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Advanced Modeling and Control of Electrically Excited Synchronous Traction Motors

By

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Declaration

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2026

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Abstract

This dissertation focuses on advanced modeling and torque control of Electrically Excited Synchronous Machines (EESMs) for high-performance automotive traction applications. Driven by the necessity to reduce reliance on rare-earth-based Permanent Magnet Synchronous Machines (PMSMs), this research investigates EESMs as a competitive, sustainable, and high-efficiency alternative for traction electric powertrains. The work addresses the inherent challenges introduced by the traction EESMs, focusing on nonlinear magnetic saturation, significant armature reaction, and the dynamic cross-coupling between stator and rotor circuits.

The scientific contribution is structured around four fundamental pillars. First, a robust three-stage experimental identification procedure is proposed to accurately characterize the stator and rotor flux linkage maps. This methodology identifies both stator and rotor flux linkages across the complete operating envelope by: time-domain integration of the rotor back-EMF during continuous sinusoidal excitation; implementation of a Particle Swarm Optimization (PSO) routine to eliminate fictitious hysteresis loops caused by thermal resistance drifts and inverter non-linearities; and determination of the integration constant through magnetic symmetry during open-rotor stator current reversals.

Second, a novel Full-Voltage-Behind-Reactance (Full-VBR) model is developed to provide a circuital digital twin capable of accurately replicating electromagnetic transients and nonlinear saturation effects. Unlike traditional models, this formulation partitions the machine into a dq -frame stator subsystem and explicit rotor state-space equations, allowing for the analytical identification of stator leakage inductance and the effective current ratio under deep saturation conditions.

Third, a Unified Torque Controller (UTC) architecture is introduced, providing a modular platform that integrates different Torque Control Approaches (TCAs), including Current Vector Control (CVC), Direct Torque Control (DTC), Direct

Flux Vector Control (DFVC), Flux Polar Control (FPC), and Flux Vector Control (FVC). The UTC employs a dual-torque reference strategy (T^* and T_f^*) to decouple stator and rotor regulation paths, optimizing global Joule losses while incorporating dynamic thermal derating to protect the powertrain components across all operating regimes.

Finally, a self-calibrated Dead-Beat Direct Flux Control (DB-DFC) methodology is proposed to be integrated in the UTC framework with minor control modules changes. By directly regulating the stator and rotor flux linkages in discrete-time, this "plug-and-play" approach eliminates the need for extensive inner-loop calibration and gain scheduling. The DB-DFC is rigorously derived in stationary $\alpha\beta$, rotor dq , and stator dqs coordinates, demonstrating superior performance and stability even at low sampling-to-fundamental frequency ratios.

Experimental validation, conducted on a 100 kW commercial traction EESM using a dSPACE MicroLabBox platform, confirms the high-fidelity of the identified maps and the robustness of the proposed control schemes. The results demonstrate a competitive torque regulation dynamic and accuracy across the entire speed range. However, different behavior are highlighted according to the different TCA implemented. This dissertation establishes a comprehensive methodological approach that enables the full performance exploitation of EESMs, providing a definitive benchmark for the next generation of electrified automotive powertrains.

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Chapter 1

Introduction

The imperative to achieve net-zero greenhouse gas (GHG) emissions by mid-century is increasingly acknowledged by governments worldwide, motivated by rising environmental concerns and the evolution of more rigorous legislative frameworks [1]. According to the most recent International Energy Agency report [1], global CO₂ emissions derived from energy combustion and industrial processes reached a record peak of 37.8 Gt in 2024. Furthermore, atmospheric CO₂ concentrations reached an unprecedented level of 422.5 ppm, approximately 50% above pre-industrial levels.

The transportation sector remains a significant contributor, accounting for approximately 15% of global energy-related CO₂ emissions [2]. As illustrated in the latest sectoral breakdown (see Fig. 1.1), emissions from transport have maintained a critical share. Notably, within the European Union, transport represents the only major sector exhibiting GHG emissions still above 1990 levels [3]. Consequently, this persistence necessitates rapid decarbonization strategies, directing scientific policy and technological development toward electrification and sustainable mobility.

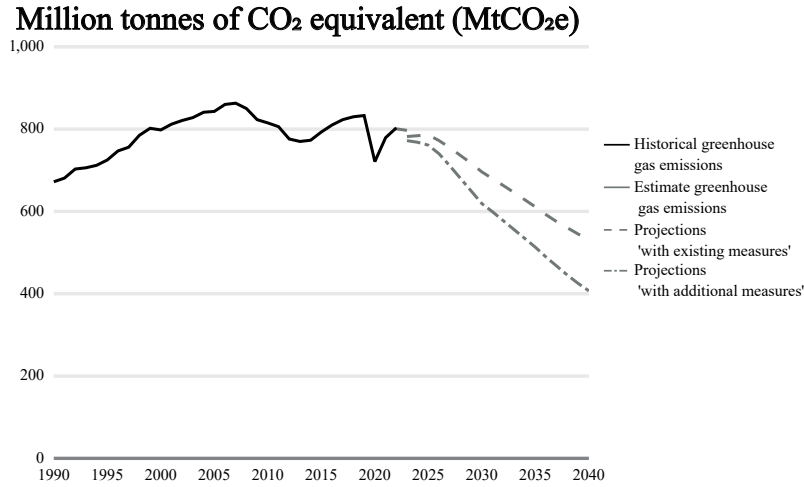


Fig. 1.1 Greenhouse gas emissions from transport in Europe [3].

1.1 Global sectoral distribution of CO₂ emissions

The analysis of the sectoral allocation of global CO₂ emissions is critical for defining decarbonization priorities and establishing the research context for sustainable propulsion technologies. According to recent comprehensive datasets provided by [1, 4, 5, 2, 6] the sectoral breakdown of the total worldwide GHG emissions are detailed as follows, as illustrated in Fig. 1.2.

- **Electricity and Heat Production:** This sector represents the largest single source, accounting for approximately 34% of global CO₂ emissions derived from fossil fuel combustion.
- **Industry:** Accounting for about 24% of emissions, this sector includes process emissions derived from cement, steel, and chemicals, along with direct energy usage.
- **Agriculture, Forestry, and Other Land Use:** This sector constitutes about 22% of emissions, primarily attributed to livestock, crop cultivation, and deforestation.
- **Transport:** It accounts for approximately 15-17% of the total, resulting almost exclusively from the combustion of petroleum-derived fuels in road, air, and marine transport.

- **Buildings:** Direct on-site fuel consumption comprises about 6% of emissions. However, this share increases when indirect emissions from electricity consumption are considered.
- **Waste and Wastewater:** It represents around 3% of total GHG emissions, comprising methane and nitrous oxide from landfills and wastewater treatment plants.

Furthermore, the transport sector is dominated by road vehicles, which account for approximately 74% of the total transport-related CO₂ emissions in 2024, while the remainder is attributed to shipping, aviation, and rail [7].

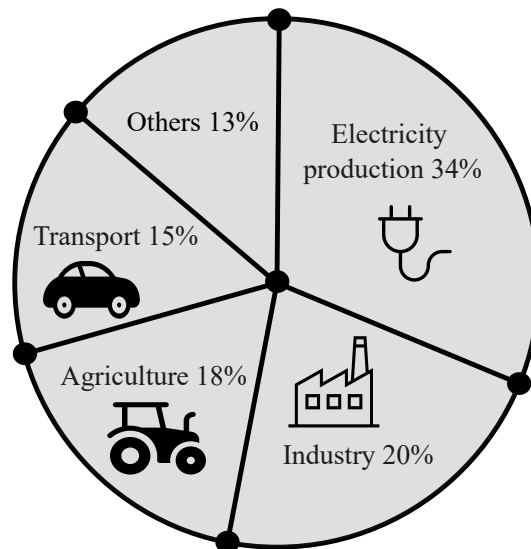


Fig. 1.2 Global sectoral distribution of CO₂ emissions.

1.2 Rare-earth elements in electric transport

Rare-earth elements (REEs), encompassing 17 chemically similar elements, are recognized as essential components for electric motors employed in transport electrification, owing to their superior magnetic characteristics [8]. Specifically, neodymium-iron-boron (NdFeB) magnets, predominantly manufactured using REEs, facilitate the production of compact, high-torque-density permanent magnets. Consequently, these materials represent the fundamental technology enabling the majority of state-of-the-art traction machines employed in electric mobility [9, 10].

However, the extraction and refinement processes of REEs are associated with significant environmental implications. Indeed, mining and processing operations generate substantial air and water pollution, along with radioactive by-products, leading to long-term ecological degradation [11, 12]. Furthermore, these environmental and social externalities are exacerbated by the currently low rates of REE recycling and the limited implementation of closed-loop supply chains within industrial practices [13].

A further critical concern involves the persistent volatility of REE markets, which exhibit high vulnerability to rapid regulatory shifts, export limitations, and geopolitical tensions. Currently, China maintains a dominant position, accounting for approximately 70% of global annual REE production and possessing over a third of global reserves [14], as illustrated in Fig. 1.3. This geographical concentration significantly exacerbates supply risks and price instability. Furthermore, the demand for REEs is projected to increase rapidly, driven by the sustained expansion of electric transportation, renewable energy, electronics, and defence sectors [14].

Consequently, these challenges require the development, validation, and deployment of alternative machine technologies capable of satisfying the environmental, economic, and strategic sustainability requirements essential for the long-term decarbonization of global transport.

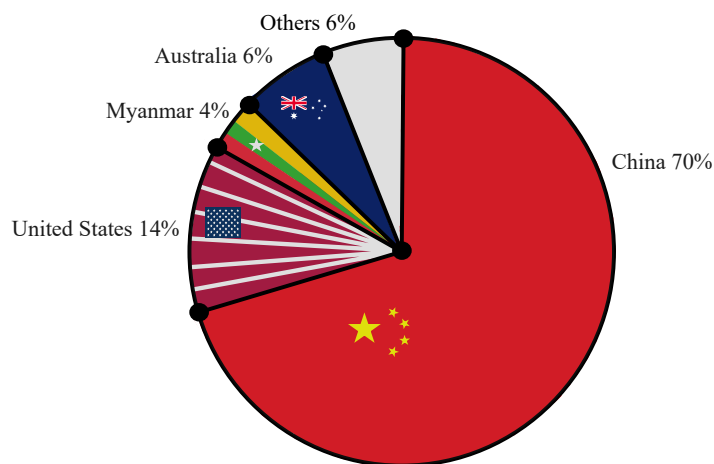


Fig. 1.3 Producers of rare-earth elements.

1.3 Electrical machines without REE

While interior permanent magnet (IPM) motors currently dominate the market due to their superior performance, their deployment depends on the extraction of rare-earth materials [15]. To achieve independence from these materials, attention has increasingly shifted toward induction motors (IM), synchronous reluctance motors (SynRM), switched reluctance motors (SRM), permanent magnet-assisted synchronous reluctance motors (PMSR), ferrite-based PM designs, and wound field solutions [16, 17].

Induction motors (IM) are characterized by a robust, low-maintenance architecture, rendering them a preferred solution for applications with moderate power and torque requirements. Torque production is achieved through the interaction of the stator flux and currents induced in the rotor cage, a mechanism capable of generating force solely due to the slip speed relative to the stator magnetomotive force [16]. However, the generation of these induced currents results in significant rotor copper losses. Consequently, this phenomenon leads to increased thermal stress on components and a reduction in the machine's overall efficiency compared to synchronous topologies.

Synchronous reluctance motors (SynRM) achieve torque production solely through magnetic reluctance, leveraging the rotor geometry without relying on magnets or rotor windings. Recent advancements have enhanced their robustness and energy efficiency, positioning the SynRM as a viable candidate for sustainable mobility, despite constraints regarding the flux-weakening range and reduced power factor [16, 17].

Permanent magnet-assisted synchronous reluctance motors (PMSR) integrate small ferrite or non-rare-earth magnets into the rotor flux barriers. This configuration improves the power factor and torque density compared to SynRM architectures, while largely avoiding the use of rare-earth magnets [18, 17]. Consequently, these machines are increasingly recognized for their balance between cost, sustainability, and drive performance.

Switched reluctance motors (SRM) represent a highly robust and magnet-free alternative, relying on the alternate energization of stator windings to sequentially align the rotor poles. While offering exceptional fault tolerance and reliability, the

SRM is challenged by pronounced torque ripple and acoustic noise profiles, which limit its suitability for passenger vehicle applications [17, 15].

Ferrite-based PM motors employ magnets composed of iron oxides rather than rare-earth elements. Although these motors typically exhibit lower flux density and torque, they offer substantial benefits in terms of environmental sustainability, cost reduction, and supply chain stability [18, 15].

Electrically excited synchronous motors (EESM), also known as wound-field synchronous motors (WFSM), are a prominent magnet-free option. By incorporating a controllable field winding on the rotor, the EESM enables real-time regulation of the excitation, resulting in excellent low-speed torque, a wide constant power operating range, and a tunable power factor. Furthermore, the elimination of rare-earth magnets, combined with increasing reliability due to innovations in brushless excitation systems, positions the EESM as a top-tier sustainable solution [16, 17, 19].

Comparative scientific studies and experimental results confirm that, although rare-earth IPM motors currently offer the highest torque density and efficiency, the performance gap is narrowing. Indeed, advances in IM, SynRM, PMaSynRM, and especially EESMs, support a significant shift toward sustainable electrified transport [16, 18, 15, 17] (see Table 1.1).

Table 1.1 Feature comparison among alternative traction machine types for passenger EVs. ● = Excellent, ○ = Average, . = Poor.

