performance

Figure 7.5 illustrates the speed tracking performance under nominal operating conditions with constant Motor load torque: Load_Trq = 0.5PU. The commanded speed and the measured motor speed are compared within the evaluation window.

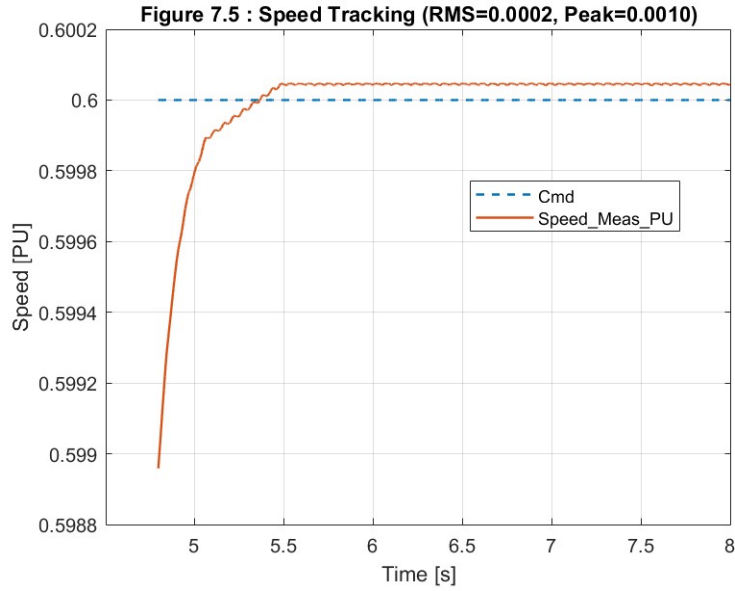


Figure 7.5 Speed command and measured speed under nominal conditions

The RMS tracking error in the evaluation window is 0.000152 PU, and the peak tracking error is 0.001041 PU. These results demonstrate that the sensorless drive achieves stable and accurate speed tracking under nominal load conditions.

7.6.1.2 Smoothness Evaluation Using q-axis Current Ripple

Figure 7.6 shows q-axis current waveform and ripple under nominal conditions. The upper plot shows the q-axis current, while the lower plot represents its ripple component after removing the mean value.

This allows independent assessment of torque magnitude and smoothness.

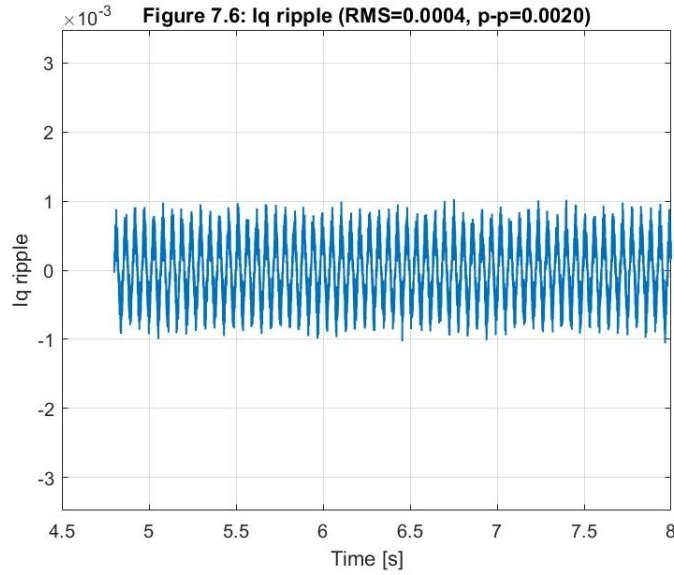


Figure 7.6 The q-axis current waveform and its ripple component under nominal conditions.

The RMS value of the q-axis current ripple is 0.000412 PU, and the peak-to-peak ripple is 0.001995 PU.

The $I_{qMean} = -0.204549$ PU and the d-axis current remains regulated lower throughout the evaluation window, supporting the validity of using q-axis current ripple as a proxy for torque ripple under field-oriented control.

7.6.2 CASE2 Increased Load Torque Condition ($TL \times 3$)

7.6.2.1 Tracking Performance under Increased Load

Figure 7.7 presents the speed tracking performance when the external load torque is increased the constant Motor load torque: $Load_Trq = 1.5$ PU by a factor of three from CASE1.

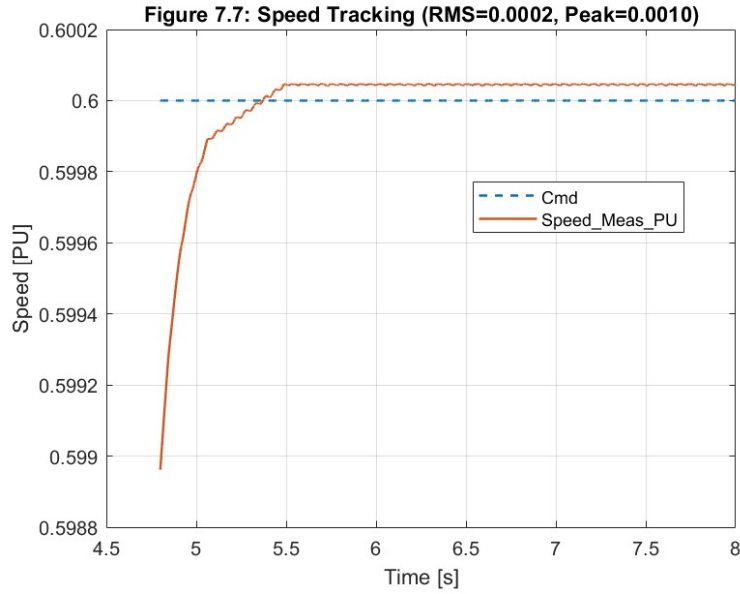


Figure 7.7 Speed command and measured speed under increased load torque

The RMS tracking error under this condition is 0.000152 PU, with a peak tracking error of 0.001038 PU. These values are nearly identical to those obtained under nominal conditions, indicating that the increased load torque does not significantly degrade tracking performance in the evaluated operating region.

7.6.2.2 Smoothness under Load Torque Variation

Figure 7.8 shows the q-axis current waveform under the increased load torque condition.

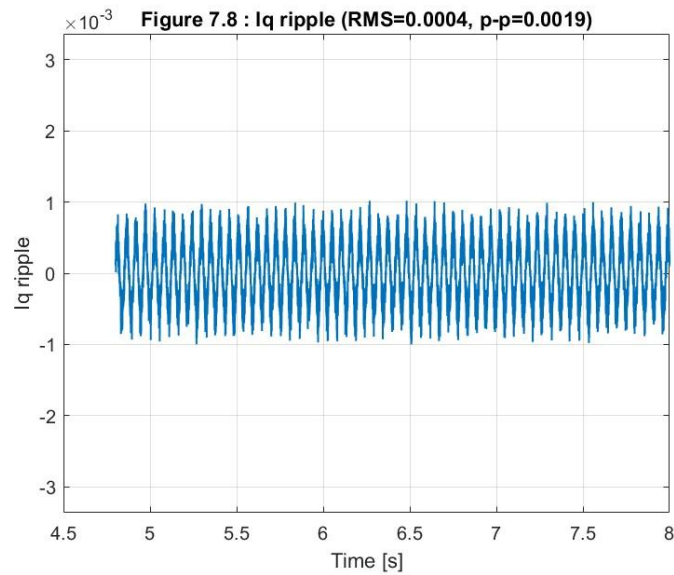


Figure 7.8 q-axis current waveform and ripple under increased load torque

While the increased load torque requires a higher mean q-axis current to satisfy the torque demand, the ripple around the mean remains comparable to the nominal case. The RMS q-axis current ripple is 0.000409 PU, and the peak-to-peak ripple is 0.001943 PU, indicating that the increased load torque does not significantly affect results from CASE1. As in the CASE1, the d-axis current remains regulated lower, confirming consistent field-oriented operation. Figure 7.9 shows Inverter and Motor Plant Model where external torque was adjusted.

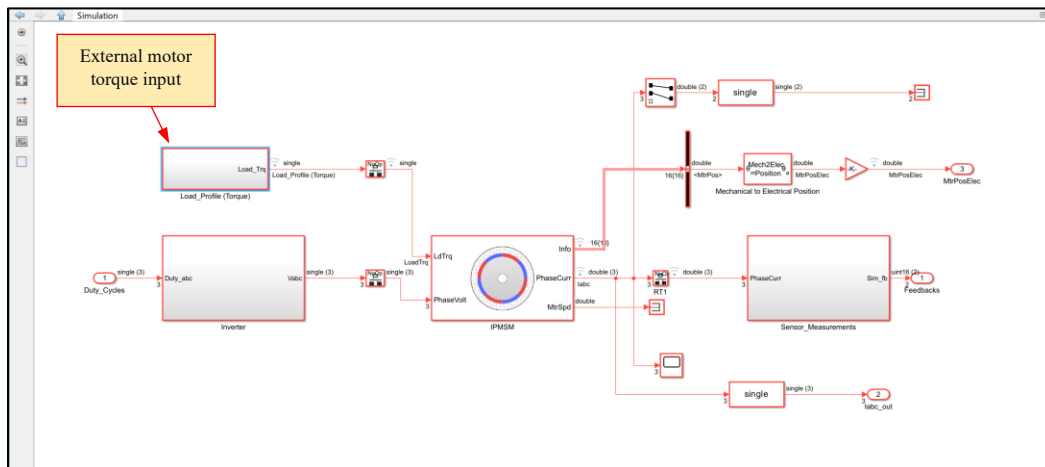


Figure 7.9 Inverter and Motor Plant Model

7.6.3 Strong Load Torque Disturbance Condition

To further evaluate actuator robustness, a strong load torque disturbance case was introduced after the motor had already reached steady operation in the medium-to-high speed region. In this test (Test Case 3), the external load torque was increased from 0.05 PU Load_Trq to 3.0 PU Load_Trq at $t = 4.8$ s, and the load-torque low-pass filter was bypassed so that the plant received a direct step-like disturbance input.

The inclusion of this test is consistent with EPS-oriented control literature. Prior studies have framed EPS control as a tracking problem under disturbance and uncertainty, and have evaluated control designs in terms of their ability to maintain performance and reject external perturbations. In those studies, the controller is required to preserve the desired assist behavior even when the steering system is subjected to road-related disturbances or parameter variations. From that perspective, the present disturbance test is an actuator-level analogue of the disturbance-rejection evaluations used in EPS research.

Figure 7.10 shows the overall speed response under the strong disturbance condition. The applied disturbance profile is shown in Figure 7.13. Although a large external torque step is applied at $t = 4.8$ s, the measured speed remains close to the command and does not exhibit visible steady-state deterioration. Refer to Figure 7.11 for the speed error.

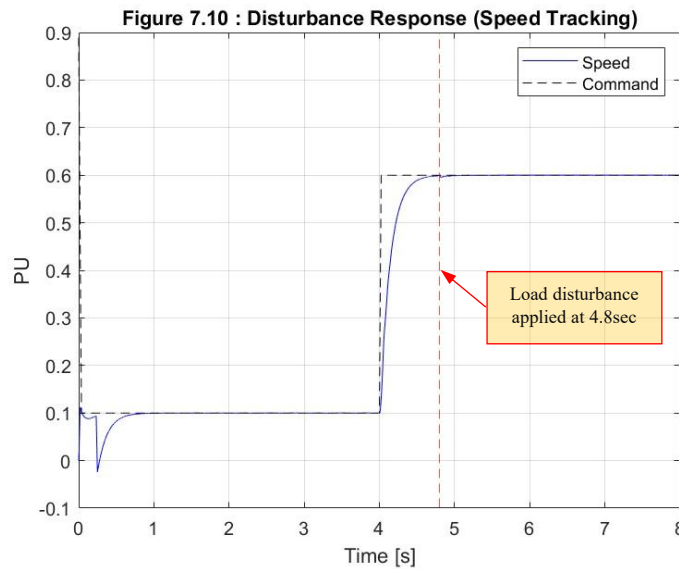


Figure 7.10 Speed command and measured speed under the strong load torque disturbance condition

Figure 7.11 shows the speed error in the disturbance window. The peak deviation is 0.0046 PU and the RMS error is 0.0012 PU. These values indicate that the disturbance-induced speed deviation remains on the order of 10^{-3} PU even under the stronger torque step.