Feature	Machine Topology				
	EESM	Ferrite-PM	SynRM	IM	SRM
Torque/power density	○	○	.	○	○
Rated operation efficiency	●	○	.	○	○
Partial-load efficiency	○	○	●	●	○
Low cost	●	○	●	○	●
Low noise/torque ripple	●	○	○	●	.
Fault Tolerance	●	.	●	●	●

1.4 Electrically excited synchronous machines in traction applications

EESMs represent a true rare-earth-free topology for next-generation automotive propulsion and heavy-duty traction applications [19, 20]. Structurally, this machine class is fundamentally distinguished from conventional PM architectures by the incorporation of a dc-excited rotor winding, as shown in Fig. 1.4.

Consequently, this configuration grants an additional degree of control freedom. Unlike PM machines, where the excitation flux is fixed, the EESM allows for the active regulation of the air-gap flux via the rotor field current. This capability thereby enables significant operational flexibility, facilitating the optimization of efficiency and performance across the entire torque-speed range [20, 15].

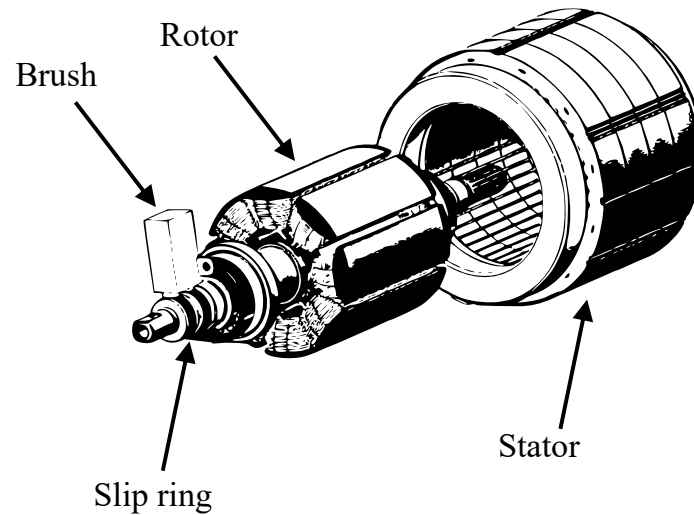


Fig. 1.4 Three-dimensional view of an EESM with slip rings.

1.4.1 EESM Pros and Cons

A primary advantage of the EESM topology lies in its capability to generate high torque at low speeds while maintaining moderate stator current requirements, thereby overcoming limitations inherent to fixed-flux machines [21, 22]. The independent regulation of rotor excitation facilitates effective flux-weakening operations, consequently extending the constant power speed range and ensuring high efficiency at elevated speeds [21, 22]. Furthermore, this additional degree of control enables

operation at a high power factor, optimizing inverter utilization and mitigating the overall cost of the traction system [22]. From a safety perspective, the capability to de-energize the rotor winding eliminates the risk of uncontrolled back-electromotive force (back-EMF) during fault conditions. This feature represents a significant advantage over PM designs, which are susceptible to persistent fault voltages and demagnetization risks [23].

However, despite these merits, the adoption of EESMs introduces specific technical challenges, primarily related to the supply of field current. This requirement leads to increased rotor losses and thermal management complexities, often exacerbated by constrained cooling pathways within the rotor structure [24–26]. While conventional brush-and-slip-ring systems are technologically mature, they are associated with mechanical wear and maintenance requirements that may constrain their suitability for automotive applications [25, 26]. Conversely, alternative brushless and wireless excitation topologies, such as rotary transformers and capacitive coupling, offer enhanced durability. Nonetheless, these solutions typically introduce increased system complexity and may compromise dynamic responsiveness [27, 28].

1.4.2 Excitation systems

The architecture of the excitation system significantly influences the performance, controllability, and maintenance intervals of the EESM [29]. Consequently, the design process must address various technical challenges, including response speed, de-excitation capabilities, and field-forcing requirements. Generally, the most prevalent excitation system topologies are categorized into static, brushless, and embedded systems (see Fig. 1.5).

Static exciters are predominantly utilized in large synchronous generators with power ratings in the order of several MVAs, where adherence to standardized dynamic performance requirements is mandatory [30–33]. In these systems, the de-excitation function is typically executed by applying a negative ceiling voltage or through the utilization of a discharge resistor.

Brushless exciters feature a rotating exciter with a three-phase winding mounted directly on the main machine shaft [34–37]. The exciter field winding is located on the stator poles and supplied by a DC source. During shaft rotation, the main air-gap field induces electromotive forces and, consequently, currents within the exciter rotor

phases. A rotating AC-to-DC rectifier, integrated onto the exciter rotor, converts the induced AC current into the DC current required to feed the main EESM rotor field winding. This topology represents a viable solution for applications that require flexible excitation control while minimizing maintenance and operational costs.

Embedded exciters represent an exciterless alternative wherein the excitation components are integrated directly into the main EESM stator [38–42]. In this configuration, excitation power is generated by an auxiliary single-phase harmonic winding embedded within the stator, which induces an electromotive force from the rotating field harmonics in the air-gap space.

The selection of the optimal system architecture remains strictly dependent on specific application requirements. While classical static exciters remain the preferred choice for the majority of large generators (power levels of order MVA), novel brushless topologies are emerging as competitive alternatives. Conventional brushless exciters dominate the market for smaller EESMs; however, harmonic-integrated exciters are enabling the development of more compact solutions. Furthermore, advancements in rotating power electronics and wireless signal transmission are expected to drive the continued evolution of excitation systems to meet increasingly stringent requirements.

In the automotive sector, traditional brush-and-slip-ring configurations, such as those employed in the Renault Zoe, provide robust excitation methods, at the expense of periodic maintenance cycles [20, 25]. Conversely, cutting-edge designs in both industry and research increasingly prioritize brushless or contactless solutions. Notable implementations by Brusa and BMW leverage rotary-transformer-based and capacitive-field excitation technologies [43, 44, 28]. These advanced methodologies fundamentally mitigate mechanical wear, thereby enabling higher reliability under demanding drive cycles [27, 28].

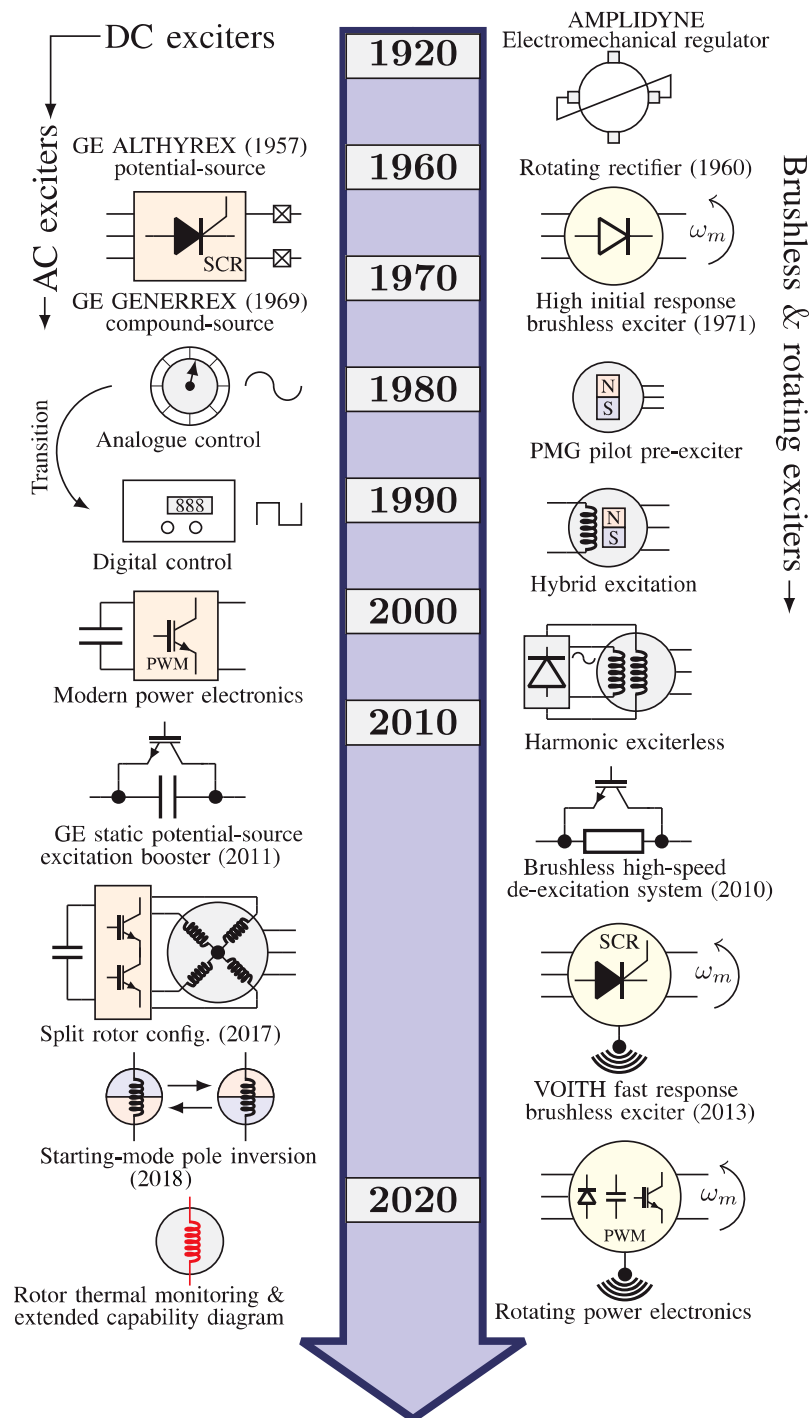


Fig. 1.5 Development history of excitation systems for wound-field synchronous machines [29].

1.5 Digital twin modeling

The concept of the digital twin has emerged as a paradigm-shifting methodology in the design, optimization, and real-time monitoring of electric drive systems. Fundamentally, a digital twin is a high-fidelity virtual counterpart of a physical asset, maintaining dynamic synchronization with the system's state, behavior, and performance characteristics via the integration of advanced sensing, physical modeling, and adaptive data-driven algorithms (see Fig. 1.6).

In the specific context of electric traction drives, digital twin frameworks facilitate predictive analytics, virtual commissioning, control strategy optimization, and fault diagnostics, thereby obviating the immediate reliance on physical prototyping or intrusive experimental validation. Regarding electrical machines, these virtual models provide a comprehensive simulation environment capable of accurately reproducing complex electromagnetic behaviors, thermal dynamics, mechanical transients, and the intricate interactions between the power converter and the motor.

Such high-fidelity modeling is of paramount importance for EESMs designed for traction applications, where inherent multi-physics attributes and highly nonlinear magnetic cross-coupling effects pose significant challenges to conventional modeling strategies. Precise digital twin implementations enable the rigorous validation of advanced control algorithms under realistic operating regimes, accounting for deep magnetic saturation and cross-coupling phenomena. Consequently, digital twin are increasingly recognized as indispensable instruments in the design and development of next-generation traction control systems.

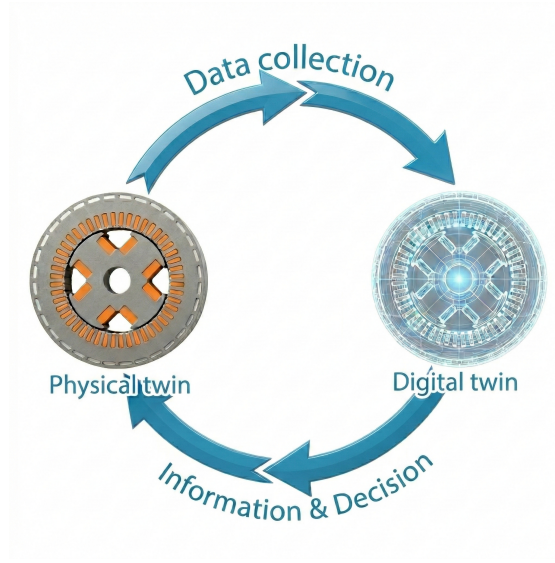


Fig. 1.6 Digital twin of an EESM: workflow.

1.5.1 EESM digital twin modeling

The accurate characterization of nonlinear magnetic behavior in PMSMs typically relies on the construction of current-to-flux and flux-to-current relationships [45–47]. These maps, referred to as *direct* and *inverse* flux maps, provide a comprehensive framework for modeling saturation, cross-saturation, and spatial harmonics. Their ability to capture the full multidimensional flux surfaces makes them essential when standard linear machine models are insufficient.

The growing interest in rare-earth-free traction technologies has driven the extension of these mapping techniques to EESMs [48–51]. Unlike PMSMs, EESMs feature an actively controlled rotor winding, which introduces an additional degree of freedom and significantly increases the complexity of the magnetization characteristics. Consequently, the multidimensional flux maps must account for the coupling between all stator and rotor currents, requiring advanced identification and interpolation methods.

For comprehensive simulation of electric powertrains, digital twin frameworks require machine models that accurately represent both converter-switching dynamics and electromagnetic transients. Circuit-oriented modeling approaches are particularly suitable for this purpose, as they facilitate integration into system-level simulations. The Voltage-Behind-Reactance (VBR) formulation [52] is a practical approach that

partitions the machine into stator and rotor subsystems. In this structure, the stator equations are defined in the dq frame, with flux linkages as state variables, and expressed in circuit form. In contrast, explicit state-space equations describe the rotor dynamics. This modular approach enables seamless integration into simulation environments such as PLECS [53], where the machine windings are represented by branches that can be directly connected to the power converter circuit with the desired topology [54].

Historically, EESMs have been primarily utilized as synchronous generators in industrial and power generation sectors [55–58]. In these applications, magnetic nonlinearity was typically approximated using one-dimensional saturation curves relating the magnetizing current to the flux linkage [59–61]. While adequate for generator operation, these simplified models fail to capture the complex cross-saturation and magnetic saturation effects inherent in EESMs designed for traction applications [62–64]. Therefore, there is a clear need for VBR-based EESM models that can account for the full nonlinear flux interactions across all stator and rotor axes.

1.6 Torque control

In recent years, the transport electrification sector has significantly accelerated the advancement of electric drives (eDrives), facilitating their widespread integration across the automotive [65, 66, 10, 67], railway [68, 69], and aircraft [70–72] industries. Currently, eDrives utilizing both synchronous and induction motors represent the standard for state-of-the-art of hybrid and fully electric vehicles [73–75]. Consequently, this integration introduces novel and stringent requirements for the associated torque controllers for eDrives, particularly within the demanding context of automotive applications [75].

Traditional industrial adjustable-speed drives typically entail significant simplifications. In these applications, the AC motor operates exclusively in speed-control mode, where high-accuracy torque control is not strictly required since steady-state torque errors are automatically compensated by the speed control loop. Furthermore, the requirement for high dynamic performance is minimal, as rapid and large torque variations are infrequent. This characteristic facilitates the calibration of inner control loops, often neglecting magnetic saturation effects, and enables the implementation

of straightforward sensorless techniques for torque regulation [76–79]. Conversely, for automotive eDrive and traction applications, the technical requirements are considerably more rigorous [80]:

1. The significant magnetic saturation inherent in automotive electrical machines cannot be neglected; rather, it must be accounted for through rigorous identification procedures to fully characterize the machine’s magnetic model, specifically the current-to-flux-to-torque relationships [81, 82].
2. Effective torque regulation is required, with steady-state errors maintained within rigorous tolerances. Furthermore, the torque reference must be constrained by the inverter voltage and current limits.
3. The performance of every inner control loop must remain independent of the operating point (torque-speed), thereby avoiding the need for controller tuning procedures that require dedicated test rigs [83].
4. The controller is required to sustain high dynamic performance under all operating conditions, including deep flux-weakening (FW) regions with maximum torque per voltage (MTPV) limitation [74, 84].
5. The control architecture should minimize the number of modules to facilitate the automatic code generation process.

According to these requirements, several torque control approaches (TCAs) have been established as benchmarks for traction applications.

- *Current Vector Control (CVC)* regulates torque by directly controlling the dq components of the stator current (i_d, i_q), almost universally implemented via Field Oriented Control (FOC) in rotor coordinates [83, 85–100]. Look-up tables-based CVC controllers in automotive contexts often use multidimensional tables to provide dq current references directly (see Fig. 1.7), incorporating all drive limits (maximum current, DC-link voltage, MTPV) [88].

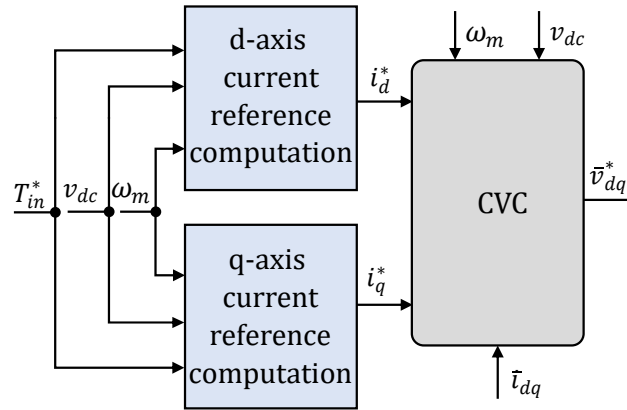


Fig. 1.7 CVC scheme.

- *Direct Torque Control (DTC)* directly regulates the amplitude of the stator flux vector and the torque (λ, T), generally implemented with inner hysteresis regulators and switching tables [101–107]. Can be integrated with pulse-width modulation (PWM) [108, 105], as shown in Fig. 1.8.

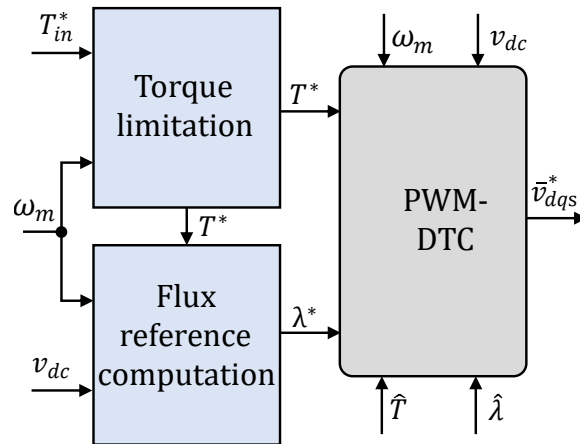


Fig. 1.8 PWM-DTC scheme.

- *Direct Flux Vector Control (DFVC)* controls both the stator flux amplitude and the torque-producing current (λ, i_{qs}) [109–115] (see Fig. 1.9), synthesizes the strengths of both CVC and DTC as a high-dynamic, cross-machine control solution [109].

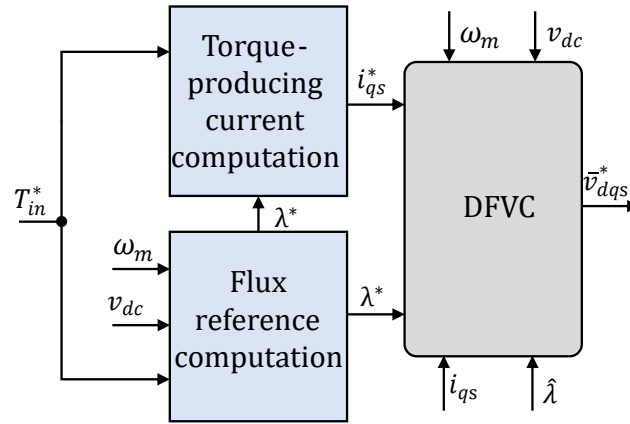


Fig. 1.9 DFVC scheme.

- *Flux Polar Control (FPC)* regulates both the amplitude and the load angle of the stator flux vector (λ, δ) [80] (see Fig. 1.10). Recently developed, provides a methodology independent of machine inductance.

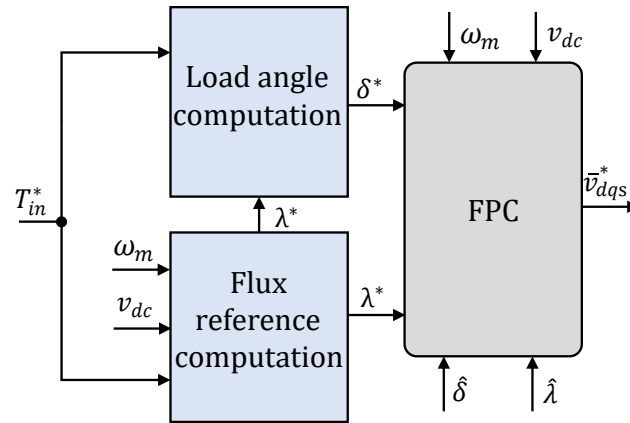


Fig. 1.10 FPC scheme.

- *Flux Vector Control (FVC)* determines torque by controlling the dq -components of the stator flux vector (λ_d, λ_q) [116] (see Fig. 1.11). Also providing the advantage of independency from machine's working point.

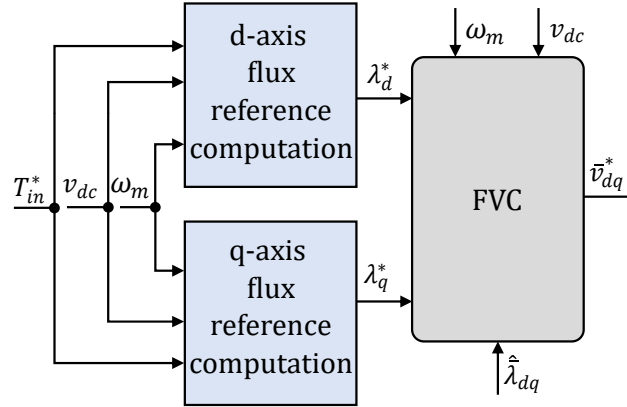


Fig. 1.11 FVC scheme.

CVC strategies represent the dominant solution in industrial applications. However, conventional industrial implementations often rely on external voltage regulation loops to manage FW operation. This approach typically necessitates heuristic calibration procedures and may compromise both torque accuracy and dynamic performance [91–99]. Conversely, advanced CVC schemes, particularly those tailored for automotive traction, employ multidimensional Look-Up Tables (LUTs) to determine the dq -current references. This methodology ensures accuracy of torque regulation while strictly adhering to the drive's operational limits [83, 85–91, 95, 100]. In contrast, DTC and DFVC schemes inherently obviate the need for LUTs to achieve torque linearization, even during FW operation with MTPV constraints. This advantage arises from their implementation in the stator dqs reference frame [109, 110, 115], where the regulation of stator flux amplitude and electromagnetic torque is fully decoupled [110]. Consequently, this decoupling significantly simplifies the management of voltage margins and the formulation of the underlying control equations. However, a challenge affects the aforementioned control strategies: tuning inner control loops to ensure dynamic invariance with respect to magnetic saturation. Indeed, the regulator gains are intrinsically dependent on the local differential inductance and the machine's saturation behavior [117, 45]. While adaptive gain tuning and online differential inductance estimation techniques provide partial solutions for specific cases [118, 116], a comprehensive methodology ensuring robust large-signal performance under rapid operating point transitions (e.g., FW and MTPV) is currently absent in the literature for these TCAs. Consequently, extensive and time-consuming calibration procedures remain a necessity. Notably, while DTC and DFVC remain the most widely recognized unified solutions

for AC motor torque control [102, 105, 109], they typically require heuristic outer controllers for MTPV and load-angle management, entailing significant calibration efforts.

The FPC scheme integrates the benefits of DTC and DFVC: unified control, decoupled voltage margins, and high dynamic bandwidth, while introducing torque regulation independence from inductance and saturation [119, 80]. Furthermore, FPC regulator gains depend exclusively on the digital controller parameters and the desired closed-loop bandwidth [119, 80], thereby establishing a robust, tuning-less framework applicable to all AC motor topologies. Similarly, FVC exhibits independence from magnetic saturation phenomena. However, its implementation in the rotor dq reference frame yields coupled voltage margins, which limit its performance relative to stator-frame strategies. Despite their effectiveness, these schemes typically employ proportional-integral (PI) regulators. Therefore, their dynamic performance is fundamentally constrained by the regulator bandwidth, which is ultimately determined by the digital controller sampling frequency.

Dead-beat (DB) control strategies are based on the utilization of an inverse machine model to compute the requisite stator voltage vector necessary to nullify the error between the actual and reference values of torque within a two sampling intervals. Consequently, the calculated voltage command is derived over the switching period using PWM techniques. These strategies have been established as highly effective solutions for minimizing torque response latency, theoretically achieving the target reference within two sampling intervals [120, 121], provided that the power converter is capable of delivering the required voltage. Furthermore, DB controllers are characterized by robust operation under low sampling-to-fundamental frequency ratios and the capability to enforce strict voltage and current constraints, thereby offering superior performance compared to conventional synchronous reference frame PI controllers. However, emerging DB solutions designed to maximize dynamic performance [122–125] typically necessitate detailed knowledge of machine parameters and reliable predictive logic for stator flux and torque [125]. To mitigate this parameter sensitivity, the Dead-Beat Direct Flux Control (DB-DFC) strategy [126] has recently emerged as a viable alternative, distinguished by its independence from the machine's inductances.

1.6.1 EESM torque control

The EESM distinguishes itself from PMSM topologies by providing an additional degree of control freedom through the adjustable rotor field current. This feature facilitates sustained operation in the constant power region over an extended speed range, while simultaneously permitting the optimization of efficiency and FW performance [62]. Consequently, the automotive industry is increasingly investigating EESM architectures as a viable, rare-earth-free alternative for high-performance traction applications [127, 20, 50].

Despite these favorable attributes, the implementation of effective torque control in EESMs imposes significant challenges [128]. A primary complexity arises from the inherent magnetic cross-coupling between the stator d -axis and the rotor excitation circuit, which precludes decoupled current regulation and necessitates the adoption of advanced multivariable control strategies [129]. In traction-oriented designs, the rotor winding typically exhibits an inductance value with orders of magnitude higher than that of the stator one. This physical characteristic results in a rotor current dynamic response that is considerably slower than that of the stator, even when the exciter operates at its voltage ceiling. Moreover, the pursuit of maximum efficiency often requires continuous modulation of the excitation current, which can further degrade the drive's overall torque dynamic response [130].

To address these dynamic limitations, advanced thermal management strategies have been proposed to permit temporary overloading of the stator currents during transient manoeuvres. This approach leverages the faster dynamics of the stator circuit to compensate for the rotor field's latency [131]. However, to ensure stable and generalizable performance, the control architecture must explicitly account for the non-linear magnetic coupling between the stator and rotor subsystems [132].

Despite the extensive research on AC drive control strategies, few studies have explicitly focused on torque control for EESMs, with most implementations relying on dq current control [133, 134, 130, 131, 135, 128]. A more detailed analysis of EESM torque control is presented in [136], where a CVC scheme is discussed in stator dqs coordinates.

In [137, 138], a deadbeat-direct torque and flux control strategy is proposed, wherein a field flux linkage command is designed to achieve unity displacement power factor in steady-state conditions. However, since machine losses are not neces-

sarily minimized at unity power factor, further optimization algorithms are required to enhance overall efficiency [139, 140, 134, 130, 131, 135, 141]. Moreover, most deadbeat approaches rely on current-estimation methods [121], thereby increasing the complexity of the control architecture.