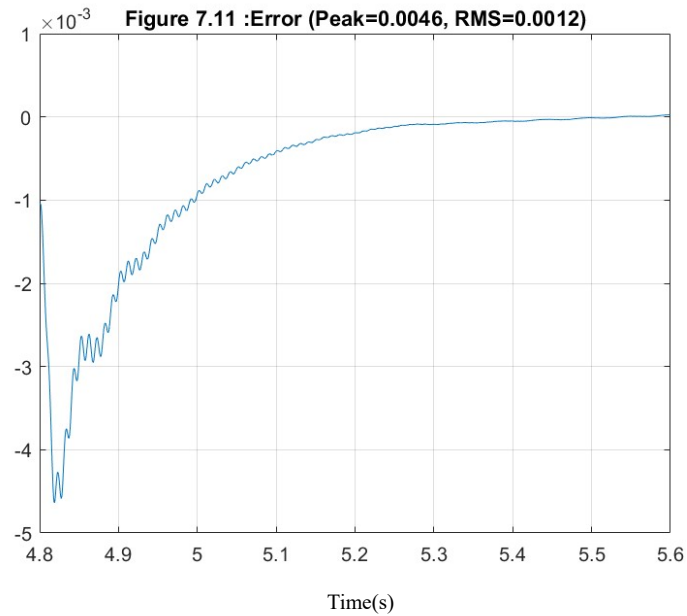


Figure 7.11 Speed error in the disturbance window under the strong load torque disturbance condition

Figure 7.12 shows the q-axis current response in the disturbance window. Immediately after the disturbance, the q-axis current changes to generate the additional electromagnetic torque required for compensation, and the response is subsequently damped toward a new steady-state level. The q-axis current ripple RMS and peak-to-peak values are 0.003752 PU and 0.04742 PU, respectively.

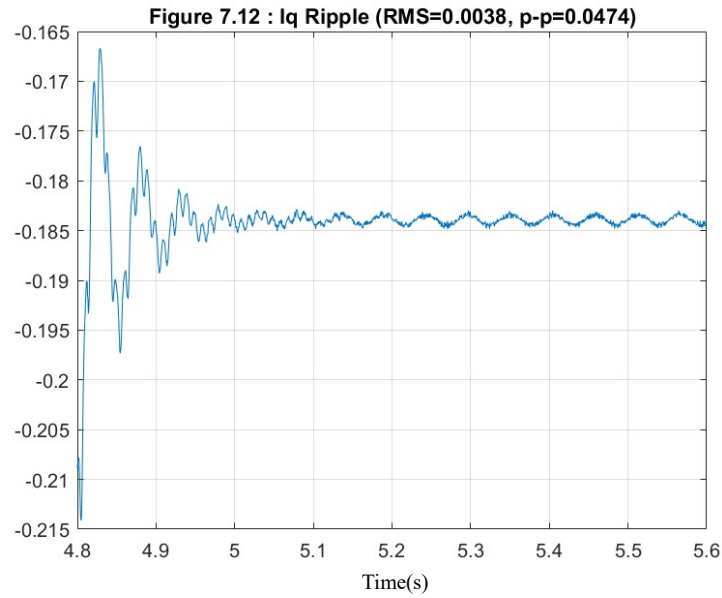


Figure 7.12 q-axis current response and ripple in the disturbance window under the strong load torque disturbance condition

Figure 7.13 confirms the disturbance injection itself. Because the load-torque filter was bypassed in this case, the commanded load torque step is directly reflected in the plant input, validating that the observed response corresponds to a genuine step disturbance rather than a filtered load variation.

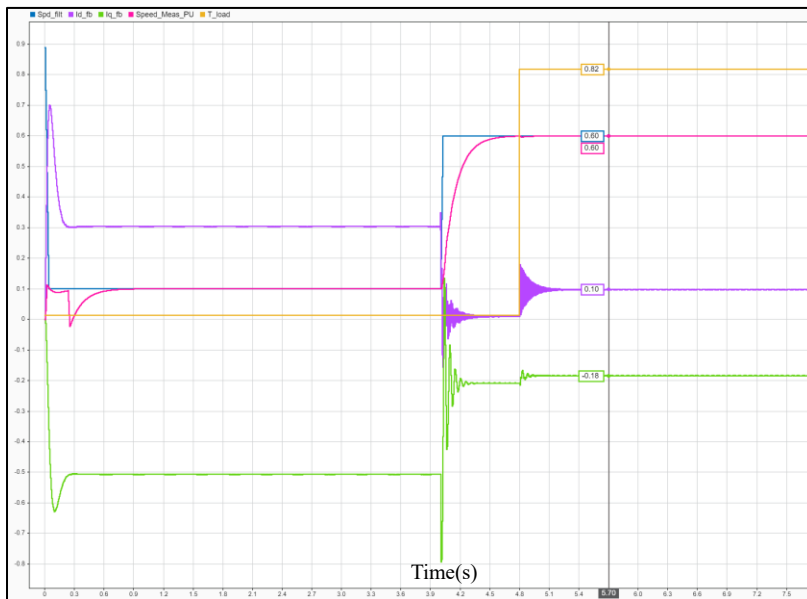


Figure 7.13 Applied load torque step and plant load torque under the strong load torque disturbance condition

Table 7.3 shows that the resulting CASE1-3 MATLAB metrics are summarized.

Case 1

Command Window		
==== CASE Evaluation Metrics === Run Name = MH_R1_24V_nominal Evaluation window = 4.800 to 8.000 s RMS_tracking = 0.000152 PU Peak_tracking = 0.001041 PU IqRippleRMS = 0.000412 IqRipplePP = 0.001995 MeanAbs_Id = 0.022584 PeakAbs_Id = 0.023563 --- Copy/paste to Table X.5 --- RMS_tracking=0.000152, Peak_tracking=0.001041, IqRippleRMS=0.000412, IqRipplePP=0.001995 IqMean = -0.204549 MeanAbs_Iq = 0.204549 MeanAbs_Id = 0.022584 PeakAbs_Id = 0.023563 fx >>		
Workspace		
Name -	Value	
cmd	16001x1 single	
cmd_win	6401x1 single	
e_tracking	6401x1 single	
fig1	1x1 Figure	
fig2	1x1 Figure	
fig3	1x1 Figure	
fig4	1x1 Figure	
getSig	@(name)Simulink.sdi.get...	
half_span	0.0035	
id	16001x1 single	
id_win	6401x1 single	
IdMeanAbs	0.0226	
IdPeakAbs	0.0236	
idx	16001x1 logical	
iq	16001x1 single	

Case 2

Command Window		
==== CASE Evaluation Metrics === Run Name = MH_R2_24V_TLx3 Evaluation window = 4.800 to 8.000 s RMS_tracking = 0.000152 PU Peak_tracking = 0.001038 PU IqRippleRMS = 0.000409 IqRipplePP = 0.001943 MeanAbs_Id = 0.052080 PeakAbs_Id = 0.053032 --- Copy/paste to Table X.5 --- RMS_tracking=0.000152, Peak_tracking=0.001038, IqRippleRMS=0.000409, IqRipplePP=0.001943 IqMean = -0.195699 MeanAbs_Iq = 0.195699 MeanAbs_Id = 0.052080 PeakAbs_Id = 0.053032 fx >>		
Workspace		
Name -	Value	
cmd	16001x1 single	
cmd_win	6401x1 single	
e_tracking	6401x1 single	
fig1	1x1 Figure	
fig2	1x1 Figure	
fig3	1x1 Figure	
fig4	1x1 Figure	
getSig	@(name)Simulink.sdi.get...	
half_span	0.0034	
id	16001x1 single	
id_win	6401x1 single	
IdMeanAbs	0.0521	
IdPeakAbs	0.0530	
idx	16001x1 logical	
iq	16001x1 single	

Case 3

Command Window		
==== Run3 Disturbance Rejection Metrics === Peak deviation = 0.004636 PU RMS error = 0.001247 PU Settling time = 0.000000 s IAE = 0.000562 Iq ripple RMS = 0.003752 Iq ripple peak2peak = 0.047429 fx >>		
Workspace		
Name -	Value	
band	0.0120	
cmd	16001x1 single	
cmd_win	1600x1 single	
err	1600x1 single	
getSig	@(name)Simulink.sdi.get...	
IAE	5.6165e-04	
id	16001x1 single	
idx	16001x1 logical	
iq	16001x1 single	
iq_mean	-0.1841	
iq_PP	0.0474	
iq_ripple	1600x1 single	
iq_RMS	0.0038	
iq_win	1600x1 single	
k	1	
Id	16001x1 single	
MAKE_PLOTS	1	
Peak_dev	0.0046	
r	1x1 Run	
ref	0.6000	
RMS_err	0.0012	
RUN_NAME	"MH_R3_24V_Tlstep"	
runID	4	
SAVE_PLOTS	0	
settle_idx	1	

Table 7.3

It should be noted that in CASE3 the reported settling time of 0.000000 s does not imply physically instantaneous recovery. Rather, it means that the disturbance-induced deviation remained within the predefined tolerance band throughout the evaluated disturbance window. Therefore, the result should be interpreted as strong disturbance attenuation rather than zero dynamic effect. Taken together, these results indicate that the proposed SMO-based sensorless FOC system maintains stable actuator-level behavior even under a strong step-like load torque disturbance.

7.6.4 Comparative Analysis of Nominal, Increased Load, and Disturbance Conditions

Case	Run Name	Eval Window [s]	Load Torque		RMS Speed Error		Peak Speed Error	
			PU	N-m	PU	RPM	PU	RPM
Case 1	MH_R1_24V_nominal	4.000–8.000	0.5	0.136	0.000152	0.070528	0.001041	0.483024
Case 2	MH_R2_24V_TLx3	4.000–8.000	1.5	0.408	0.000152	0.070528	0.001038	0.481632
Case 3	Run3 Disturbance	4.800–8.001	0.05→3.0	0.0136→0.81600000	0.001247	0.578608	0.004636	2.151104

Case	Settling Time [s]	IAE	Iq Mean		Iq Ripple RMS	
			PU	A	PU	A
Case 1			-0.204549	-2.04549	0.000412	0.00412
Case 2			-0.195699	-1.95699	0.000409	0.00409
Case 3	0	0.000562	-0.18	-1.8	0.003752	0.03752

Case	Iq Ripple p-p		Mean [Id]		Peak [Id]	
	PU	A	PU	A	PU	A
Case 1	0.001995	0.01995	0.022584	0.22584	0.023563	0.23563
Case 2	0.001943	0.01943	0.05208	0.5208	0.053032	0.53032
Case 3	0.047429	0.47429				

Table 7.4 Summary of evaluation metrics for Nominal, Increased Load, and Disturbance Conditions (PU normalized quantity vs Physical quantity).

The comparative results for all CASE1-3 MATLAB metrics are in Table7.4. As for CASE2, it shows that increasing the external load torque by a factor of three has negligible impact on tracking performance and does not significantly affect torque smoothness, as evaluated by q-axis current ripple. This behavior indicates that the proposed SMO-based sensorless control maintains robust operation against load torque disturbances in the medium-to-high speed region.

When the nominal case (CASE1), the increased steady-load case(CASE2), and the strong disturbance (CASE3) are considered together, the results show that the proposed control system maintains stable actuator-level performance over a broader range of operating conditions than would be demonstrated by nominal tracking alone.

Under the nominal and increased steady-load conditions, the tracking and smoothness metrics remain nearly unchanged, indicating that a steady increase in mechanical load does not significantly degrade performance in the evaluated operating region.

Under the strong disturbance condition (CASE3), the response exhibits a larger deviation than in the two steady-load cases, as expected. However, the resulting speed error still remains in the 10^{-3} PU range, and the q-axis current response remains stable and well damped. This indicates that the controller compensates for abrupt external torque changes primarily through torque-producing current regulation while preserving stable sensorless operation.

This interpretation is consistent with EPS control literature, in which tracking quality and robustness under disturbance / uncertainty are treated together rather than independently. Therefore, the present results support not only nominal actuator accuracy but also actuator robustness against abrupt load torque perturbations in the medium-to-high speed region.