The performance of the aforementioned TCAs are summarized in Table 1.2. Their characteristics are fundamentally determined by the analytical framework adopted for the machine modeling, which will be deeply analyzed in the next Chapter.

Table 1.2 Feature comparison among TCAs for traction synchronous motors.

● = Strength, ○ = Intermediate, · = Weakness.

Torque Regulation	Control Strategy						
	CVC	PWM DTC	DFVC	FPC	FVC	DB DTC	DB DFC
Accuracy	○	●	●	○	○	○	○
Dynamic performance	·	○	○	●	●	●	●
Simplicity	●	○	○	●	○	○	●
Tuning efforts	·	○	○	●	●	○	●

1.6.2 Thermal Management

A critical issue affecting the reliability of synchronous motors is the potential for failure due to uncontrolled temperature rise [142]. To mitigate this risk, advanced methodologies for thermal analysis, such as lumped parameter thermal networks and Finite Element Analysis (FEA), have been widely adopted [143]. As detailed in [144], the windings represent the most thermally sensitive components, primarily due to the thermal degradation properties of the insulation materials. Notably, exposure to elevated temperatures can severely compromise these materials, potentially reducing their operational lifespan by up to 50% [145].

Consequently, the implementation of robust thermal management strategies for motor windings is imperative in electric vehicle applications [145]. These strategies are essential to safeguard critical components against excessive temperature excursions [146–148]. The prevailing approach described in the literature entails dynamic

derating of current limits for stator windings and, specifically for EESMs, rotor windings, thereby ensuring that operation remains within safe thermal boundaries.

1.7 Research contributions

This dissertation presents a comprehensive framework for the analysis, design, and implementation of advanced modeling and torque control strategies for EESMs in traction applications. The research provides a systematic structure incorporating all essential elements for the thorough analysis and optimal performance exploitation of traction EESMs, spanning from the experimental identification of multidimensional flux linkage maps to the development of a high-fidelity circuital digital twin and a unified torque control architecture. A central theme throughout this work is the rigorous treatment of magnetic cross-coupling and armature reaction, identified as the primary phenomena governing the machine's nonlinear dynamics.

The primary contributions of this research are structured around three fundamental pillars:

- The definition of a robust three-stage experimental identification procedure for 3D flux linkage maps, ensuring high-fidelity characterization across the entire operating range.
- The design of a high-fidelity Full-VBR digital twin capable of accurately replicating nonlinear magnetic saturation, stator leakage flux dynamics, and dynamic cross-coupling effects.
- The development of a UTC architecture designed to optimize global Joule losses while providing a modular and versatile platform for the implementation of diverse TCAs according to variable stator and rotor current limits.

The proposed methodologies were experimentally validated using a 100 kW commercial traction EESM. A systematic comparative analysis of the steady-state torque accuracy and transient performance of the considered TCAs is reported, providing a performance benchmark for modern high-efficiency EESM powertrains.

1.8 Abstracts of the Chapters

This dissertation addresses the identification, modeling, and control of EESMs for automotive traction applications. The research progresses from theoretical formulation and experimental identification procedure (Chapter 2) to advanced digital-twin prototyping (Chapter 3) and high-performance torque controller definition (Chapter 4). The performance of the advanced torque control algorithm is extensively tested experimentally, and the results are presented in Chapter 5. Finally, Chapter 6 concludes the dissertation, highlighting the main findings. The specific contributions of each chapter are summarized below.

1.8.1 Chapter 2 - EESM Modeling and Identification

This chapter establishes the mathematical modeling of EESMs as the foundation for advanced control development. State-space formulations are derived in stationary $\alpha\beta$, rotor synchronous dq , and stator flux-oriented dqs frames, incorporating nonlinearities through a full differential inductance matrix. The modeling is substantiated through a Renault ZOE R135 case study, whose experimental identification of high-fidelity 3D flux linkage maps is described. To circumvent the risks of conventional locked-rotor tests for the rotor flux identification, a three-stage procedure is proposed: (1) numerical integration of back-EMF under sinusoidal excitation, (2) PSO-based compensation for thermal drift and inverter non-linearities, and (3) initial state determination via magnetic symmetry. Validation against FEA and experimental data confirms the accuracy of the identified maps, with a residual rotor current tracking error of less than 3%. These profiles enable the extraction of variable differential inductances for robust, saturation-aware torque regulation.

1.8.2 Chapter 3 - EESM Digital Twin Modeling

This chapter presents a high-fidelity digital twin framework for EESMs that addresses the limitations of conventional models in reproducing deep saturation effects. A novel Full-Voltage-Behind-Reactance (Full-VBR) formulation is proposed, which integrates high-resolution flux maps to compute a state-dependent 4×4 incremental inductance matrix. This approach enables the precise representation of nonlinear electromagnetic behavior and dynamic stator-rotor cross-coupling within a circuit-

oriented simulation environment. Additionally, a rigorous analytical method for extracting stator leakage inductance directly from global flux maps is introduced. The framework is validated through comparative simulation benchmarking and a "digital twin" experimental campaign, confirming the model's predictive accuracy under fast torque transients and armature reaction tests.

1.8.3 Chapter 4 - EESM Torque Control

This chapter presents in detail the development of a Unified Torque Controller (UTC) for traction EESMs. The proposed architecture provides a reconfigurable framework that supports multiple TCAs, including CVC, DTC, DFVC, FPC, and FVC, under varying thermal and electrical constraints. An offline optimization algorithm generates reference trajectories to minimize total copper losses, while a linearization procedure based on normalized LUTs addresses magnetic nonlinearity. Furthermore, a novel "plug-and-play" DB-DFC strategy is proposed. Implemented across the $\alpha\beta$, dq , and dqs reference frames, this predictive approach eliminates the need for extensive regulator tuning, ensuring high-bandwidth torque regulation that is invariant to magnetic saturation.

1.8.4 Chapter 5 - Experimental validation

This chapter presents the comprehensive experimental validation of the proposed UTC structure comprising different PI-regulated TCAs as DFVC, DTC, FVC, FPC, CVC and DB-DFC strategies for EESMs. The validation campaign was conducted on a custom test rig featuring a commercial Renault ZOE R135 motor, controlled via a dSPACE platform. The experimental protocol was designed to rigorously assess the controller's performance across the entire torque-speed range, specifically targeting high-power operation, deep flux-weakening capability, and dynamic robustness under severe constraints. A comparative analysis was performed against all the TCAs considered. Special emphasis was placed on evaluating the "plug-and-play" nature of the proposed DB-DFC architecture, which utilizes FEA derived maps and heuristic tuning to demonstrate robustness against parameter mismatches. Furthermore, the chapter verifies the efficacy of advanced control features, including the online adaptation of current limits for thermal management and active MTPV limitation to prevent machine pull-out.

1.9 List of publications

1.9.1 Journal papers

- (J1) A. Ionta, S. Rubino, F. Graffeo and R. Bojoi, "High-Fidelity Voltage-Behind-Reactance Model of Electrically Excited Synchronous Machines Using Flux Maps," in *IEEE Open Journal of Industry Applications*, vol. 6, pp. 603-618, 2025.
- (J2) A. Ionta, S. Rubino, F. Graffeo, F. Mandrile, R. Bojoi and E. Armando, "Direct Flux and Torque Control of Electrically Excited Synchronous Traction Motors," in *IEEE Transactions on Industry Applications* (Early Access).
- (J3) A. Ionta, S. Rubino, F. Graffeo, F. Stella and R. Bojoi, "A Unified Approach for the Design of Accurate Torque Controllers for Electrically Excited Synchronous Motors," in *IEEE Transactions on Industry Applications* (Early Access).
- (J4) S. Rubino, A. Ionta, F. Graffeo and R. Bojoi, "Deadbeat Direct Flux Control: Plug-and-Play Torque Controllers for Electrically Excited Synchronous Motors," in *IEEE Transactions on Power Electronics* (Early Access).

1.9.2 Conference papers

- (C1) A. Ionta, F. Righetto, M. Saeed, N. Bianchi, R. Bojoi and F. Briz, "Optimal Estimation of the Parameters and Losses of Induction Motors Working in Full-Wave," 2025 IEEE Energy Conversion Conference Congress and Exposition (ECCE), Philadelphia, PA, USA, 2025, pp. 1-7.
- (C2) A. Ionta, S. Rubino, F. Graffeo and R. Bojoi, "Unified Torque Control of Electrically Excited Synchronous Machines for Traction Applications," 2025 IEEE Industry Applications Society Annual Meeting (IAS), Taipei, Taiwan, 2025, pp. 1-8.
- (C3) A. Ionta, S. Rubino, F. Mandrile, F. Graffeo, R. Bojoi and E. Armando, "Direct Flux Vector Control of Electrically Excited Synchronous Machines for Electrical Vehicles," 2024 IEEE Energy Conversion Congress and Exposition (ECCE), Phoenix, AZ, USA, 2024, pp. 6431-6438.

Chapter 2

EESM Modeling and Identification

2.1 Electromagnetic behavior modeling

A synchronous machine can be represented in either the stationary $\alpha\beta$, the rotor dq or the stator dqs coordinates. When adopting the stator frame, the ds -axis is defined to be aligned with the stator flux linkage vector $\bar{\lambda}$, as depicted in Fig. 2.1. The angle separating the stator flux vector from the rotor d -axis is named the load angle δ .

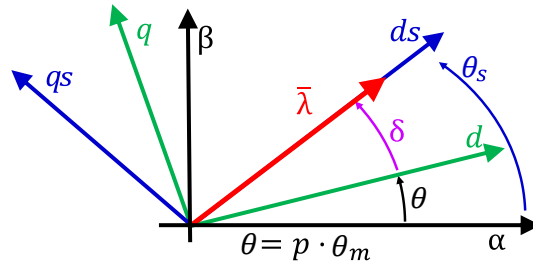


Fig. 2.1 Schematic view of the stationary- $\alpha\beta$, rotor- dq , and stator- dqs frames for a generic synchronous motor.

Control methodologies such as CVC and FVC regulate the stator current and stator flux, respectively, within the rotor dq reference frame. Conversely, approaches such as DTC, DFVC, and FPC operate in the stator dqs reference frame. Accordingly, the electromagnetic equations and the state-space formulations in both dq and dqs coordinate systems for EESM are presented hereafter.

2.1.1 Electromagnetic model in rotor dq coordinates

The stator voltage equations in the rotor dq reference frame can be formulated as in [149], by including the rotor terms:

$$\bar{v}_{dq,f} = \begin{bmatrix} v_d \\ v_q \\ v_f \end{bmatrix} = \begin{bmatrix} R & 0 & 0 \\ 0 & R & 0 \\ 0 & 0 & R_f \end{bmatrix} \cdot \begin{bmatrix} i_d \\ i_q \\ i_f \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \lambda_d \\ \lambda_q \\ \lambda_f \end{bmatrix} + \omega \cdot \begin{bmatrix} -\lambda_q \\ \lambda_d \\ 0 \end{bmatrix} \quad (2.1)$$

where the subscript notation d and q denotes the stator and f the rotor components of the respective voltage "v", current "i", and flux linkage " λ " vectors in the dq reference frame, and ω represents the electrical angular velocity of the rotor.

The corresponding electromagnetic torque is defined by:

$$T = \frac{3}{2} \cdot p \cdot \bar{\lambda} \times \bar{i} = \frac{3}{2} \cdot p \cdot (\lambda_d \cdot i_q - \lambda_q \cdot i_d), \quad (2.2)$$

where p is the number of pole pairs and $\bar{\lambda}$, $\bar{i}$ are the stator flux and current vector, respectively.

For the derivation of the state-space model, it is necessary to explicitly establish the relationships between the stator voltage components in dq coordinates and the chosen state variables. In CVC schemes, torque is regulated by directly manipulating the dq components of the stator current vector. To linearize the current dynamics around a given operating point, the differential inductances are introduced as local linearizing parameters. These are defined as the partial derivatives of the flux linkages with respect to the stator and rotor current components, thus capturing the local, possibly nonlinear, electromagnetic characteristics of the machine [117]. The same concept is applied to define the state-space equation of the rotor current. For EESMs, the stator differential inductance matrix is given by:

$$[l_{dq,f}] = \begin{bmatrix} l_{dd} & l_{dq} & l_{df} \\ l_{qd} & l_{qq} & l_{qf} \\ l_{fd} & l_{fq} & l_{ff} \end{bmatrix} = \begin{bmatrix} \frac{\partial \lambda_d(i_d, i_q, i_f)}{\partial i_d} & \frac{\partial \lambda_d(i_d, i_q, i_f)}{\partial i_q} & \frac{\partial \lambda_d(i_d, i_q, i_f)}{\partial i_f} \\ \frac{\partial \lambda_q(i_d, i_q, i_f)}{\partial i_d} & \frac{\partial \lambda_q(i_d, i_q, i_f)}{\partial i_q} & \frac{\partial \lambda_q(i_d, i_q, i_f)}{\partial i_f} \\ \frac{\partial \lambda_f(i_d, i_q, i_f)}{\partial i_d} & \frac{\partial \lambda_f(i_d, i_q, i_f)}{\partial i_q} & \frac{\partial \lambda_f(i_d, i_q, i_f)}{\partial i_f} \end{bmatrix} \quad (2.3)$$

In this formulation, the cross-coupling between the stator and rotor is fully accounted for. It should be noted that the dq, f inductances are not only functions

of the instantaneous d and q stator currents but also strongly depend on the rotor excitation current i_f . Accurate modeling and control require that this dependence be explicitly taken into account.

By substituting the differential inductance matrix (2.3) into (2.1), the stator voltage equation becomes:

$$\bar{v}_{dq,f} = \begin{bmatrix} R & 0 & 0 \\ 0 & R & 0 \\ 0 & 0 & R_f \end{bmatrix} \cdot \begin{bmatrix} i_d \\ i_q \\ i_f \end{bmatrix} + [l_{dq,f}] \cdot \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \\ i_f \end{bmatrix} + \omega \cdot \begin{bmatrix} -\lambda_q \\ \lambda_d \\ 0 \end{bmatrix} \quad (2.4)$$

Performing some manipulations on (2.4), the state-space equations of the stator dq currents in a generic machine's operating point are computed as follows:

$$\begin{cases} l_{eq,d} \cdot \frac{d}{dt} i_d = (v_d - R \cdot i_d + \omega \cdot \lambda_q) - k_{dq} \cdot (v_q - R \cdot i_q - \omega \lambda_d) + k_{df} \cdot (v_f - R_f \cdot i_f) \\ l_{eq,q} \cdot \frac{d}{dt} i_q = (v_q - R \cdot i_q - \omega \cdot \lambda_d) - k_{qd} \cdot (v_d - R \cdot i_d + \omega \lambda_q) - k_{qf} \cdot (v_f - R_f \cdot i_f) \\ l_{eq,f} \cdot \frac{d}{dt} i_f = (v_f - R_f \cdot i_f) + k_{fd} \cdot (v_d - R \cdot i_d + \omega \lambda_q) - k_{fq} \cdot (v_q - R \cdot i_q - \omega \cdot \lambda_d) \end{cases} \quad (2.5)$$

where the terms indicated with k denote the cross-coupling factors that quantify the influence of the interactions between the axes and are defined as:

$$\begin{cases} k_{dq} = \frac{l_{qd} \cdot l_{ff} - l_{qf} \cdot l_{fd}}{l_{qq} \cdot l_{ff} - l_{qf} \cdot l_{fq}}; & k_{df} = \frac{l_{qd} \cdot l_{fq} - l_{qq} \cdot l_{fd}}{l_{qq} \cdot l_{ff} - l_{qf} \cdot l_{fq}}; \\ k_{qd} = \frac{l_{dq} \cdot l_{ff} - l_{df} \cdot l_{fq}}{l_{dd} \cdot l_{ff} - l_{df} \cdot l_{fd}}; & k_{qf} = \frac{l_{dd} \cdot l_{fq} - l_{dq} \cdot l_{fd}}{l_{dd} \cdot l_{ff} - l_{df} \cdot l_{fd}}; \\ k_{fd} = \frac{l_{dq} \cdot l_{qf} - l_{df} \cdot l_{qq}}{l_{dd} \cdot l_{qq} - l_{dq} \cdot l_{qd}}; & k_{fq} = \frac{l_{dd} \cdot l_{qf} - l_{df} \cdot l_{qd}}{l_{dd} \cdot l_{qq} - l_{dq} \cdot l_{qd}}; \end{cases} \quad (2.6)$$

while $l_{eq,d}$, $l_{eq,q}$ and $l_{eq,f}$ are the equivalent differential inductances along the corresponding axes and computed as:

$$\begin{cases} l_{eq,d} = l_{dd} - l_{dq} \cdot k_{dq} + l_{df} \cdot k_{df} \\ l_{eq,q} = l_{qq} - l_{qd} \cdot k_{qd} - l_{qf} \cdot k_{qf} \\ l_{eq,f} = l_{ff} + l_{fd} \cdot k_{fd} - l_{fq} \cdot k_{fq} \end{cases} \quad (2.7)$$

The cross-coupling factors k_{dq} , k_{df} , k_{qd} and k_{qf} can frequently be neglected, as their effect on the system dynamics is minor and does not significantly impact the overall model accuracy in most practical cases. However, it is important to note that

applying this simplification does not decouple the stator and rotor systems. Indeed, their correlation is still considered, as shown in (2.3), the differential inductances are functions of the instantaneous values of the d and q stator currents and the rotor current i_f . This reasoning cannot be applied to the rotor equation, where the cross-coupling factors k_{fd} and k_{fq} have a crucial role in the definition of the dynamic behavior of the rotor current i_f . The cross-coupling terms act as an additive disturbance in the rotor control loop. It is also noted that for synchronous machines optimized for traction operation, the resistive voltage drops in the stator windings are often negligible due to their small relative magnitude.

Applying these simplifications and rearranging (2.5), the simplified state-space equations for the dq current dynamics can be written as:

$$\begin{cases} l_{dd} \cdot \frac{d}{dt} i_d \simeq v_d + \omega \lambda_q \\ l_{qq} \cdot \frac{d}{dt} i_q \simeq v_q - \omega \lambda_d \end{cases} \quad (2.8)$$

FVC operates by directly regulating the dq flux components. The state-space equations governing this type of control, derived from (2.1), are:

$$\begin{cases} \frac{d}{dt} \lambda_d \simeq v_d + \omega \lambda_q \\ \frac{d}{dt} \lambda_q \simeq v_q - \omega \lambda_d \end{cases} \quad (2.9)$$

These state-space representations enable a systematic analysis of the relative strengths and limitations of dq -frame-based torque controllers. While CVC strategies require regulators to be adaptively tuned to local differential inductances, FVC schemes are inherently more robust to parameter variations and often permit simpler controller design. Nonetheless, both approaches encounter limitations under voltage-constrained conditions typical of deep flux-weakening operation; in these regimes, the increasing back-EMF ($\omega \lambda_q$, $-\omega \lambda_d$) couples the dq axes, necessitating adaptive management of the available voltage margins in accordance with the machine operating point and the load angle δ [80].

2.1.2 Electromagnetic model in stator dqs coordinates

The stator voltage and electromagnetic torque expressions in the dqs stator flux-oriented reference frame are applicable to all synchronous machine topologies comprising the EESM, and are formalized as [109, 80, 63]:

$$\bar{v}_{dqs} = R \cdot \bar{i}_{dqs} + \frac{d}{dt} \begin{bmatrix} \lambda \\ 0 \end{bmatrix} + \lambda \cdot \begin{bmatrix} 0 \\ \omega + \frac{d\delta}{dt} \end{bmatrix} \quad (2.10)$$

$$T = \frac{3}{2} \cdot p \cdot \bar{\lambda} \times \bar{i} = \frac{3}{2} \cdot p \cdot \lambda \cdot i_{qs} \quad (2.11)$$

The rotor component is not reported as it is the same in every reference frame. Furthermore, the simplified model neglecting the cross-coupling factors k_{dq} , k_{df} , k_{qd} and k_{qf} is considered. Equation (2.10) illustrates that the amplitude of the stator flux linkage λ is regulated through the ds -axis voltage component, while the qs -axis voltage component directly influences the electromagnetic torque. Control strategies such as DFVC, DTC, and FPC inherently exploit these relationships, operating within the stator flux frame to facilitate intuitive decoupling of torque and flux regulation.

Furthermore, Equation (2.11) provides an alternative mechanism for torque regulation. Rather than directly adjusting the load angle δ , regulation of the qs -axis current i_{qs} offers a more straightforward and linearized framework for managing electromagnetic torque production. This current-oriented approach enables a direct proportional relationship between i_{qs} and torque, thereby mitigating complexities associated with load angle-based control.

To establish the direct correspondence between dqs voltage components and the specific control variables for DFVC (λ, i_{qs}), DTC (λ, T), and FPC (λ, δ) schemes, the associated state-space equations are systematically derived. It is noted that the flux amplitude equation results from the ds -axis voltage equation of (2.10):

$$\frac{d\lambda}{dt} = -R \cdot i_{ds} + v_{ds} \quad (2.12)$$

The state-space equation of the qs -axis current is instead computed by applying the rotational transformation on (2.7) [80] and using the load angle as the rotational

angle, leading to:

$$l_{qs} \frac{di_{qs}}{dt} \simeq -i_{ds} \cdot l_{qs} \cdot \frac{d}{dt} \delta + h_{ds} \cdot v_{ds} + (v_{qs} - \omega \lambda) \quad (2.13)$$

where the equivalent qs -axis differential inductance is defined as:

$$l_{qs} = \frac{2 \cdot l_{dd} \cdot l_{qq}}{l_{dd} + l_{qq} - (l_{qq} - l_{dd}) \cdot \cos(2\delta)} \quad (2.14)$$

with l_{qs} , l_{dd} , and l_{qq} denoting the relevant local differential inductances, while the voltage coefficient h_{ds} corresponds to:

$$h_{ds} = \frac{(l_{qq} - l_{dd}) \cdot \sin(2 \cdot \delta)}{(l_{dd} + l_{qq}) \cdot \cos(2\delta) - (l_{dd} + l_{qq})} \quad (2.15)$$

It should be considered that the load angle variation acts as an additive disturbance on the qs -axis control loop. Moreover, its action vanishes in steady-state conditions. Similar considerations are valid for the ds -axis voltage v_{ds} , whose value is negligible in practice as it must compensate only for the resistive voltage drops along the ds -axis according to (2.10). Therefore, the qs -axis voltage equation (2.13) is reformulated, resulting in:

$$l_{qs} \cdot \frac{di_{qs}}{dt} \simeq v_{qs} - \omega \lambda \quad (2.16)$$

To further characterize system dynamics, the time derivative of torque (2.11) leads to the state-space formulation:

$$\frac{d}{dt} T = \frac{3}{2} \cdot p \cdot \left(\frac{d\lambda}{dt} \cdot i_{qs} + \lambda \cdot \frac{di_{qs}}{dt} \right) \rightarrow \frac{2l_{qs}}{3p\lambda} \cdot \frac{dT}{dt} \simeq v_{qs} - \omega \lambda \quad (2.17)$$

These equations clarify the advantages of DFVC and DTC schemes according to the dqs model framework. Notably, one of the major benefits is the preservation of a linear torque regulation characteristic across the entire operating range, including the flux-weakening region. In DFVC, for any preassigned flux amplitude, the qs -axis current reference can be easily computed without the need for non-linear LUTs, a complication typically encountered in CVC implementations. This benefit holds for DTC as well.

Above base speed, the stator flux amplitude can be managed using model-based rules that remain robust to magnetic saturation, enabling straightforward implementation of flux-weakening. Moreover, the closed-loop dynamic response of the stator flux amplitude is invariant over the machine's full operating range (2.10).

A further advantage is that the inner regulation loops for flux amplitude and qs -axis feature independent voltage limits. Only a minimal portion of the inverter voltage margin is allocated to regulating the ds -axis (for resistance drop compensation), enabling the qs -axis regulator to utilize nearly the entire available voltage. This arrangement guarantees superior torque dynamic response, even under severe voltage constraints.

The FPC-based controller regulates torque via direct manipulation of the stator flux amplitude λ and the load angle δ . In this case, the qs -axis voltage governs the closed-loop behavior of δ , yielding the following:

$$\lambda \frac{d\delta}{dt} \simeq v_{qs} - \omega \lambda \quad (2.18)$$

The FPC methodology inherits the aforementioned advantages of DTC and DFVC, specifically the decoupled and robust control of torque and flux. However, unlike DTC and DFVC, FPC does not intrinsically maintain strict linearity in torque regulation, as magnetic saturation effects preclude a direct analytical relationship between torque and load angle δ . Therefore, practical FPC implementations require LUTs to map desired torque to a corresponding load angle reference, ensuring accurate response. However, FPC schemes are inherently more robust to parameter variations because they are independent of the machine's inductances. In summary, while the FPC strategy achieves dynamic performance robustness against parametric variations, its torque regulation accuracy remains dependent on the precision of the LUTs derived according to the machine's parameters.

2.1.3 Electromagnetic model in stationary $\alpha\beta$ coordinates

The stator voltage equations in the stationary $\alpha\beta$ reference frame are given by [149]:

$$\bar{v}_{\alpha\beta} = R \cdot \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \lambda_\alpha \\ \lambda_\beta \end{bmatrix} \quad (2.19)$$

This mathematical representation inherently decouples the α - and β -axis voltage components. With respect to the dq and dqs reference frames, where system variables in steady-state appear as dc quantities, variables in the $\alpha\beta$ reference frame remain ac-valued even under steady-state operating conditions. Consequently, the effective implementation of conventional PI controllers is precluded in this coordinate system.

Despite this limitation, the $\alpha\beta$ model structure is well-suited for DB control formulations. It is important to emphasize that DB control may also be effectively realized within the dq and dqs reference frames, subject to the aforementioned considerations regarding the coupling of controllable voltage margins between axes.

2.2 EESM cross-coupling

Effective torque control in EESMs involves several nontrivial challenges arising from the intrinsic electromagnetic coupling between the stator and rotor windings [128]. For a better understanding of this concept for the machine under consideration (MUC), the relationship between the d -axis flux λ_d component with the rotor current i_f and the relationship between the rotor flux λ_f component with the stator current i_d are reported in Fig. 2.2 and Fig. 2.3, respectively. One of the fundamental issues is the interaction between the stator d -axis current and the rotor excitation system, which produces cross-coupling effects that complicate the independent regulation of the dq current components [129]. This phenomenon is primarily attributed to armature reaction, whereby variations in stator currents induce voltage in the rotor field winding through the shared magnetic flux paths. Although the stator and rotor windings are galvanically isolated, their magnetic circuits are tightly coupled, resulting in bidirectional magnetic field exchange and dynamic interactions that must be explicitly considered in the control design, as shown in (2.5).

The impact of this coupling is particularly pronounced in traction-oriented EESM designs, where the exciter winding is intentionally configured with a much larger inductance than the stator windings. This design choice increases the machine's field-producing capability while improving robustness against current ripple; however, it simultaneously causes the excitation circuit to exhibit significantly slower electrical dynamics. Even when the exciter operates at its maximum allowable voltage, the rotor current time constant remains one to two orders of magnitude larger than that

of the stator currents. Consequently, the excitation system cannot respond rapidly to changes in the torque reference, leading to a torque-generation bottleneck [130].