7.6.5 Consistency with EPS-Oriented Evaluation Criteria

The evaluation results are consistent with actuator-level performance expectations commonly applied to EPS systems:

- Tracking performance reflects the ability of the actuator to respond accurately and stably to command inputs, a fundamental requirement in EPS applications.
- Torque smoothness, evaluated through q-axis current ripple, corresponds to torque ripple characteristics that directly influence steering feel and NVH behavior in EPS systems.
- Disturbance rejection performance reflects the actuator's ability to tolerate abrupt external torque perturbations, which are relevant to road disturbance, rack friction variation, and steering torque spikes.
- The evaluation window selection aligns with established theory indicating reliable SMO operation in the medium-to-high speed region.

Accordingly, the obtained results satisfy EPS-relevant actuator-level performance criteria within the defined scope of this study

7.7 Discussion

7.7.1 Interpretation of Tracking Performance Results

The evaluation results show that the tracking performance remains essentially unchanged between the nominal condition (**Test Case 1**) and the increased load torque condition (**Test Case 2**). Both the RMS and peak tracking errors are nearly identical across the two test cases, despite a threefold increase in external load torque.

This behavior can be interpreted by considering the hierarchical control structure of the drive system. The outer speed control loop regulates the motor speed based on the tracking error, while the inner current

control loop ensures fast and stable torque production. When the external load torque increases, the controller compensates primarily by increasing the mean q-axis current, thereby generating the additional electromagnetic torque required to maintain the commanded speed.

Importantly, the similarity of the tracking metrics between **Test Case 1** and **Test Case 2** indicates that, within the defined medium-to-high speed evaluation window, the sensorless position estimation does not become a limiting factor for speed control. This observation provides indirect evidence that the SMO-based position estimation remains sufficiently accurate under increased load torque conditions.

7.7.2 Interpretation of Smoothness Results Using q-axis Current Ripple

The smoothness evaluation based on q-axis current ripple reveals that the ripple characteristics remain comparable between the nominal and increased load torque cases. Although the mean (DC) q-axis current increases in response to the higher torque demand, the ripple around the mean does not exhibit a significant increase.

This result suggests that the additional torque demand is primarily accommodated through the DC component of the q-axis current, while the current control and modulation mechanisms continue to suppress high-frequency oscillations effectively. As a result, torque smoothness, which is closely related to current ripple under field-oriented control, is preserved even when the external load torque is increased.

Furthermore, the d-axis current remains regulated lower in both operating conditions. This consistent regulation reinforces the validity of using q-axis current ripple as a proxy for torque ripple, as the electromagnetic torque is predominantly determined by the q-axis current in this operating regime.

7.7.3 Robustness Against Load Torque Disturbances

The comparative results between **Test Case 1** and **Test Case 2** demonstrate that the proposed SMO-based sensorless control scheme exhibits robustness against load torque disturbances in the evaluated operating region.

Despite a substantial increase in external load torque, neither the tracking performance nor the smoothness metric shows noticeable degradation.

From an EPS actuator perspective, this robustness is a desirable characteristic, as the actuator must respond reliably to varying mechanical loads while maintaining stable and smooth operation. Although the present evaluation does not incorporate vehicle-level dynamics or steering feel, the actuator-level results indicate that the sensorless drive can tolerate load variations without compromising fundamental control performance.

It should be emphasized that the robustness observed here applies specifically to the medium-to-high speed region, where the observability conditions required by SMO-based estimation are satisfied. The results therefore support the applicability of the proposed sensorless control approach within this operating range.

The additional strong disturbance in **Test Case 3** case provides a clearer interpretation of actuator robustness than the steady-load comparison alone. In this case, a direct step load torque input was applied after the drive had already reached steady operation in the medium-to-high speed region, so that the observed response primarily reflects disturbance rejection behavior rather than command-step transient behavior.

Even under this stronger disturbance, the speed deviation remained on the order of 10^{-3} PU, and the response stayed within the predefined tolerance band throughout the evaluated disturbance window. At the same time, the q-axis current showed a physically reasonable compensation response: it changed rapidly after the disturbance, generated the additional torque required by the load increase, and then converged to a stable new operating level without instability or divergence.

This result is meaningful in the EPS context. Prior EPS studies have shown that actuator control is commonly formulated as motor-current tracking relative to a predefined EPS characteristic curve, while robustness under disturbance and uncertainty is treated as an equally important requirement. In that literature, external effects such as road excitation, friction variation, and model uncertainty are explicitly recognized as performance-limiting factors, and disturbance rejection is evaluated as a core control property.

The present disturbance test is consistent with that evaluation philosophy and demonstrates that the proposed sensorless control structure has meaningful robustness at the actuator level. Furthermore, because the d-axis current remains lower at the disturbance application point, the observed after steady-state q-axis current response can still be interpreted as a valid proxy for torque-generation behavior.

Therefore, the disturbance result strengthens the earlier conclusion that the proposed control system preserves not only tracking performance and torque smoothness, but also actuator-side disturbance robustness in the medium-to-high speed region.

7.7.4 Scope Limitations of the Present Evaluation

The discussion in this chapter is based on an evaluation framework that is intentionally limited in scope. The present study focuses on actuator-level behavior and does not address several aspects that are relevant to full EPS system performance, including low-speed and on-center steering feel, vehicle and rack dynamics, thermal behavior, efficiency, and functional safety.

These limitations are a direct consequence of the modeling scope and the objective to establish a clear and controlled evaluation of sensorless motor drive behavior. By restricting the evaluation to a well-defined

operating region and a limited set of performance metrics, the influence of load torque variations on sensorless control performance can be isolated and interpreted without confounding factors.

7.7.5 Implications for EPS Applications

Within the defined scope, the results of this study suggest that SMO-based sensorless control can provide stable and smooth actuator behavior under varying load torque conditions in the medium-to-high speed region. This finding is relevant to EPS applications, where the actuator is required to operate reliably under changing mechanical loads while maintaining consistent dynamic response.

Although direct conclusions regarding steering feel or vehicle-level performance cannot be drawn from the present evaluation, the actuator-level robustness demonstrated here supports the feasibility of sensorless control strategies as a component of future EPS system designs.

The disturbance rejection result obtained in the strong load torque disturbance case in **Test Case 3** further supports this applicability from an actuator robustness perspective.

7.7.6 Future Work

Several directions for future work are identified based on the limitations of the present study. First, the evaluation can be extended to the low-speed region by incorporating high-frequency injection techniques to enhance position observability. Second, the control framework can be expanded from speed control to torque (assist) control, which is more representative of practical EPS operation.

Finally, integration with vehicle and rack models would enable assessment of steering feel and NVH characteristics at the system level.

7.8 HFI Modeling: Achievements and Limitations

This section discusses the implementation of high-frequency injection (HFI)–based position estimation within the sensorless control framework and clarifies its achieved design scope and current limitations. Rather than replacing the existing sliding mode observer (SMO), HFI is designed as an auxiliary mechanism to enhance position observability in operating regions where back-EMF–based estimation becomes unreliable.

7.8.1 Implemented HFI Signal Processing Path

High-frequency injection is commonly employed to improve rotor position observability in low-speed regions, where back-EMF magnitude is insufficient for reliable observer-based estimation. In this study, an HFI signal processing path was implemented to complement the SMO-based estimation framework.

The implemented HFI path consists of the following functional blocks:

- Injection of a high-frequency voltage component: *cos, sin*
- Transformation of measured currents into the γ - δ reference frame: *GammaDeltaTransform*
- Synchronous demodulation using in-phase and quadrature components: *demod_sin, demod_cos*
- Low-pass filtering to extract position-related information, *num(z)/den(z)*
- Generation of an estimated position error signal: *e_theta_out*

Figure 7.14 illustrates the implemented *HFI_Path* subsystem. This figure shows the complete signal processing chain, from the injected excitation to the extracted position-related signal, as implemented in the Simulink model.

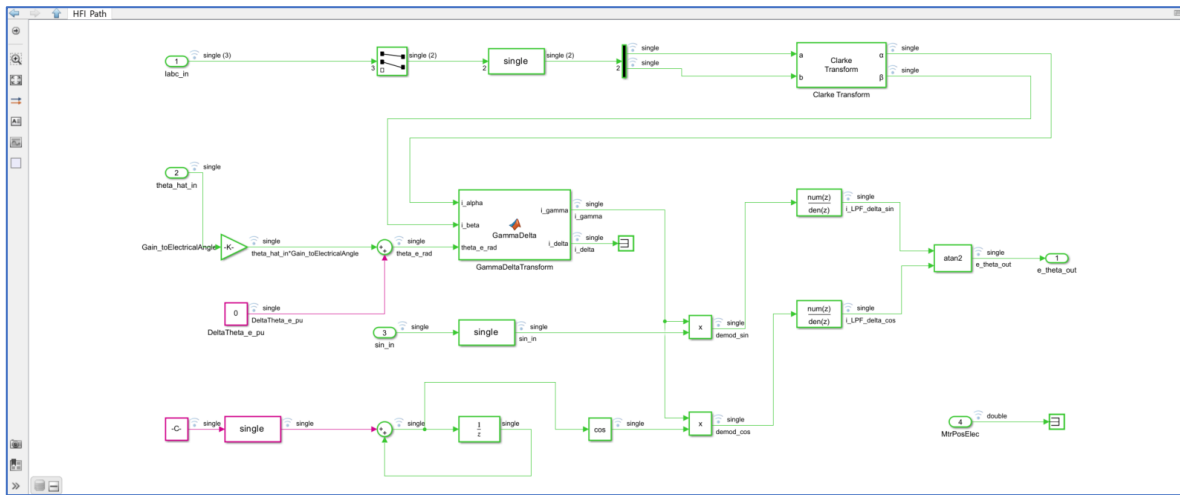


Figure 7.14 Implemented HFI signal processing path

The HFI signal processing path was implemented as an independent subsystem, allowing it to be enabled or disabled without modifying the core sensorless control structure. This design choice ensures that the baseline SMO-based control remains intact and can be evaluated independently, as presented in the previous section.

7.8.2 Integration with the Sensorless Control Loop

The HFI path was integrated into the sensorless control framework as a supplementary correction mechanism rather than as a standalone estimator. Specifically, the HFI-derived position-related signal is designed to act as an additional error input within the closed-loop position estimation structure.

Figure 7.15 shows the closed-loop integration point of the HFI path relative to the SMO output. The figure highlights how the HFI-related signal is combined with the SMO-estimated position in a feedback structure,

forming a unified sensorless estimation loop.

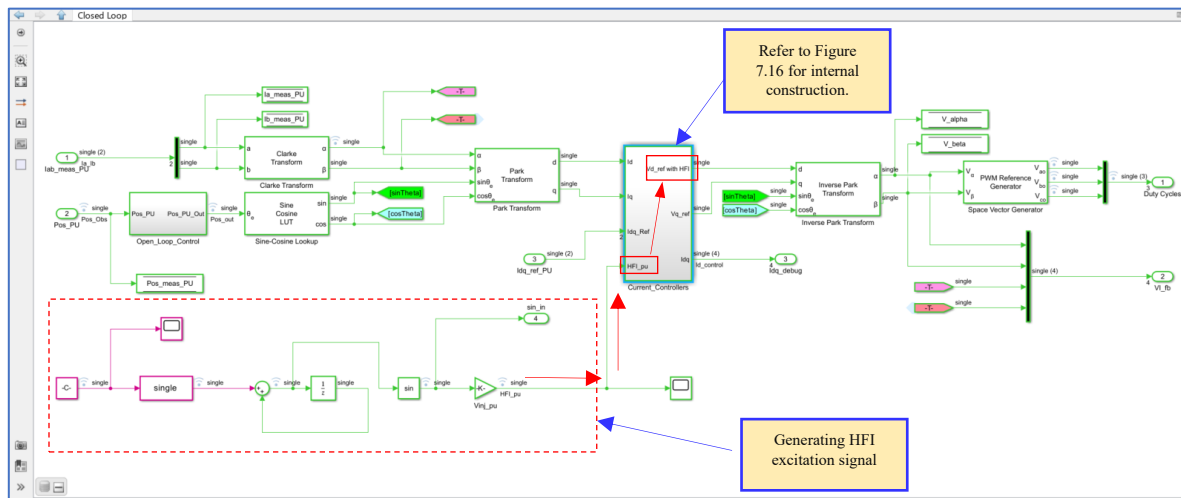


Figure 7.15 Closed-loop integration of SMO and HFI-based correction

In parallel, the HFI excitation is injected at the current control level in a manner that preserves the field-oriented control (FOC) structure. The injected signal is superimposed on the d-axis voltage reference, ensuring that the torque-producing q-axis current is not directly disturbed.