Furthermore, high-efficiency torque control strategies typically rely on the continuous adaptation of the rotor field current to maintain the machine on optimal flux trajectories, such as the maximum torque per ampere (MTPA). Since these trajectories can vary rapidly with speed, load, and temperature, slow rotor current dynamics impose an additional limitation on the achievable torque bandwidth.

To partially mitigate these constraints, modern traction drives incorporate advanced thermal management and overcurrent-handling strategies that permit short-term stator current overloading during transient torque demands. Since the stator current loops possess much faster dynamics, the control system should rely on the stator's higher bandwidth for rapid torque modulation [131]. This approach, however, provides only a partial remedy, as it does not fundamentally resolve the coupling-induced limitations in the rotor excitation loop.

Therefore, for a control solution that is both robust and generalizable, it is essential to explicitly model and leverage the magnetic coupling between the stator and rotor subsystems. Advanced control approaches have been shown to exploit this coupling to improve dynamic torque performance and ensure stable operation across the entire operating range of the EESM [132].

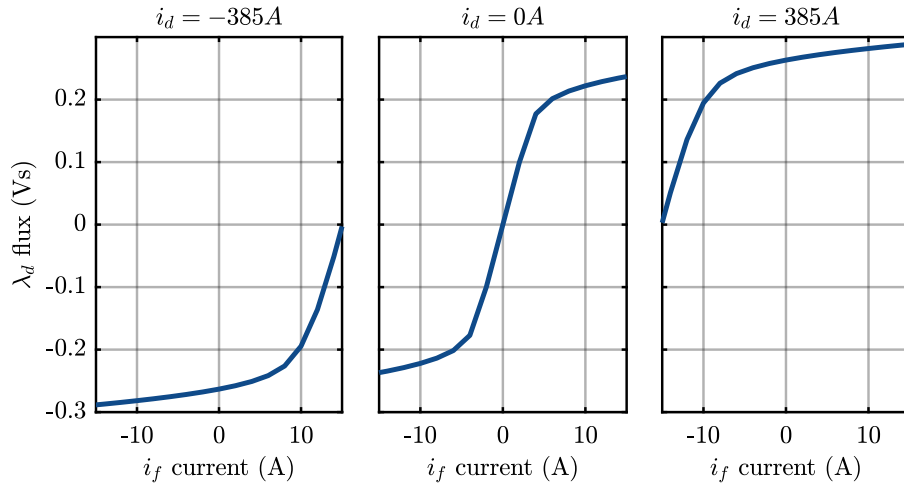


Fig. 2.2 Stator d -axis flux λ_d as function of the d -axis i_d and rotor i_f currents ($\lambda_d(i_d, 0, i_f)$).

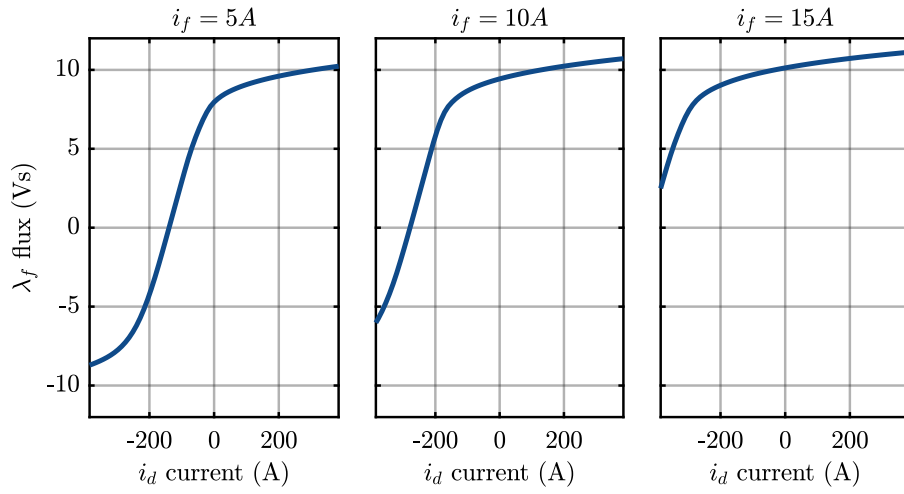


Fig. 2.3 Rotor flux λ_f as function of the d -axis i_d and rotor i_f currents ($\lambda_f(i_d, 0, i_f)$).

2.3 EESM under consideration

The MUC for the thesis work is the commercial EESM employed by the Renault ZOE R135 electrical vehicle (see Fig. 2.4). In particular, the MUC is a 3-phase, 4-pole motor of 100 kW with a base speed of 4000 rpm [150], whose main characteristic are shown in Table 2.1 while its geometrical data set is reported in Table 2.2 and in Fig. 2.5. Further details can be found in [151].

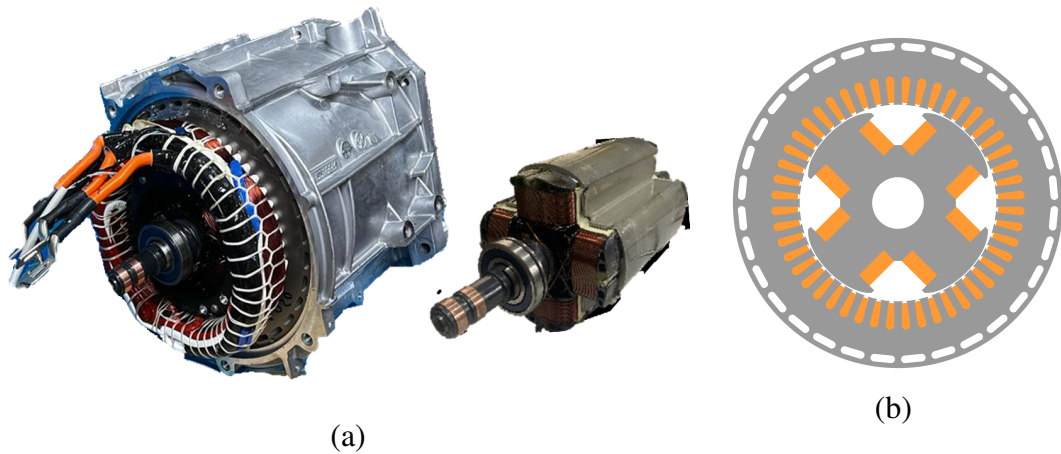


Fig. 2.4 The EESM case study: (a) prototype picture and (b) cross-section view [152].

Table 2.1 The EESM case study primary.

Machine characteristics	
Number of pole pairs	2
Winding type	stranded
Type of cooling	air
Peak power	100 kW
Rated power	51 kW
Peak torque	245 Nm
Base speed	4000 rpm
Maximum speed	11000 rpm
Peak phase current	385 Apk
DC-link voltage	350 V
Maximum excitation current	15 A
Stator resistance @20°C	9.8 mΩ
Rotor resistance @20°C	5.7 Ω

Table 2.2 Design data of the EESM under test.

Constructional details	
Number of stator slots	48
Slots per pole and per phase	4
Number of conductors in one slot	108
Number of external parallels	4
Number of internal parallels	12
Machine axial length	175.0 mm
Stator outer diameter	262.2 mm
Stator inner diameter	156.2 mm
Rotor outer diameter	154.6 mm
Rotor inner diameter	41.0 mm

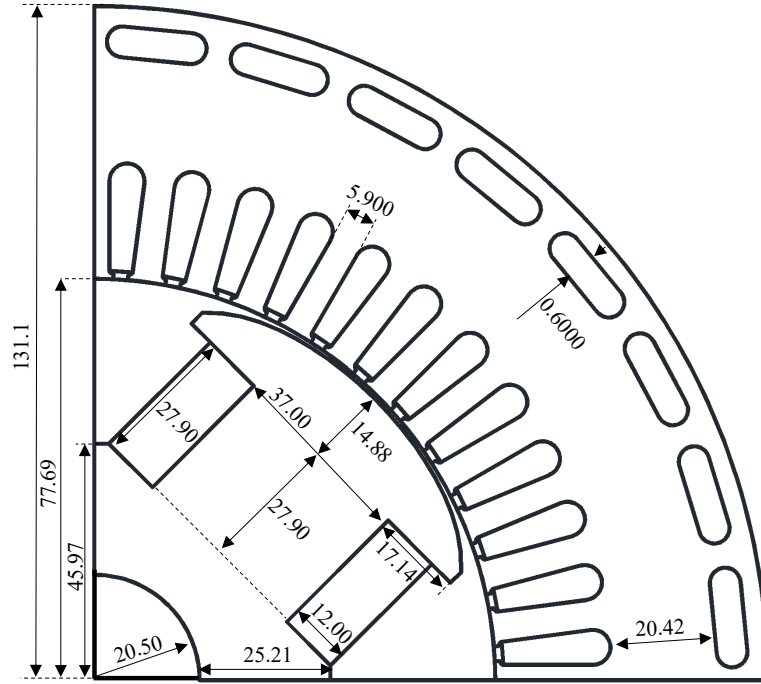


Fig. 2.5 Dimensions of the EESM under consideration (quotes in mm) [151].

The characterization of traction EESMs requires multidimensional flux linkage maps to accurately capture nonlinear saturation and magnetic cross-coupling. While PMSM modeling typically utilizes 2D relationships, the independently controlled rotor winding in EESMs introduces an additional degree of freedom, requiring generalized 3D surfaces (i_d, i_q, i_f) . Such high-fidelity mapping is essential for circuit-oriented formulations like the VBR model, enabling precise representation of electromagnetic transients and armature reaction within digital twim frameworks.

Furthermore, the information extracted from these flux maps is essential during machine design and also for advanced control implementations [74, 153, 154]. From a control perspective, high-performance torque and flux regulation over a wide operating range requires an accurate knowledge of the machine parameters depending on the operating point. In advanced TCAs, it is particularly important to account for saturation-induced parameter variations and dq -axis cross-coupling, as neglecting these effects degrades current control performance, reduces torque accuracy, and may lead to stability issues in field-weakening or high-load conditions.

2.4 Experimental identification of stator and rotor flux linkages maps

Flux linkage maps represent the standard paradigm for high-fidelity machine modeling [46, 155] and the systematic extraction of fundamental control trajectories [156, 157]. While PMSM characterization is sufficiently addressed through stator flux linkage definitions, EESM architectures need the concurrent identification of rotor flux linkage. This comprehensive characterization is critical for modeling the dynamic behavior of traction EESMs, specifically for quantifying the nonlinear magnetic cross-coupling between the stator and rotor windings [129, 132, 158], a requirement that is particularly important for the development of high-fidelity digital twins [158].

While stator flux identification can be effectively addressed using established protocols inherited from the PMSM literature [81, 159–162], rotor flux characterization remains largely unresolved. Despite renewed research interest in EESMs [163–166], existing methodologies predominantly rely on complex locked-rotor tests [163, 164]. Furthermore, these conventional approaches are often incompatible with traction-optimized designs; the confluence of low stator inductance and pronounced magnetic cross-coupling renders these methods impractical, as detailed below:

- Locked-rotor identification is strictly valid for the specific angular alignment of the machine. Consequently, capturing the spatial harmonics and slotting effects requires a discrete repetition of the identification procedure across multiple rotor positions to accurately map the saliency-dependent magnetic behavior.
- The significant magnetic cross-coupling triggered by rotor voltage steps induces severe transient perturbations in the stator current regulation loops. This interaction can compromise the stability margins of the closed-loop system, potentially leading to transient instability even when utilizing robustly tuned PI controllers.
- Under locked-rotor conditions, high-dynamic transients risk the structural failure of mechanical couplings and necessitate specialized, high-stiffness instrumentation to prevent hardware degradation.

To address these limitations, this work proposes a robust experimental identification procedure for the stator and rotor flux linkages of traction EESMs that eliminates the requirement for mechanical rotor locking. The proposed methodology formalizes the extension of consolidated PMSM stator flux identification procedures to the EESM topology. The novel rotor flux identification methodology uses current and torque transients specifically tailored to the operational and design constraints of traction-grade EESMs. In this approach, spatial harmonic effects resulting from the slot configuration are accurately incorporated by averaging the measurements over a full mechanical period. Furthermore, the work presents the first explicit experimental procedure for the high-precision evaluation of the rotor flux linkage at zero field current, ensuring the accuracy of the magnetic maps across the entire operating range.

2.4.1 Stator flux linkage identification procedure

Experimental identification of stator flux linkages is a mature topic within the context of PMSMs [81, 159–162]; these methodologies can be straightforwardly extended to EESMs with minor procedural modifications.

The identification protocol established in [81] is adapted here to accommodate the specific requirements of the EESM topology. Since the stator flux vector λ is dependent on the complete current state vector (i_d, i_q, i_f) , the characterization domain is expanded from a 2D plane to a 3D volume. Consequently, the standard (i_d, i_q) test grid is iterated across discrete rotor current levels, utilizing a recommended resolution of $5\% \cdot i_{f,max}$.

To mitigate excessive dynamic cross-coupling between the stator and rotor regulators, the rotor current is regulated to its setpoint prior to executing the three-pulse stator current evaluation. Furthermore, to ensure thermal stability and prevent insulation degradation, excitation is actively disabled during inter-test idle intervals. A single cycle of the described experimental sequence is illustrated in Fig. 2.6.