Figure 7.16 presents the current controller subsystem, indicating the location where the HFI voltage component is injected. This configuration maintains compatibility with standard dq current control while enabling HFI-based observability enhancement.

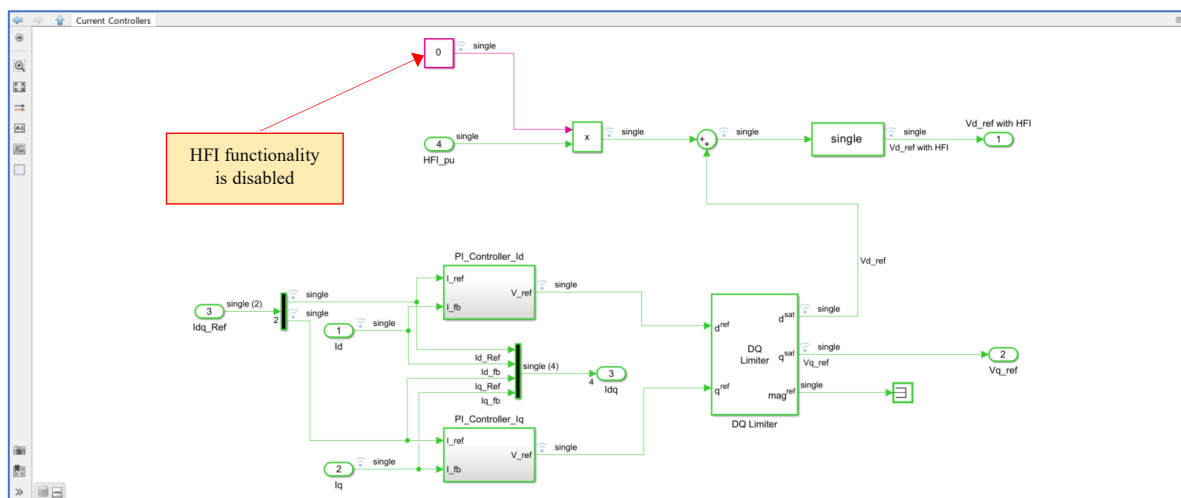


Figure 7.16 Current controller structure with HFI voltage injection

It is important to note that, for the evaluation results presented in Section 7.6, the HFI functionality was disabled. The SMO-based estimation was therefore assessed in isolation, ensuring that the reported performance reflects the baseline sensorless control capability without HFI assistance.

7.8.3 Observed Limitations in the Present Study

While the HFI signal processing path and its integration points were successfully implemented, the present study intentionally limited the scope of HFI evaluation. The primary limitation arises from the focus on medium-to-high speed operation, where SMO-based estimation is known to be effective. As a result, detailed closed-loop evaluation of HFI-assisted position estimation in the low-speed region was not conducted.

Additionally, although the closed-loop structure for combining SMO and HFI signals was implemented, a comprehensive stability and robustness analysis of this combined estimator was beyond the scope of this work. Such an analysis would require systematic tuning and validation under low-speed and standstill conditions, which were not prioritized in the current evaluation framework. These limitations reflect deliberate scope constraints rather than deficiencies in the implemented design.

7.8.4 Future Work

Based on the achieved implementation and observed limitations, several directions for future work are identified. First, a detailed evaluation of the HFI-assisted closed-loop estimator in the low-speed region should be conducted to quantify improvements in position estimation accuracy. Second, stability analysis and tuning of the combined SMO–HFI estimator should be addressed to ensure robust operation across the full speed range.

The control framework can be extended from speed control to torque (assist) control, which is more representative of practical EPS operation. Such an extension would enable direct assessment of steering assist performance and further integration with vehicle- and rack-level models.

Overall, the modeling and evaluation results confirm that the proposed SMO-based sensorless FOC system achieves stable and accurate actuator-level operation in the medium-to-high speed region.

The nominal and increased-load evaluations demonstrated that the system maintains accurate speed tracking and low q-axis current ripple under steady operating conditions. In addition, the strong load torque disturbance case showed that the system can tolerate an abrupt actuator-side torque perturbation while keeping the resulting speed deviation within the 10^{-3} PU order and maintaining stable current regulation.

These results collectively indicate that the proposed control structure satisfies the fundamental actuator-level performance requirements relevant to EPS applications, including tracking accuracy, torque smoothness, and disturbance robustness, within the defined modeling scope.

This interpretation is also consistent with EPS-oriented control literature, which commonly treats tracking performance and disturbance rejection as coupled requirements rather than independent evaluation axes.

While low-speed enhancement using HFI remains future work, the present work establishes a quantitatively validated and practically relevant foundation for sensorless EPS motor control in the medium-to-high speed region. [7-1], [7-2], [7-3]

Chapter 8. Conclusions

8.1 Summary of Technical Contributions

This thesis has investigated sensorless control for automotive Electric Power Steering (EPS) systems from a systems engineering perspective, with particular emphasis on the architectural implications of removing rotor position sensors and replacing them with a hybrid estimation framework.

The central technical contribution of this work is not merely the implementation of a sliding mode observer (SMO), but the structured evaluation of sensorless EPS as an architectural transformation rather than an algorithm substitution. The research integrates motor control theory, inverter hardware considerations, AUTOSAR-based software architecture, and functional safety principles into a coherent framework for assessing the feasibility of sensorless EPS deployment in sedan-class vehicles.

The key technical contributions can be summarized as follows:

1. Structured Comparison of Sensorless Estimation Strategies for EPS

A comprehensive comparison of four mainstream sensorless estimation approaches—Extended Kalman Filter (EKF), Sliding Mode Observer (SMO), Model Reference Adaptive System (MRAS), and Phase-Locked Loop (PLL)—was conducted under EPS-specific constraints.

Unlike traction-drive-oriented analyses, the evaluation criteria were tailored to steering systems, emphasizing:

- Low-speed observability requirements
- Deterministic real-time execution on automotive ECUs
- Robustness to parameter drift and inverter nonidealities
- Torque smoothness and NVH sensitivity
- Mass-production feasibility

This structured comparison demonstrated that, among back-EMF-based approaches, SMO provides the most favorable robustness-to-complexity tradeoff for sedan-class EPS applications, while low-speed operation requires saliency-based enhancement (e.g., HFI).

2. Architectural Framing of Sensorless EPS as a System-Level Change

Chapter 6 introduced an impact analysis across five subsystems:

- Mechanics and dynamics

- Electrical hardware (ECU and inverter)
- Control architecture (SMO + HFI hybrid)
- Software architecture (AUTOSAR-based integration)
- Functional safety and diagnostics

The transition from sensor-based to sensorless EPS was shown to:

- Convert rotor angle from a hardware boundary signal into a software-generated safety-relevant variable
- Increase dependency on signal integrity and inverter nonidealities
- Require explicit transition logic between low-speed (HFI) and medium/high-speed (SMO) operation
- Introduce new diagnostic and plausibility mechanisms for estimator validation

This systems-engineering framing is a primary contribution of the thesis: it demonstrates that sensorless control must be justified not only at the control-theory level, but across mechanical NVH, ECU hardware, AUTOSAR scheduling, and ISO 26262 safety arguments.

3. Development of a Simulation Framework Based on an IPMSM Sensorless FOC Architecture

A MATLAB/Simulink-based modeling framework was developed using a modified version of the MathWorks reference model *mcb_pmsm_foc_sensorless_f28379d*.

Key modifications included: IPMSM parameterization ($L_d \neq L_q$)

- Explicit separation of control and evaluation paths
- SMO-based rotor position estimation for medium-to-high speed
- Structured integration interface for future HFI enhancement

The modeling approach was deliberately scoped to motor-drive-level validation, consistent with the thesis objective of subsystem verification within a broader EPS systems context.

4. Quantitative Evaluation of SMO-Based Sensorless Control in the Medium-to-High Speed Region

Simulation results demonstrated:

- Stable speed tracking under nominal and increased load torque conditions

- Comparable RMS tracking error under $3\times$ load variation
- Low q-axis current ripple as a proxy for torque smoothness
- Robust operation within a defined medium-to-high speed evaluation window

Importantly, the study does not claim full-range validation.

The low-speed HFI channel was implemented structurally but not fully closed-loop validated within this work.

Thus, the validated contribution is:

Demonstration of robust SMO-based sensorless FOC performance in the medium-to-high speed region relevant to EPS actuator operation.

5. Integration of Functional Safety and Diagnostic Thinking into Sensorless Architecture

The thesis connects sensorless control to:

- ISO 26262 concept and system-phase requirements
- Estimator plausibility monitoring
- Cross-check logic between SMO and HFI channels
- Safe-state strategies under estimator inconsistency

Rather than presenting sensorless control as a purely performance-driven improvement, the work positions it within a structured safety and diagnostic framework suitable for production EPS environments.

Overall Contribution

The overall contribution of this thesis is the establishment of a coherent and scalable architectural argument:

Sensorless SMO + HFI hybrid estimation is not only algorithmically feasible but systemically integrable into modern sedan-class EPS architectures, provided that mechanical, electrical, software, and safety impacts are explicitly addressed.

The thesis therefore advances sensorless EPS from an algorithm-focused topic to a system-architecture-level engineering proposition.

8.2 Architectural Implications for Production EPS

The findings of this thesis indicate that the transition from sensor-based to sensorless EPS is not merely a control-algorithm substitution, but an architectural redistribution of responsibility across hardware, software, and safety domains.

In a conventional resolver-based EPS system, rotor electrical angle is a hardware-delivered boundary condition. The control algorithm consumes it as a deterministic signal with known error bounds and diagnostic interfaces. When that boundary is removed, the estimation function becomes part of the actuation chain itself. Consequently, angle estimation quality directly influences torque generation, steering smoothness, and safety-relevant behavior.

From a production EPS perspective, the architectural implications can be summarized as follows:

1. Hardware-Level Implications

Sensor removal reduces motor-side cost and harness complexity. However, the burden shifts toward:

- Higher signal integrity requirements for current sensing
- Compensation for inverter nonidealities (dead-time, voltage drops)
- Tighter timing synchronization between PWM, ADC sampling, and estimator updates
- Increased EMC robustness requirements

Thus, sensorless EPS is economically attractive only when the ECU hardware is engineered as an estimator-grade measurement platform.

2. Control Architecture Implications

The hybrid SMO + HFI architecture represents a partitioned observability strategy:

- HFI for low/zero speed observability
- SMO (back-EMF-based) for medium-to-high speed operation
- Transition logic ensuring angle continuity

The architectural challenge is not estimator accuracy alone, but:

- Stable cross-channel arbitration
- Delay compensation
- Robustness to parameter drift

- Minimization of estimator-induced torque ripple

The simulation results of Chapter 7 confirm that SMO-based estimation in the medium-to-high speed region satisfies actuator-level performance requirements under load disturbances. However, low-speed closed-loop HFI validation remains future work.

3. Software Architecture Implications (AUTOSAR Context)

Virtualizing rotor position transforms it into a software-derived safety-relevant signal.

Therefore:

- The estimator becomes a high-criticality software component.
- Timing determinism must be explicitly analyzed.
- Diagnostic exposure via Dem/Dcm must include estimator plausibility.
- Mode transitions must be state-managed rather than threshold-triggered.

The feasibility analysis shows that SMO + HFI can be integrated into an AUTOSAR-based ECU architecture provided that:

- Fast control loops remain tightly synchronized
- Diagnostic and monitoring logic is structurally separated
- Freedom-from-interference and timing budgets are respected

4. Functional Safety Implications

The safety argument shifts from “sensor fault detection” to:

- Estimator plausibility detection
- Cross-channel consistency monitoring
- Angle dynamic consistency checks
- Torque-direction validation

Sensorless EPS does not automatically simplify the safety case.
It alters it.

The estimator becomes a safety-relevant function whose failure modes must be addressed within the ISO 26262 lifecycle, and whose non-fault insufficiencies fall into the SOTIF domain.

8.3 Limitations of the Present Study

To maintain technical integrity, the limitations of this work must be explicitly stated.

1. Low-Speed Closed-Loop Validation Is Incomplete

Although an HFI signal-processing path was implemented and structurally integrated, the thesis does not present full closed-loop validation of HFI-assisted control in the zero/low-speed region.

Specifically:

- Stability of the combined SMO + HFI estimator was not quantitatively validated.
- Acoustic noise impact of HFI was not measured.
- Transition-region performance was not experimentally characterized.

Therefore, claims of “full-speed validated hybrid control” are not made in this thesis.

2. Vehicle-Level Steering Feel Is Not Simulated

The modeling scope is limited to motor-drive-level behavior.

Excluded:

- Rack and pinion dynamics
- Steering column compliance
- Tire-road interaction
- On-center steering feel
- Full NVH vehicle response

Thus, qualitative claims regarding steering feel remain architectural hypotheses supported by actuator-level robustness, not full vehicle simulation results.

3. Hardware-Level EMC Validation Is Analytical, Not Experimental

EMC and signal-integrity implications were discussed using literature and architectural reasoning.

However:

- No physical prototype was built.
- No CISPR 25 or ISO 11452 testing was conducted.
- No conducted transient testing per ISO 7637 was performed.

Hardware feasibility conclusions are therefore theoretical and literature-supported, not experimentally validated.

4. Safety Case Is Conceptual, Not Certified

The safety discussion aligns with ISO 26262 lifecycle principles.

However:

- No formal HARA document is produced.
- No quantitative FMEDA is calculated.
- No ASIL decomposition claim is made.

The functional safety discussion establishes architectural compatibility, not certification-level evidence.

8.4 Future Research Roadmap

The work opens several directions for continuation.

1. Full-Speed Closed-Loop Hybrid Validation

Future work should:

- Complete HFI closed-loop validation at standstill and parking conditions.
- Quantify angle error and torque ripple during mode transitions.
- Evaluate injection frequency and amplitude scheduling strategies.

2. Integration with Assist-Torque Control Architecture

The present simulation uses speed control as a structured validation framework.

Future work should:

- Replace outer speed loop with assist torque control logic.
- Evaluate steering-reversal behavior.

- Introduce return-to-center test scenarios.

3. Vehicle-Level Co-Simulation

Integration with:

- Rack-and-pinion dynamic model
- Steering compliance model
- Vehicle lateral dynamics

would allow direct evaluation of steering feel and NVH implications.

4. Functional Safety Quantification

Future research could:

- Develop formal HARA for sensorless EPS.
- Quantify diagnostic coverage of SMO–HFI cross-check logic.
- Explore ASIL decomposition feasibility with dual estimation channels.

5. Toward Steer-by-Wire Architectures

Sensorless estimation is structurally aligned with steer-by-wire systems where mechanical fallback may not exist.

Future work should explore:

- Redundant sensorless architectures
- Fail-operational estimation logic
- Integration with autonomous steering controllers

Final Concluding Statement

In conclusion, this thesis demonstrates that a hybrid SMO + HFI sensorless control architecture can serve as a technically coherent and production-feasible foundation for sedan-class EPS systems when evaluated from a systems engineering perspective.

The validated medium-to-high speed SMO performance, combined with the structured architectural analysis across mechanical, electrical, software, and safety domains, supports the argument that sensorless EPS is not

merely a cost-driven alternative but a scalable engineering pathway—provided its limitations and safety implications are explicitly addressed.

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Appendices

Appendix A

ASPICE - Automotive SPICE (Automotive Software Process Improvement and Capability determination)

A.1 Overview of Automotive SPICE and Process Capability Model

Automotive SPICE (hereinafter referred to as ASPICE) is a software process assessment model specialized for the automotive field, designed to quantitatively evaluate and improve the development process capabilities of organizations.

ASPICE is an automotive-specific extension based on the international standards ISO/IEC 15504 (currently ISO/IEC 330xx series) for the Process Reference Model (PRM) and the Process Assessment Model (PAM). It defines each process in the software development life cycle and evaluates its execution status by assigning capability levels ranging from 0 to 5.

Capability within ASPICE is expressed using a two-dimensional model:

- Horizontally, it organizes process areas such as requirements definition, design, and testing.
- Vertically, it defines capability maturity indicators from Level 0 (Incomplete) to Level 5 (Optimizing).

For instance:

- Level 1 signifies that the process is basically performed,
- Level 3 signifies that a standardized and managed process has been established within the organization,
- Level 5 indicates that continuous improvement is achieved based on quantitative feedback.

Typically, ASPICE Level 2 (managed implementation of standardized processes) is considered the minimum desirable level, and Level 3 (established quantitative management process) is regarded as an industry standard for excellence.

Conversely, achieving Levels 4 and 5 remains an ambitious goal pursued primarily by large enterprises; reaching these levels is not mandatory for all organizations.

Notably, ASPICE does not mandate specific development methods or tools; instead, it serves as a guideline to allow each company to devise best practices suited to its context.

In terms of global trends, the German Automotive Industry Association (VDA) has effectively made ASPICE the standard for supplier evaluations, and other international OEMs have followed suit. In fact, it has been pointed out that “for modern automotive development companies, development compliant with ASPICE is already indispensable”. Consequently, many OEMs and Tier 1 suppliers have been working actively to establish their processes and enhance capabilities based on the ASPICE model.

A.2 Application of ASPICE to EPS Development and Case Studies

Since steering control is critical to vehicle safety, rigorous process management is required in the development of EPS.

At J-TEKT Corporation, a major Japanese supplier of EPS systems, it has been reported that "the definitions and improvements of development processes have been carried out through ASPICE activities even before the introduction of functional safety measures in EPS development".

This demonstrates that the company implemented ASPICE-compliant processes early on for EPS development and integrated them with ISO 26262 functional safety processes.

Specifically, JTEKT has built a development process based on the V-model, covering all stages from Hazard Analysis and Risk Assessment (HARA) to system requirements definition, software requirements definition, design, and testing.

Through this process, the company has successfully realized a development flow compliant with the stringent ASIL-D safety requirements, thereby enhancing the reliability of EPS control software.

Moreover, even outside Japan, such practices are becoming widespread. For example, in South Korea, research has been conducted proposing an optimization method for EPS software task scheduling using AUTOSAR.

This approach defines and achieves quality goals such as real-time performance and torque ripple reduction within the development process.

Thus, EPS development process demonstrates the trend of adopting ASPICE-based extended processes that integrate safety and performance objectives.

A.3 Process Assessment Methods and Capability Maturity Evaluation

Process assessment based on ASPICE is conducted through objective evaluations by certified assessors (holders of initial qualification).

Specifically, assessment involves examining artifacts generated during the development process (e.g., requirements specifications, design documents, test reports) and interviewing involved members to evaluate whether each process meets its defined base practices and achieves the corresponding Process Attributes (PAs) at the required capability level.

ASPICE PAM 3.1 defines 16 major processes, including:

- System Requirements Analysis (SYS.2),
- Software Requirements Analysis (SWE.1),
- Software Design (SWE.2/3),
- Software Integration Testing (SWE.5), etc.

Each process is scored on evaluation scales such as “Fully Achieved” and “Largely Achieved”.

For example, to achieve Level 2, all relevant processes must have Process Attributes 1.1 (Work Product Quality) through 2.2 (Monitoring and Adjustment) evaluated as “Largely Achieved” or higher.

The results of the evaluation identify organizational and project strengths and weaknesses and are fed back into process improvement plans.

From the viewpoint of capability maturity model analysis, ASPICE Level 3 roughly corresponds to Level 3 of the conventional CMMI model (Defined Process).

However, since ASPICE includes processes specific to the automotive domain (such as functional safety and supplier management), each organization must set evaluation indicators tailored to its development context.

In recent years, extensions incorporating the requirements of ISO 26262 (Functional Safety) and ISO/SAE 21434 (Cybersecurity) into ASPICE have been proposed.

For instance, Doba et al. (2021) proposed a development lifecycle model that extends ASPICE with cybersecurity processes and verified its effectiveness in EPS system development.

Thus, EPS development organizations are evolving process assessments from mere quality checklists into comprehensive maturity evaluations integrating both safety and cybersecurity.

A.4 Company-Specific ASPICE Adoption Cases and Capability Comparisons

The degree of ASPICE adoption has historically varied depending on the region and company size. However, in recent years, its global proliferation has accelerated.

European OEMs, particularly German manufacturers, were early adopters of ASPICE and began requiring suppliers to implement processes equivalent to ASPICE Level 3. Simultaneously, they actively promoted organizational process improvement internally.

In Germany, ASPICE PAM (Process Assessment Model) has effectively become a mandatory prerequisite for supplier selection, and companies such as the Volkswagen Group routinely demand that suppliers achieve at least ASPICE Level 2.

Meanwhile, within Japan, there have been reports since around 2017 indicating that "some manufacturers have already begun ECU software development based on AUTOSAR, with full-scale consideration of ASPICE introduction underway".

In fact, JTEKT Corporation, a Toyota Group supplier, has introduced ASPICE-compliant processes in parallel with ISO 26262 compliance efforts.

Through responses to European process and expanding global business, Japanese suppliers have increasingly consolidated ASPICE-adherent development practices.

In addition, the automotive industries in countries like India and China have also accelerated ASPICE adoption recently, aiming to meet global OEMs' required quality standards.

However, the approaches to ASPICE implementation vary greatly depending on corporate size and cultural background. Some companies have explored harmonization with agile development methodologies.

For instance, one case study reported a hybrid model where ASPICE-based process compliance was combined with Scrum iterative development and Continuous Integration (CI) practices, achieving a successful balance between process maturity and development speed.

Furthermore, there has been increasing incorporation of recent ASPICE extensions, such as Automotive SPICE for Cybersecurity and ASPICE for Mechanical Engineering, by advanced companies seeking to further sophisticate their process maturity models.

In summary, for systems critical to safe driving like EPS, achieving approximately ASPICE Level 3 has become a common goal across various organizations, demonstrating the industry's commitment to securing quality, safety, and cybersecurity. [9-1],[9-2],[9-3]

Appendix B

AUTOSAR

B-1 Implementation and Utilization Cases of Standardized OS and Development Environment Using AUTOSAR

B.1.1 Overview of AUTOSAR Classic and Adaptive Platforms

AUTOSAR (Automotive Open System Architecture) is an industry-standard architecture that provides standardized frameworks for in-vehicle ECU software. Since its establishment in 2003, AUTOSAR has released four major versions centered around the Classic Platform (CP).

The AUTOSAR Classic Platform operates on a static real-time operating system based on OSEK/VDX, designed for embedded control applications that execute deterministic periodic tasks directly on the vehicle's microcontroller.

In contrast, the AUTOSAR Adaptive Platform (AP), released in 2017, operates on dynamic POSIX-compliant operating systems (e.g., Linux, QNX) and is tailored for use cases requiring high-performance computing, such as advanced driver assistance systems (ADAS) and autonomous driving.

Importantly, the Adaptive Platform is not an upward-compatible evolution of the Classic Platform; rather, the two are designed to coexist cooperatively.

In practice:

The Classic Platform primarily targets sensor/actuator ECUs (e.g., EPS, braking, engine control) and emphasizes hard real-time behavior and high reliability.

Typically, applications are implemented in C language as monolithic binaries, directly executed from ROM, with minimal memory protection (i.e., single address space with basic MPU/MMU usage).

In contrast:

The Adaptive Platform adopts a service-oriented architecture and allows multiple applications, often written in C++, to be dynamically loaded and executed.

Each application runs in a separate virtual address space, with full memory protection enforced by the OS (MMU usage).

Although the real-time requirements for Adaptive are less stringent (targeting millisecond-level responsiveness), certain high-priority threads can ensure real-time performance for critical autonomous driving functions.

Regarding communication:

The Classic Platform is optimized for connectionless, signal-oriented communication (event-triggered, periodic broadcasts) over CAN/LIN networks,

Whereas the Adaptive Platform assumes service-oriented communication (e.g., SOME/IP, DDS) and high-speed Ethernet as the primary medium.