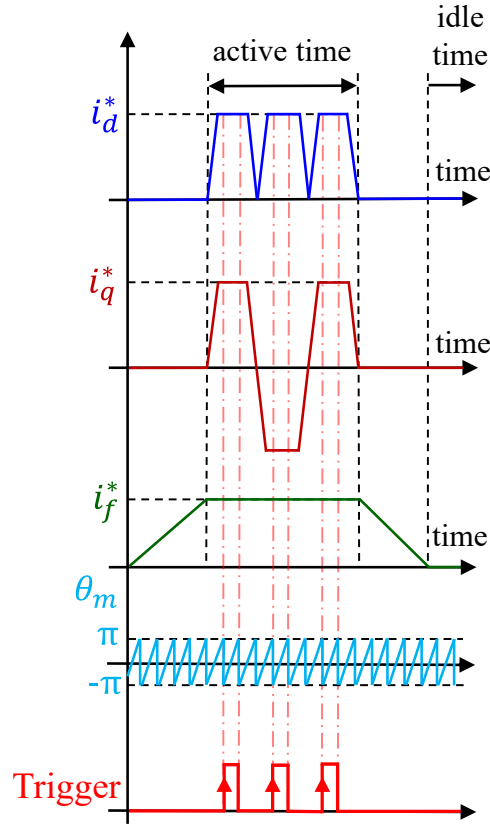


Fig. 2.6 Example of one cycle of the i_d, i_q, i_f pulse sequence.

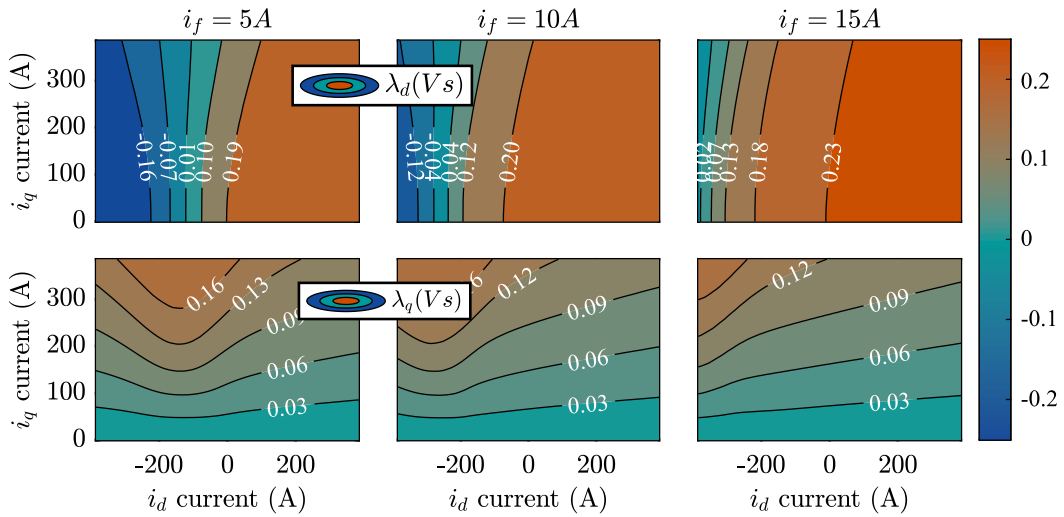


Fig. 2.7 Experimental stator current-to-flux maps for the EESM under consideration.

Once the 3D stator flux linkage volumes are identified in the (i_d, i_q, i_f) current domain (see Fig. 2.7), the machine's steady-state performance is systematically

characterized through fundamental electromagnetic relations. The electromagnetic torque is computed via (2.11) utilizing the identified flux vectors and specified current states. Consequently, the optimal control trajectories are extracted as surfaces within the current space, as illustrated in Fig. 2.8. In addition to the conventional MTPA and Maximum Torque Per Volt (MTPV) loci, the Maximum Efficiency Per Torque (MEPT) trajectory is identified. The MEPT locus is defined by the minimization of global Joule losses, comprising the simultaneous dissipation in both the stator and rotor field windings. The resulting machine performance characteristics are directly derived from these optimal paths. Fig. 2.9 presents the torque profiles corresponding to MTPA operation. Furthermore, the torque-speed and power-speed curves for a 350 V DC-link voltage are computed across discrete rotor excitation levels following the MEPT strategy, as reported in Fig. 2.10.

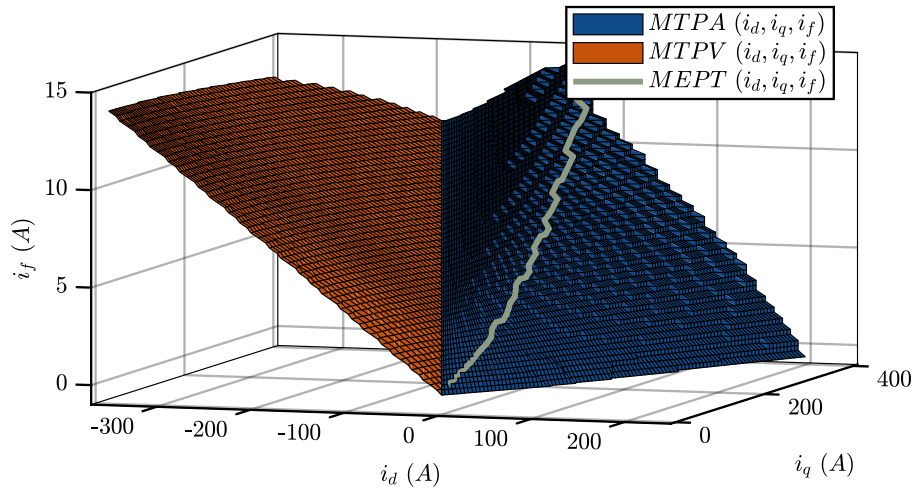


Fig. 2.8 Control trajectories (MTPA, MEPT, MTPV) for the EESM under consideration.

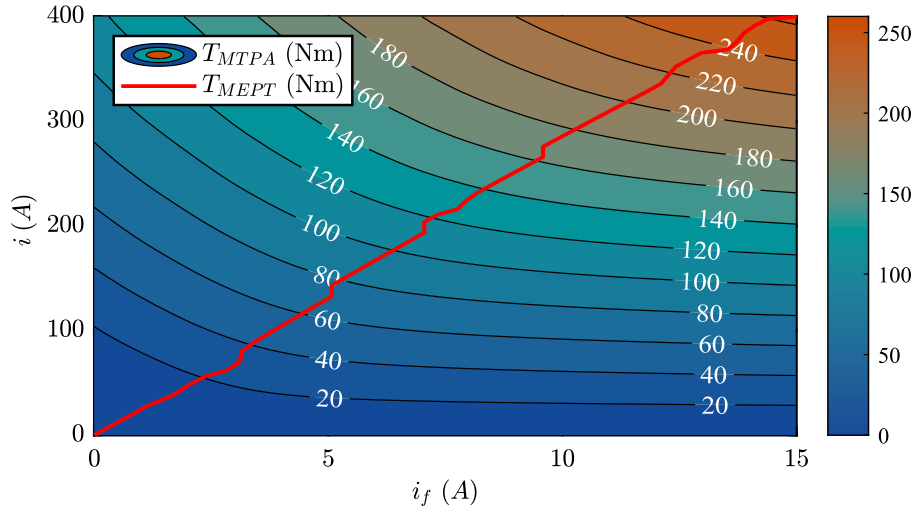


Fig. 2.9 Torque performance along the MTPA and MEPT control trajectories for the EESM under consideration.

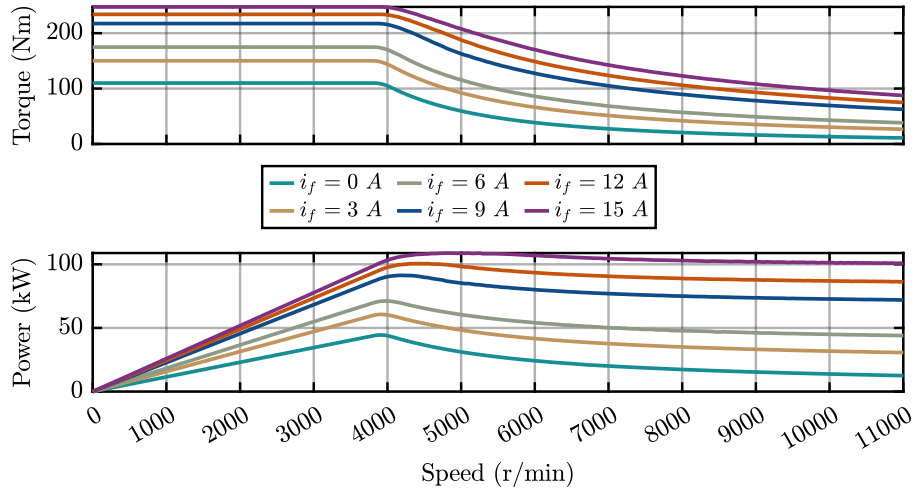


Fig. 2.10 Output torque vs speed curves (top) and output power vs speed curves (bottom) at 350 V dc-link voltage and different rotor current levels.

2.4.2 Rotor flux linkage identification procedure

While the experimental characterization of stator flux linkage maps is well-established through the extension of consolidated PMSM protocols, the specific challenges governing rotor flux linkage identification in traction EESMs remain largely unaddressed in the literature. Specifically, the use of square-wave excitation voltages [164] is

impractical for traction EESMs due to the combination of low stator inductances and pronounced magnetic cross-coupling.

To confirm this limitation, a simulation was performed for the MUC of a rotor-voltage step under locked-rotor conditions at the maximum torque point (see Fig. 2.11). In this analysis, safety limits on current magnitudes were disabled to observe the full unconstrained transient dynamics. As illustrated in Fig. 2.11, two critical issues are identified:

- **Control instability:** The intensive armature reaction triggered by the rotor voltage step induces severe transient disturbances in the stator current regulation loops. This interaction can cause a transient instability scenario, even with robustly tuned controllers.
- **Mechanical integrity:** Even assuming ideal current regulators with infinite bandwidth capable of maintaining stator current regulation stability, the resulting electromechanical transient remains extreme. For the MUC, the electromagnetic torque reaches its peak value of 245 Nm in 0.2 s, corresponding to a torque slew rate of 1225 Nm/s. Such rapid dynamics under locked-rotor conditions risk the structural failure of mechanical couplings and necessitate specialized, high-stiffness instrumentation to prevent hardware damage.

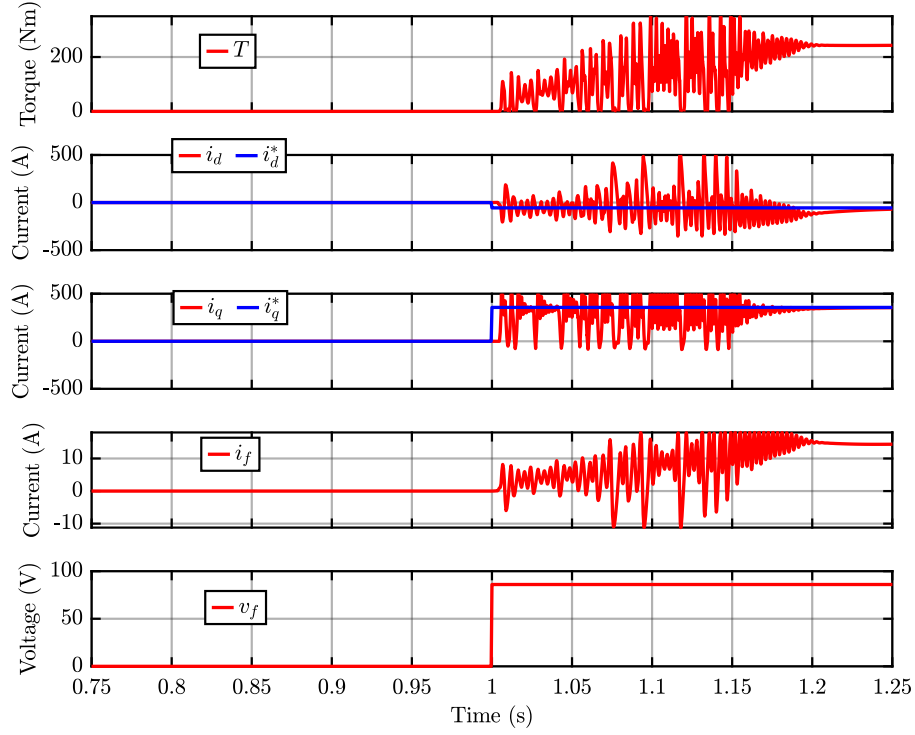


Fig. 2.11 Simulation of a rotor voltage step to obtain the maximum rotor current $i_f = 15$ A, with locked rotor.

With reference to this context, this work proposes an innovative, robust three-step methodology that accounts for practically executable current and torque transients according to traction EESMs designs.

The proposed identification procedure is schematically illustrated in Fig. 2.12 and comprises the following three stages:

1. **Rotor flux linkage characterization:** In this phase, the stator dq current loops are actively regulated to their references ($i_d = i_d^*, i_q = i_q^*$) while the rotor current i_f is controlled to follow a sinusoidal trajectory. The amplitude of this sinusoid corresponds to the maximum admissible rotor current $i_{f,max}$. This process is iterated across the entire (i_d, i_q) operating range to capture the full magnetic dependency.
2. **Error correction:** To mitigate inaccuracies deriving from measurement noise and parameter mismatches in the raw experimental data, a PSO algorithm is employed. This post-processing stage minimizes identification errors by

leveraging the redundancy inherent in the data acquired over a full sinusoidal period of the rotor current.

3. **Initial magnetic state characterization:** The rotor circuit is operated in an open-circuit condition while the stator dq currents are regulated to a predefined operating point. Subsequently, a rapid reversal of the d -axis current is executed to induce a measurable transient voltage in the rotor circuit, enabling the characterization of the initial magnetic state.

Each step is detailed in the following.