Thus, sensor/actuator ECUs like EPS are still generally implemented using the Classic Platform today.

However, in future architectures, centralized server units running on the Adaptive Platform are expected to perform vehicle dynamics computations and issue steering commands to EPS ECUs, suggesting a hybrid architecture combining Classic and Adaptive elements.

In summary, while the AUTOSAR Classic and Adaptive Platforms differ in purpose and characteristics, both play vital roles in enhancing software reusability and interoperability across ECUs and manufacturers through standardization.

B.1.2 Application of AUTOSAR to EPS Control and ECU Architecture

Typically, EPS ECUs are equipped with 32-bit automotive microcontrollers, such as those from the Renesas RH850 or Infineon AURIX families, incorporating peripherals for motor control, including timers, ADCs, and CAN interfaces.

By applying AUTOSAR Classic to EPS systems, the hardware-dependent portions of the microcontroller can be abstracted through the Microcontroller Abstraction Layer (MCAL), enabling higher-level application software to be described independently of hardware specifics.

The AUTOSAR standard OS (based on OSEK) provides priority-based scheduling for multitasking.

Typically, EPS control software is divided into multiple tasks, including:

High-frequency control tasks: Executed every 1 ms, performing torque control calculations based on steering torque sensor values (via A/D conversion) and outputting motor target current or PWM values to hardware registers.

Communication tasks: Executed approximately every 10 ms, managing the reception and transmission of CAN messages, dynamically adjusting assist characteristics based on vehicle speed or ESC operation received from other ECUs.

Diagnostic monitor tasks: Executed approximately every 100 ms, handling internal self-diagnostics, range checks of sensor values, temperature and voltage monitoring, and triggering fail-safe measures when abnormalities are detected.

Each task is scheduled by the AUTOSAR OS with static priorities to guarantee real-time performance.

Memory allocation is abstracted through AUTOSAR's memory services and drivers, efficiently managing code, data, and calibration parameters stored in Flash ROM.

Internally, EPS ECUs often incorporate dual cores or lock-step redundant cores. For instance, one core may handle main control tasks while another independently runs monitoring tasks.

AUTOSAR OS provides multi-core support extensions to enable such architecture.

Additionally, with the increasing sophistication of steering angle sensors and torque sensors, EPS systems require the fusion of multiple sensor signals.

Under AUTOSAR architecture, such signal processing is modularized as Software Components (SW-Cs), with communication between SW-Cs and with external ECUs managed via the Runtime Environment (RTE).

By designing EPS control software using AUTOSAR, portability is enhanced, allowing SW-Cs to be reused across different microcontroller platforms or vehicle models.

In Japan, for instance, Mitsubishi Electric Software Corporation reported experiments involving the migration of EPS ECU software developed on AUTOSAR R3.2 to AUTOSAR R4.2, verifying multi-core support and the addition of security functions.

B.1.3 Utilization of Vector/MICROSAR Development Environment

For AUTOSAR-compliant development, the use of advanced tool chains is indispensable.

Particularly, Vector's MICROSAR is a globally recognized implementation of AUTOSAR-compliant Basic Software (BSW) and is widely utilized in EPS development projects.

MICROSAR provides a comprehensive suite of software modules, including:

- OS (Operating System),
- Communication stacks (COM stack for CAN/LIN communication),
- Memory services,
- Diagnostic services (UDS communication layer), all compliant with AUTOSAR standards.

- Developers use Vector's DaVinci Configurator Pro tool to configure various MICROSAR BSW modules — for example, specifying:
- Which CAN IDs to receive,
- Task priorities within the OS,
- Memory driver mappings —and to integrate the generated source code into their protect builds.

Additionally, the DaVinci Developer tool supports the design of interfaces between Software Components (SW-Cs), assisting in the definition of ports and runnable (execution units).

In a case study by Mitsubishi Electric Software Corporation, the migration of an EPS ECU from AUTOSAR 3.2 to AUTOSAR 4.2 was performed through a sequence of phases:

Requirements analysis → Configuration → Code generation → Build → Debugging.

During the requirements analysis phase, key decisions such as microcontroller types, AUTOSAR versions, and BSW module selections were finalized.

Subsequently, during configuration, the previous (3.2) settings were carefully adapted for 4.2 compatibility, automatic code generation was performed, and new functions (such as activation of SecOC) were tested.

Throughout this migration process, Vector's toolchain facilitated automatic conversion and error checking, enabling rapid adaptation and minimizing manual effort.

Furthermore, recent Vector products include:

- The VT System (Hardware-in-the-Loop simulation),
- CANoe (comprehensive ECU and network simulation/testing platform), allowing developers to connect EPS ECUs to simulation environments that mimic steering loads, CAN traffic, and perform automated testing of control and diagnostic functions.

Thus, the utilization of Vector's MICROSAR development environment significantly enhances the efficiency and reliability of EPS ECU development and is widely adopted in both domestic and international process.

Thus, AUTOSAR Classic-based architecture enables EPS ECUs to seamlessly combine real-time motor control with vehicle network communication, supporting modularity, reusability, and safety.

B.1.4 Development and Validation Using Virtual ECUs (V-ECU)

Recently, Virtual ECU (V-ECU) technology, which allows ECU software to be executed on a PC without requiring actual hardware, has attracted significant attention.

Particularly for EPS systems, which involve vehicle dynamics control, V-ECUs are increasingly utilized for early software verification and integration into Continuous Integration/Continuous Deployment (CI/CD) pipelines.

A notable example is the case of Tata Elxsi, where a Level 3 Virtual ECU environment was constructed for an EPS system supplier using dSPACE tools.

Approximately 80% to 90% of the EPS control software (excluding hardware-dependent portions) was executable on a standard PC environment.

This setup enabled sophisticated simulations, including:

- CAN communication,
- Diagnostic services, without relying on physical hardware availability, thus allowing early detection of software bugs and integration into CI/CD processes.

In the reported case, the adoption of the Virtual ECU led to a dramatic increase in testing efficiency and achieved approximately 20% cost reduction in development.

Furthermore, a point project with a U.S. municipal agency demonstrated experiments where cloud-based EPS V-ECUs were deployed for remote software updates and validations, thus pioneering cloud-driven automotive software verification (related discussions are elaborated in Section 3.4 on OTA/FOTA).

In implementing Virtual ECUs:

- Simulation of the microcontroller's instruction sets,
- Accurate timing emulation,
- Peripheral I/O (ADC, PWM, CAN) emulation via tools like CANoe are crucial.

Given that AUTOSAR-compliant software abstracts much of the hardware dependency, it is relatively easier to run AUTOSAR-based EPS control software on host PCs.

However, low-level modules such as Complex Device Drivers and hardware interrupt handling still require special modeling or stubbing.

Nevertheless, the ability to perform integration testing before hardware availability presents significant advantages in EPS development.

Moreover, when bugs are detected, the V-ECU environment allows direct memory dump analysis and debugger attachment, facilitating detailed troubleshooting that would be difficult with physical hardware.

Thus, Virtual ECU technology is increasingly regarded as an innovative method for significantly streamlining parts of the EPS software development process.

B.1.5 Practical Examples of BSW Configuration and Task/Memory Design

This section provides practical examples regarding the configuration of Basic Software (BSW) and the design of tasks and memory layout in EPS ECUs based on the AUTOSAR Classic Platform.

Consider a mass-production EPS ECU with the following typical task structure:

Control Task: Executed at a fast cycle of 1ms.

It reads steering torque sensor values via A/D conversion, performs torque control algorithms (such as PID control or torque mapping), and outputs the motor target current values as PWM duty cycles to hardware registers.

This task is assigned the highest priority (e.g., OS priority 1) to ensure immediate execution without delay.

Communication Task: Executed every 10ms.

It handles the reception and transmission of CAN messages via the AUTOSAR COM stack and CAN Interface/Driver modules.

For instance, it dynamically adjusts assist characteristics based on vehicle speed or ESC operation information received from other ECUs and sends steering angle and fault information to other systems. This task is assigned a mid-level priority (e.g., OS priority 2).

Diagnostic Task: Executed every 100ms.

It conducts self-diagnostics, sensor value range checking, temperature and voltage monitoring, and sets Diagnostic Trouble Codes (DTCs) when abnormalities are detected.

It also handles UDS (Unified Diagnostic Services) requests as described later.

While assigned a lower priority (e.g., OS priority 3), it is capable of event-driven execution to promptly respond to critical faults.

Monitoring Task: Executed on a separate core or at a high-priority interrupt level.

It performs independent recalculations to verify the results of the main control task (Dead Man Monitoring), monitors CPU traps, and manages watchdog communications with Safety ICs.

To ensure high reliability, this task is designed to operate within an independent execution context as much as possible.

Based on such task designs, the various parameters of the AUTOSAR BSW — including OS scheduling tables, CAN reception task assignments, Watchdog driver configurations, and ADC driver settings — are defined accordingly.

Regarding memory configuration:

- EPS ECUs compliant with ASIL-D typically employ Error Correction Code (ECC) protected memories.
- Flash ROM stores control codes, calibration parameters, and diagnostic-related data.
- RAM holds runtime variables and buffers.
- To ensure robustness, AUTOSAR's Memory Protection Mechanism (Memory Partitioning) is used to prevent catastrophic impacts from memory corruption. For instance, critical control loop variables are stored redundantly in separate memory areas and cross-checked for consistency.

Such safety measures are implemented in accordance with ISO 26262 guidelines. Moreover, Non-Volatile RAM (NVRAM) management (NVM) is crucial for EPS ECUs. For example, learned data such as steering neutral position offsets must be preserved even after power-off.

By utilizing AUTOSAR's NVM services, the ECU can automatically flush important data from RAM to Flash memory before shutdown.

Thus, the configuration of AUTOSAR BSW modules is performed carefully to meet EPS-specific requirements such as:

- High-frequency real-time control,
- Safety-critical monitoring,
- Persistent data retention.
- Task and memory design are optimized in tandem to support these requirements.

Although the above presents only an example, actual EPS products from different manufacturers demonstrate similar design considerations according to various technical reports. [9-4],[9-5],[9-6]

Appendix C

CAN

C.1 The Role and Utilization of CAN Network Systems in EPS

C.1.1 Overview of EPS and In-Vehicle Networks

EPS systems must collaborate with other in-vehicle systems, and the primary means of communication is through the Controller Area Network (CAN).

CAN is a high-speed communication bus developed specifically for automotive applications under the ISO 11898 standard.

EPS ECUs use CAN to exchange data with other ECUs, such as:

- Engine control ECUs,
- Brake control ECUs,
- Vehicle speed sensors,
- ADAS units.

For example:

- Vehicle speed information is critical for controlling the steering assist torque.
- Speed sensor data is transmitted as CAN messages to the EPS ECU, which adjusts the assist amount (e.g., reducing assist at high speeds) accordingly.

Additionally, systems such as Electronic Stability Control (ESC) or automated parking systems may send steering torque or steering angle targets to the EPS ECU.

In modern autonomous parking applications:

- The EPS may receive direct commands specifying target steering angles from the ADAS ECU,
- The EPS motor drives accordingly, emphasizing the importance of trustworthy and secure communication links (cybersecurity aspects will be discussed in Chapter 4).

Moreover, the EPS ECU periodically transmits its own internal status, such as:

- Steering angle,
- Assist torque,
- Fault information,
over the CAN bus to the vehicle's Gateway ECU, which may further distribute it to driver warning systems or diagnostic devices.

Thus, the EPS system plays a central role within the vehicle's CAN network, both as a receiver of external sensor and command data and as a transmitter of its internal state.

C.1.1.2 Support for CAN FD and Improvement of Communication Performance

In recent years, the extension of CAN, known as CAN-FD (CAN with Flexible Data-rate), has been realized and is increasingly being applied to in-vehicle networks, including EPS systems.

Originally proposed by Bosch around 2012, CAN-FD improves upon classical CAN by:

- Expanding the maximum payload size from 8 bytes to 64 bytes per frame, and
- Increasing communication speed from 500 kbps up to 2 Mbps or even 5 Mbps.

For ECUs like EPS that frequently transmit and receive analog quantity data (e.g., torque sensor readings, motor control commands), CAN-FD offers significant benefits:

- A larger amount of data can be sent at once,
- Reducing the number of frames required,
- Shortening communication latency,
- Enhancing the responsiveness of real-time control loops.

For example, Infineon's next-generation EPS chipset integrates CAN-FD compliant transceivers (TLE9251/TLE9252) to achieve reliable, high-speed data transfer.

In experimental demonstration boards:

- Dual-redundant EPS architectures were constructed using six-phase motors,
- Multiple CAN-FD channels were used to interconnect subsystems,

- Enabling fail-operational behavior, where steering assist is maintained even in case of a subsystem failure.

Specifically:

- Two independent control circuits and motor windings are deployed,
- Each connected via separate CAN-FD networks,
- If one network or subsystem fails, the other seamlessly continues providing steering assistance.

This type of dual-CAN redundant architecture is being considered for highly reliable steering systems, particularly in Level 3 and above autonomous driving vehicles.

Typically:

- The same steering commands are sent over both CAN networks,
- The EPS ECU monitors both channels,
- In case of a discrepancy or failure in one channel, it switches to the backup channel without interruption.

In current mass-production vehicles, dual-CAN-port EPS ECUs have already been adopted for high-safety applications, such as heavy-duty trucks.

In these cases, main and backup CAN channels are both prepared to meet the stringent safety requirements of ASIL-D under ISO 26262.

Thus, by introducing CAN-FD and redundant CAN architectures, the communication systems of EPS are achieving ever higher levels of reliability and performance, laying the foundation for future steer-by-wire era vehicles.

C.1.3 Application of UDS Diagnostic Communication Protocol to EPS

EPS systems are also connected to the vehicle's diagnostic system, and communication for diagnostics typically follows the Unified Diagnostic Services (UDS) protocol, as specified in ISO 14229.

UDS defines a standardized framework for performing:

- Fault diagnosis,
- Memory readouts,

- Actuator control tests,
- ECU reprogramming,
through request/response message exchanges over the CAN bus.

EPS ECUs, similar to engine or transmission ECUs, fully support UDS services, including:

- Storage and retrieval of Diagnostic Trouble Codes (DTCs),
- On-demand reading of sensor values,
- Actuator control operations,
- Software updates.

Specific examples of UDS services applicable to EPS include:

- Service 0x14 (Clear Diagnostic Information): Clears stored DTCs,
- Service 0x22 (Read Data By Identifier): Reads current values, such as steering torque,
- Service 0x31 (Routine Control): Executes routines, such as warning lamp activation for end-user notifications.

UDS messages are typically transported over CAN using the ISO-TP (ISO 15765-2) protocol.

When a UDS message is received:

- The EPS ECU's diagnostic service module within AUTOSAR Basic Software parses the request,
- Executes the corresponding action (e.g., referencing DTC memory),
- And returns a properly formatted UDS response.

Diagnostic testers (scantools) use these interactions to assess the health of the EPS system, identify fault locations, and guide appropriate repair actions.

Common EPS-related DTCs include:

- U0001 (CAN communication bus fault),
- C1B00 (Steering torque sensor malfunction), etc.

EPS ECUs store these DTCs in internal non-volatile memory and provide UDS-based readouts upon request.

Moreover, UDS supports software reprogramming functions, using services such as:

- 0x34 (Request Download) and
- 0x36 (Transfer Data),
allowing new firmware to be written into the ECU over the CAN bus.

This capability enables in-service software updates for EPS systems during recalls or maintenance campaigns without requiring physical ECU replacement.

Thus, UDS diagnostic communication is an indispensable function for the maintenance, repair, and software update processes of EPS, and is consistently implemented within the AUTOSAR Diagnostic Stack.

C.1.4 Case Studies of OTA/FOTA Software Updates for EPS

Traditionally, EPS ECU software updates were performed via wired connections at factories or service centers.

However, with the advancement of Over-The-Air (OTA) update technologies in recent years, initiatives to update EPS firmware wirelessly have begun.

A notable example is Tesla, a pioneer in OTA technology.

In 2023, Tesla remotely distributed a software fix via OTA to address a defect in certain vehicles where EPS assist was lost at low speeds.

In this case:

- A bug was discovered where EPS assist could suddenly deactivate when the vehicle speed approached zero,
- Tesla released a correction in software update version 2023.38.4.

This OTA deployment enabled users to improve their EPS functionality without the need to visit service centers, demonstrating the effectiveness of OTA in remotely correcting safety-critical steering issues.

Other manufacturers, such as General Motors (GM) and Ford, are also adopting OTA frameworks by:

- Equipping new vehicle models with communication modules and
- Secure gateways, enabling remote reprogramming of multiple ECUs, including the EPS ECU

Nevertheless, updating safety-critical ECUs like EPS via OTA requires careful design to mitigate risks:

- If power loss or communication interruptions occur during an update, EPS functionality could be lost, posing serious hazards.
- Therefore, fallback mechanisms and dual-bank flash memory architectures are necessary to allow rollback to a safe software version in case of update failure.

Precautionary measures also include:

- Ensuring the vehicle is parked in a safe condition before initiating updates,
- Maintaining EPS ECU power-on status during the update process.

From a security perspective:

- OTA update packages must be cryptographically signed,
- The EPS ECU must verify the signature to authenticate the update before accepting it.

In Tesla's case, it is inferred that OTA updates incorporated cryptographic validation before overwriting EPS firmware.

Moreover, OTA technology enables manufacturers to continuously improve EPS control logic based on field-collected data:

- For instance, tuning the steering feel under specific driving conditions,
- And delivering such improvements to vehicles already in use via OTA updates.

Toyota has also reportedly experimented with OTA deployment of EPS parameter changes for commercially available vehicles.

Thus, OTA/FOTA technology is advancing into the domain of EPS systems, contributing to enhanced user experiences and swift resolution of functional defects, albeit with the necessity to balance high safety and cybersecurity requirements meticulously.

C.1.5 Redundant CAN Designs and Fail-Safe Communication for EPS

As discussed earlier, ensuring high reliability is critical for EPS systems, particularly in the context of Steer-by-Wire (SbW) systems and autonomous vehicles.

Accordingly, communication redundancy is being actively considered and implemented.

The basic concept of redundant CAN design is to:

- Use two independent CAN buses,
- Send either identical or complementary messages on both buses, and
- Monitor both channels at the EPS ECU side.

If an error or delay occurs on one bus:

- The system automatically switches to the other channel to maintain continuous steering control.

This architecture enhances resilience against single-point failures such as:

- Wiring disconnections,
- Short circuits,
- Bus overloads, etc.

Moreover, on the vehicle side, dual-CAN capable gateways and sensor networks provide redundant inputs to the EPS ECU.

In some high-end vehicles:

- Even the steering angle sensor values are transmitted redundantly through multiple network paths.

At the protocol level, although time-triggered networks such as TTP/C and FlexRay once offered redundancy, due to cost and industry trends, CAN redundancy has become the de facto mainstream approach.

Under ISO 26262:

- For systems requiring ASIL-D classification (the highest automotive safety level),
- Fallback through redundant communication channels is highly recommended.

A real-world example is found in some GM vehicles, where:

- A dedicated secondary CAN network is used solely for safety-critical messages,
- Allowing mutual cross-communication between EPS ECUs and brake control ECUs.

In case the main CAN network fails:

- The secondary CAN network ensures that minimum essential control commands (e.g., "maintain current steering angle") are still transmitted,
- Avoiding sudden loss of steering control.

Internally, the EPS ECU compares command inputs from both CAN channels:

- If discrepancies persist beyond a defined time threshold,
- It flags a communication fault, alerts the driver, and switches to a predefined fail-safe control mode.

Such multi-layered redundant designs significantly enhance the communication reliability and safety of EPS systems.

Nevertheless, redundancy comes with trade-offs:

- Increased system complexity,
- Higher costs,
- Added weight.

Thus, risk assessments must be conducted on a per-vehicle basis to determine the appropriate level of redundancy.

In the future, as Automotive Ethernet (with Time-Sensitive Networking, TSN) becomes more prevalent:

- Redundancy concepts will evolve towards VLAN partitioning and dual-homing configurations.

In summary, EPS communication systems are progressively adopting multilayered redundancy architectures to ensure fail-operational and fail-safe steering control, even under fault or cyberattack scenarios. [9-7], [9-8], [9-9]

Appendix D

Cybersecurity

D.1 Cybersecurity Challenges and Countermeasures for EPS

D.1.1 Threat Modeling and Risk Assessment

In 2021, the international automotive cybersecurity standard ISO/SAE 21434 was issued, mandating that safety-critical systems such as EPS undergo cybersecurity risk assessments from the early stages of development.

A representative cybersecurity threat to EPS is the unauthorized manipulation of steering behavior by injecting malicious control commands externally.

Indeed, in a famous case, Miller and Valasek (2015) demonstrated that by exploiting vulnerabilities in the telematics unit,

they were able to infiltrate the CAN bus of a Jeep Cherokee via the Internet and remotely control not only infotainment systems but also steering and braking functions.

Initially, their hack was limited to manipulating steering only while the vehicle was in reverse gear, but further research enabled limited control even during highway driving.

This attack triggered a massive recall of approximately 1.4 million vehicles (software updates), sending shockwaves throughout the automotive industry.

These incidents highlight that cyberattacks targeting EPS are not mere theoretical risks but real and consequential threats.

Threat modeling systematically identifies:

- Possible attack paths (e.g., via external CAN ports, telematics units, or adjacent ECUs),
- Assesses the impact and likelihood of each threat scenario.

In the case of EPS:

- The impact is categorized as "Severe" because it could lead to the complete loss of steering control,
- Thus presenting a critical safety risk.

Regarding likelihood:

- Although modern vehicles implement multiple layers of network defense,
- The increasing connectivity (e.g., OTA updates, V2X communication) raises the exposure surface,
- Requiring careful analysis of threat actors, attack difficulty, and existing vulnerabilities.

After comprehensive threat modeling, a Threat Analysis and Risk Assessment (TARA) is conducted:

- High-risk items that are not acceptable are mapped to security goals and requirements,
- Which are then implemented within the system architecture.

For instance, security requirements may include:

- "Detection and nullification of forged EPS control messages,"
- "Protection of internal ECU flash memory against unauthorized code execution".

Doba et al. (2021) proposed a security-extended ASPICE development lifecycle model and verified its applicability for EPS development, emphasizing the importance of conducting threat modeling, vulnerability analysis, and security requirement definition at both system and subsystem levels.

Thus, security risk assessments for EPS are systematically conducted in line with ISO/SAE 21434, in parallel with conventional functional safety (ISO 26262) processes.

D.1.2 Vulnerability Analysis and Practical Attack Examples (on EPS ECUs)

Following threat analysis, vulnerabilities in EPS systems can be broadly categorized into two areas:

- Communication channels, and
- Internal ECU software.

For communication channels:

Since the CAN protocol itself lacks intrinsic authentication and encryption mechanisms, if an attacker gains access to the in-vehicle CAN network, they can forge or tamper with EPS control frames without detection.

In the Jeep Cherokee hacking incident:

- The attackers first exploited vulnerabilities in the telematics unit to gain CAN bus access,
- Then transmitted forged steering torque commands to the EPS ECU, enabling remote steering manipulation.

Such unauthorized message injection attacks can occur either:

- Directly from external access points (e.g., telematics units, OBD-II ports), or
- Indirectly via compromised adjacent ECUs.

To address communication-related vulnerabilities, automakers are increasingly introducing:

- ECU authentication mechanisms and
- Message Authentication Codes (MAC) (refer to the following Section 4.3 on SecOC).

For internal ECU software vulnerabilities:

Common issues include:

- Buffer overflows,
- Exploitable debugging interfaces,
- Bootloader vulnerabilities.

In a notable example, a white-hat hacker group successfully reverse-engineered the EPS ECU firmware of a Toyota vehicle, exploited improper bootloader validation, and achieved arbitrary code execution.

Their method involved:

- Physically accessing the ECU,
- Extracting the internal flash memory contents,
- Reverse engineering the firmware to recover cryptographic keys used for SecOC authentication.

Subsequently, they modified the bootloader mode to inject unauthorized firmware, allowing aftermarket devices (such as aftermarket autopilot units) to gain control over the EPS ECU.

This case illustrates that even if communication security (e.g., SecOC) is properly implemented, the entire EPS system can be compromised if internal ECU software security measures (e.g., key management, boot authentication) are inadequate.

Although such sophisticated physical attacks require technical expertise and physical access, they represent a realistic threat, particularly in cases of vehicle theft or targeted attacks.

Therefore, manufacturers must perform rigorous security validation during development, including:

- Penetration testing and
- Fuzz testing,
to detect and eliminate latent vulnerabilities.

A noteworthy trend here is the introduction of attacker-centric FMEA (Threat-oriented FMEA) methodologies, which extend conventional Failure Mode and Effects Analysis to include security attack perspectives for evaluating systems like EPS.

Thus, cybersecurity risk assessment and countermeasures are becoming an integral part of EPS system development.

D.1.3 Intrusion Detection Systems (IDS) and Application of Secure Communication Protocols

To defend EPS systems against cyberattacks, manufacturers are increasingly adopting:

- Intrusion Detection Systems (IDS) and
- Secure onboard communication protocols.

Regarding IDS:

- Some systems deploy centralized IDS across the in-vehicle network backbone,
- While others embed lightweight IDS modules inside each ECU individually.

For EPS, embedding a lightweight IDS directly within the ECU is a practical approach.

The IDS can continuously monitor steering command or torque request patterns in real-time.

Specifically:

- Normally, steering torque requests from systems like ESC or ADAS remain within predictable frequency and range limits.
- If malicious messages cause unusually frequent or extreme value fluctuations, the IDS can detect the anomaly and trigger safety fallback actions (e.g., ignore the suspicious input, prioritize manual driver input).

Recent research has proposed various IDS techniques, such as:

- Timing anomaly detection based on statistical learning of CAN message inter-arrival patterns,
- Correlation-based detection comparing steering angle measurements with yaw rate sensor outputs to infer illegitimate steering manipulations.

In addition to IDS, the Secure Onboard Communication (SecOC) mechanism defined by AUTOSAR standards plays a pivotal role.

SecOC enhances communication security by:

- Attaching a Message Authentication Code (MAC) to each message,
- Verifying the MAC at the receiving ECU based on pre-shared secret keys,
- Ensuring data authenticity and freshness (preventing replay attacks).

The cryptographic keys (typically using AES-CMAC algorithms) are securely shared among ECUs during vehicle manufacturing.

Thus, unless attackers can compromise the keys:

- Forged steering control messages would be detected and rejected by the EPS ECU's SecOC module.

Many OEMs, including Toyota, have already implemented SecOC for major ECU communications, including EPS control systems, in vehicles launched in the 2020s.

Nevertheless, SecOC is not a panacea:

- As demonstrated by the aforementioned hacking case,
- If an attacker extracts the cryptographic keys from an ECU's memory (e.g., by physical teardown and flash memory reading),
- They could forge valid SecOC-authenticated messages.

To prevent such attacks, automotive microcontrollers increasingly incorporate a Hardware Security Module (HSM):

- Keys are securely stored within hardware-protected areas,
- MAC calculations are executed by the HSM, not by general-purpose CPU resources [CAN-CIA.org].

Additionally, at the network perimeter level, vehicle gateway ECUs implement:

- Firewall functions to block unauthorized CAN IDs,
- Traffic shaping mechanisms to mitigate Denial-of-Service (DoS) attacks caused by message flooding.

Therefore, defense in depth is practiced at multiple levels:

- Gateway IDS at external network boundaries,

- Centralized IDS for backbone Ethernet,
- Lightweight IDS within local CAN segments,
- Secure communication protocols like SecOC embedded in ECUs.

For EPS in particular:

- The emphasis is placed on combining SecOC-based communication authentication with
- Failsafe fallback strategies triggered by anomaly detections.

For example:

- Even if an IDS fails to perfectly detect an attack,
- The EPS ECU itself may apply torque clamping (e.g., limiting excessive steering commands) and prioritize manual driver input as a defensive measure.

Thus, multi-layered protection combining cryptographic techniques, anomaly detection, and robust fallback mechanisms is the emerging standard for EPS cybersecurity.

D.1.4 Integrated Analysis of Cybersecurity and Functional Safety for EPS

In EPS development, it is crucial to harmonize cybersecurity requirements with functional safety requirements.

This necessity arises because cybersecurity breaches can directly trigger functional safety hazards.

Traditionally, ISO 26262 focused on random hardware faults.

However, ISO/SAE 21434 emphasizes that intentional malicious actions must also be considered in risk assessments.

For EPS specifically:

- The main safety goal is to ensure "proper provision of steering assist without unintended behavior."
- However, cybersecurity breaches (e.g., malicious command injection) could compromise this goal.
- For instance, "attackers disabling EPS functionality" is analogous to "hydraulic steering system failure" and is classified as a critical ASIL-D level hazard.

Thus, hazard analysis must be performed in a way that:

- Considers both hardware failures and cybersecurity attacks equally as causes for safety hazards.

Reflecting this perspective, the concept of Cybersecurity and Safety Co-Engineering has gained traction in recent years.

Specifically:

- Traditionally separated HARA (Hazard Analysis and Risk Assessment) and TARA (Threat Analysis and Risk Assessment) are now being integrated.
- Both analyses are mapped onto a common risk matrix for evaluation.

In an EPS example:

- If "loss of steering assist" is identified as a critical event,
- Then:
 - If caused by a hardware fault → Apply conventional ASIL-D countermeasures (e.g., hardware redundancy),
 - If caused by a cyberattack → Apply cybersecurity measures at Cybersecurity Assurance Level (CAL) 4 (e.g., communication authentication, IDS implementation).

Both safety and security measures are designed together to achieve robust EPS operation.

However, trade-offs between safety and security occasionally arise.

A typical example involves system behavior during faults:

- From a safety viewpoint, immediate transition to a safe state (fail-safe) is desirable,
- But from a cybersecurity perspective, aggressive fail-safe transitions may be exploited by attackers to launch Denial-of-Service (DoS) attacks that immobilize vehicles.

Thus, for EPS:

- Even when an abnormality is detected,
- Where possible, degraded operational modes (limp-home functionality) are preferred,
- Rather than abrupt cessation of steering assist.

For example:

- In the event of a torque sensor fault, the system may maintain a default steering assist torque to allow minimal controllability.

- During communication errors, the last valid command may be retained temporarily and the assist level smoothly decayed.

These techniques enable EPS systems to balance safety and security objectives.

Organizationally:

- Process integration is also advancing.
- ASPICE includes continuous risk management in its support processes,
- And some companies are incorporating cybersecurity risks into joint safety-security reviews during development cycles.

Furthermore, in Europe, UN ECE regulations now require OEMs to establish a Cybersecurity Management System (CSMS) for vehicle type approvals.

Proof of ISO/SAE 21434-compliant development processes is now mandatory.

In EPS development projects:

- Functional Safety Managers and Cybersecurity Managers are increasingly collaborating to ensure that safety and security requirements are implemented consistently and coherently.

In conclusion:

- Cybersecurity countermeasures for EPS are no longer viewed as separate from functional safety measures.
- Instead, an integrated systems engineering approach combining both domains is becoming essential.
- Only by such an approach can truly "safe even under cyberattacks" EPS systems be realized.

[9-10], [9-11],[9-12]

Appendix E

ASIL - Automotive Safety Integrity Level

As the discourse surrounding active safety systems and autonomous driving technologies continues to expand, the emphasis on functional safety requirements has become increasingly prominent. However, for many individuals both within and outside the automotive sector, the precise definition and application of functional safety standards remain complex. This appendix article examines a fundamental aspect of these requirements, namely ASIL-D.

ASIL-D (Automotive Safety Integrity Level D) is a critical component within the broader context of ISO 26262, the international standard for functional safety of electrical and electronic systems in road vehicles. As the most stringent classification level within ISO 26262, ASIL-D necessitates the highest standards of risk mitigation. Systems and components assigned to ASIL-D must be developed under rigorous safety criteria to ensure functional reliability.

Initially published in 2011 and subsequently revised in 2018, ISO 26262 provides detailed process and design recommendations applicable throughout the complete product lifecycle of road vehicles. The standard defines four levels of Automotive Safety Integrity Level (ASIL), ranging from ASIL-A (minimal risk reduction required) to ASIL-D (maximum risk reduction required). An additional classification, Quality Management (QM), applies to systems where no supplementary risk reduction beyond standard industry quality practices is necessary.

The determination of an ASIL classification for a given system is based on the evaluation of three principal factors: Severity, Exposure, and Controllability.

Severity assesses the potential impact on vehicle occupants, pedestrians, or other road users should a system failure occur.

Exposure measures the probability that operational conditions conducive to a hazard will arise concurrently with a system failure.

Severity	Exposure	Controllability		
		C1 (Simple)	C2 (Normal)	C3 (Difficult, Uncontrollable)
S1 LIGHT AND MODERATE INJURIES	E1 (Very low)	QM	QM	QM
	E2 (Low)	QM	QM	QM
	E3 (Medium)	QM	QM	A
	E4 (High)	QM	A	B
S2 SEVERE AND LIFE THREATENING INJURIES – SURVIVAL PROBABLE	E1 (Very low)	QM	QM	QM
	E2 (Low)	QM	QM	A
	E3 (Medium)	QM	A	B
	E4 (High)	A	B	C
S3 LIFE THREATENING INJURIES, FATAL INJURIES	E1 (Very low)	QM	QM	A
	E2 (Low)	QM	A	B
	E3 (Medium)	A	B	C
	E4 (High)	B	C	D

QM (Quality Management)
Development supported by established Quality Management is sufficient.

lowest ASIL
A Low risk reduction necessary
B
:
C
highest ASIL
D High risk reduction necessary

Figure Appendix E.1 ASIL categories

Controllability evaluates the likelihood that the driver or nearby individuals can prevent harm through appropriate and timely action in the event of a malfunction.

A system is classified as ASIL-D when it operates under conditions of high exposure (i.e., scenarios representing more than 10% of operational time), where a failure could result in high severity outcomes (e.g., serious injury or fatality), and where controllability is considered very low (i.e., fewer than 10% of affected individuals could successfully mitigate the risk).

Several illustrative examples provide clarity regarding these classifications.

Consider a malfunction in the dashboard speed indicator. If the indicator fails by displaying no information (e.g., remaining at zero), the failure is easily perceptible to the driver, who can adapt driving behavior accordingly. Such a failure would typically be classified under Quality Management (QM). Conversely, if the speedometer inaccurately displays a lower speed (e.g., 55 mph instead of 65 mph), the failure is less apparent but still controllable by an attentive driver, warranting an ASIL-A or ASIL-B classification.

In contrast, a failure of the vehicle's braking system during high-speed operation in congested urban conditions would pose a critical threat to vehicle control and occupant safety. Given the severity of potential consequences and the reduced ability to avoid harm, such a failure would be classified as ASIL-D.

Alternatively, if the same braking system failure occurred at low speed on an uncongested street, the driver would have a greater opportunity to slow down and bring the vehicle to a safe stop, resulting in a lower ASIL classification.

Systems integral to steering or braking operations must be designed to accommodate a broad range of scenarios, many of which necessitate compliance with ASIL-D requirements due to their critical role in maintaining vehicular control and occupant safety.

Given the fundamental role of steering and braking systems in ensuring safe vehicle operation, their development is subject to the most rigorous scrutiny under ISO 26262. Failures in these systems, whether during routine driving or emergency maneuvers, pose substantial risks, justifying their frequent classification at the ASIL-D level.

The significance of ASIL-D compliance extends beyond mere designation. Attaining ASIL-D certification necessitates that a component or system is subjected to one of the most exhaustive safety development

processes within the automotive industry. The process commences with initial hazard analysis and risk assessment activities aimed at identifying potential safety issues prior to product realization. Subsequently, stringent design controls, redundant architectures, and comprehensive validation activities are employed to guarantee system robustness even in the presence of faults.

Designing for ASIL-D encompasses the entire development of lifecycles. Development teams must systematically document their design rationales, implement proven safety mechanisms, and integrate robust fault detection and fault handling strategies. Verification activities must cover normal operation scenarios, boundary conditions, and deliberate fault injection tests to validate system resilience. Complete traceability from initial requirements through implementation to validation is mandatory and must be verifiable. [9-13], [9-14], [9-15]