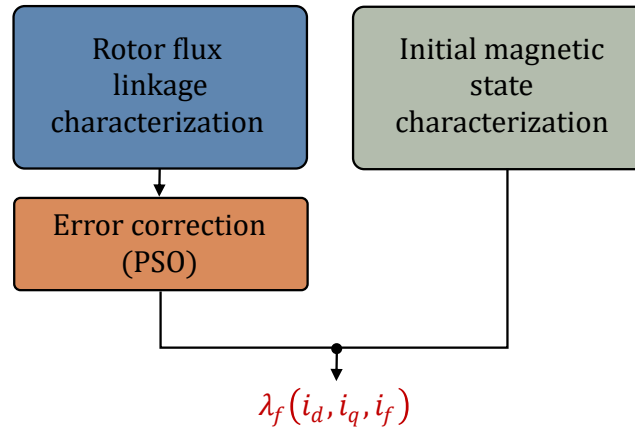


Fig. 2.12 Schematic overview of the proposed rotor flux identification procedure.

Rotor flux linkage characterization

The proposed methodology estimates the rotor flux linkage λ_f by evaluating the time-integral of the rotor back-electromotive force (back-EMF). This term is derived by decoupling the measured rotor terminal voltage v_f from the resistive drop, computed via the estimated rotor resistance $\tilde{R}_f$. Based on the fundamental voltage relationship in (2.20):

$$v_f = R_f \cdot i_f + \frac{d}{dt} \lambda_f, \quad (2.20)$$

the rotor flux linkage is formulated as:

$$\lambda_f(t) = \int_{t_0}^t (v_f(\tau) - \tilde{R}_f \cdot i_f(\tau)) d\tau \quad (2.21)$$

To facilitate experimental implementation, a dedicated test protocol is executed wherein the stator currents are regulated at constant setpoints ($i_d = i_d^*, i_q = i_q^*$) while the rotor current i_f follows a sinusoidal reference. The rotor current amplitude is set to the maximum machine rating $i_{f,max}$ at a frequency f_{if} , enabling a continuous characterization of the magnetic behavior within a single acquisition cycle. This procedure, depicted in Fig. 2.13, is swept across the entire (i_d, i_q) operating domain to capture the full cross-coupling effect.

It is important to note that the integration in (2.21) provides the flux linkage relative to an unknown offset $\lambda_{f,0}$. This constant corresponds to the initial magnetic state at the start of the integration period and must be determined to ensure the accuracy of the rotor flux map.

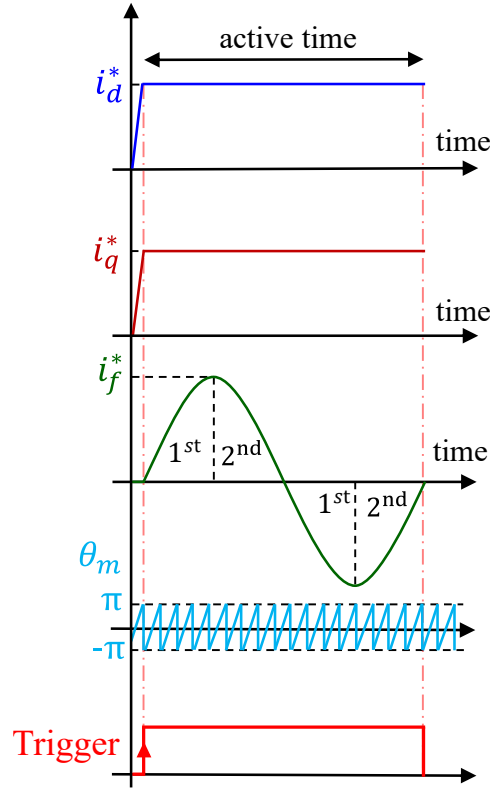


Fig. 2.13 Example of one cycle of the rotor flux linkage characterization test.

Error correction

To enhance the robustness of the identification process against experimental uncertainties and signal integration drift, the time-domain integration interval described

in (2.21) is repeated to span a full cycle of the sinusoidal rotor excitation. This temporal redundancy is used to generate multiple flux-linkage estimates for each discrete rotor-current level. From a theoretical standpoint, these estimates should coincide exactly, as they represent the same magnetic state of the machine.

Nevertheless, as depicted in Fig. 2.14, the experimental mapping of the estimated rotor flux λ_f against the current i_f exhibits an artificial hysteresis loop rather than the expected monotonic saturation characteristic. This divergent behavior is a direct consequence of cumulative errors embedded within the voltage integration. These estimation inaccuracies primarily originate from three critical sources:

- **Thermal parametric drift:** A mismatch between the nominal rotor resistance $\tilde{R}_f$ used in the calculation and the actual physical resistance, which varies dynamically with the winding temperature during the test sequence.
- **Power electronics non-linearities:** Inherent voltage distortions produced by the inverter stages, including conducting-state voltage drops, which cause the actual terminal voltage to deviate from the commanded values.
- **Dead-time effects:** Uncompensated inverter dead-time effects.

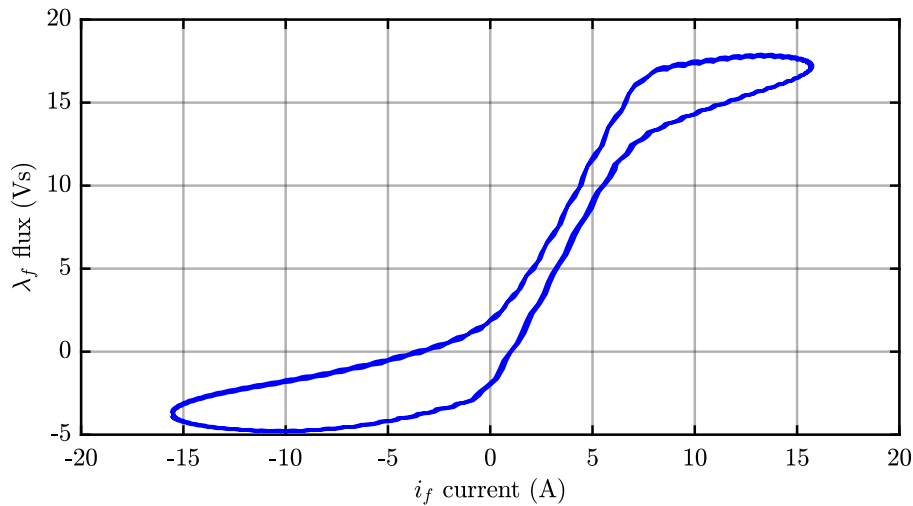


Fig. 2.14 Example result of rotor flux characterization test: $i_d^* = -100$ A, $i_q^* = 0$ A.

To mitigate these effects, the proposed framework does not rely on a static resistance value. Instead, it treats $\tilde{R}_f$ as an optimization variable within a PSO routine.

First, a reference mean trajectory is established by averaging the flux linkage values acquired over the full sinusoidal cycle. The objective function is then formulated based on the instantaneous deviation of each quarter-period's trajectory from this reference mean. In this framework, the estimated rotor resistance $\tilde{R}_f$ is treated as an optimization variable. Beyond its physical representation, $\tilde{R}_f$ is tuned to simultaneously compensate for winding resistance mismatch and the integrated influence of inverter voltage non-linearities and dead-time distortions. A PSO algorithm is utilized to iteratively refine $\tilde{R}_f$ until the fictitious hysteresis loops collapse into a single, physically consistent rotor flux linkage characteristic, as shown in Fig. 2.15.

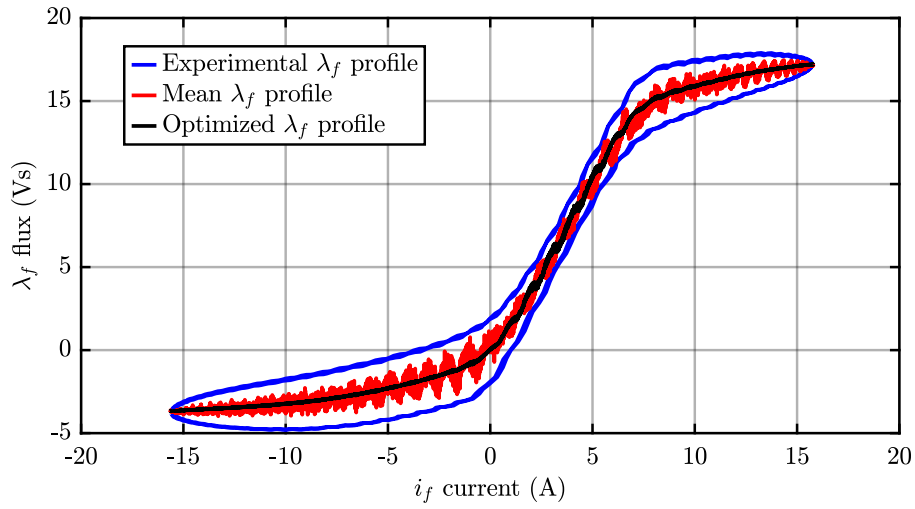


Fig. 2.15 Example result of the optimization of a rotor flux characterization test: $i_d^* = -100$ A, $i_q^* = 0$ A.

The distribution of the estimated rotor resistance $\tilde{R}_f$ across all tested operating points is illustrated in Fig. 2.16. The optimization yields a mean value of $\tilde{R}_{f,\text{mean}} = 6.2 \Omega$, with maximum and minimum estimates of $\tilde{R}_{f,\text{max}} = 6.3 \Omega$ and $\tilde{R}_{f,\text{min}} = 5.8 \Omega$, respectively. These results confirm the physical validity and high consistency of the proposed identification procedure. Specifically, the low standard deviation ($\sigma = 0.12 \Omega$, representing approximately 1.9% of the mean value) demonstrates the precision and robustness of the optimization algorithm. Furthermore, the systematic positive shift of the estimated values relative to the nominal resistance $R_f = 5.7 \Omega$ experimentally validates the expected additional voltage drops.

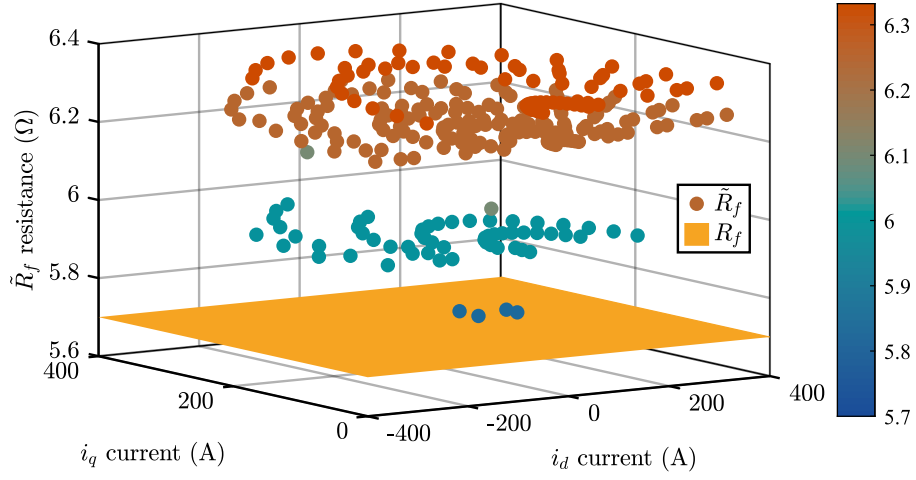


Fig. 2.16 Rotor resistance optimization values.

While this optimization may appear redundant compared to a simple arithmetic mean, it is needed by the asynchronous interaction between the fixed sampling frequency and the controlled frequency of the sinusoidal rotor current. Minor rotor current regulation deviations can result in a non-uniform sampling grid, where the discrete i_f points captured in the ascending quarter-period do not strictly coincide with those in the descending quarter-period. Consequently, the PSO-based refinement is essential to identify the correct rotor flux profile and prevent errors associated with a non-coincident sampling grid.

Initial magnetic state characterization

The third stage of the proposed procedure aims to compute the integration constant $\lambda_{f,0}$. Physically, this constant represents the rotor flux linkage component induced solely by the stator currents in the absence of rotor excitation, thus defining the initial magnetic state for the integration in (2.21).

Based on the physical boundary condition that the rotor flux linkage vanishes in the absence of any excitation ($i_d = i_q = i_f = 0 \Rightarrow \lambda_f = 0$), the integration constant $\lambda_{f,0}$ is also null in this specific scenario. However, a critical distinction must be made regarding its nature: while mathematically treated as a constant during the rotor current evolution in (2.21), $\lambda_{f,0}$ is physically dependent on the stator operating point. It effectively quantifies the magnetic cross-coupling, such that $\lambda_{f,0} = \lambda_f(i_d, i_q, 0)$.

To characterize this profile, a series of rotor open-circuit tests ($i_f = 0$) have been performed. In this condition, the resistive voltage drop in (2.20) is eliminated, and the terminal voltage v_f corresponds purely to the time derivative of the rotor flux linkage induced by stator current variations.

The identification protocol leverages the inherent magnetic symmetry of the EESM. The stator q -axis current is regulated at a reference i_q^* , while the d -axis current is regulated to $+i_d^*$ and subsequently reversed to $-i_d^*$ at a defined slew rate, as shown in Fig. 2.17. This slew rate is selected to balance the induction of a measurable rotor voltage against the requirement for a sufficient acquisition window. Given that the q -axis current contribution to the rotor flux linkage is negligible, only d -axis current reversals are performed to streamline the characterization. Under the assumption of magnetic symmetry, reversing the stator magnetizing current reverses the polarity of the induced rotor flux:

$$\lambda_f(i_d, i_q, 0) = -\lambda_f(-i_d, i_q, 0) \quad (2.22)$$

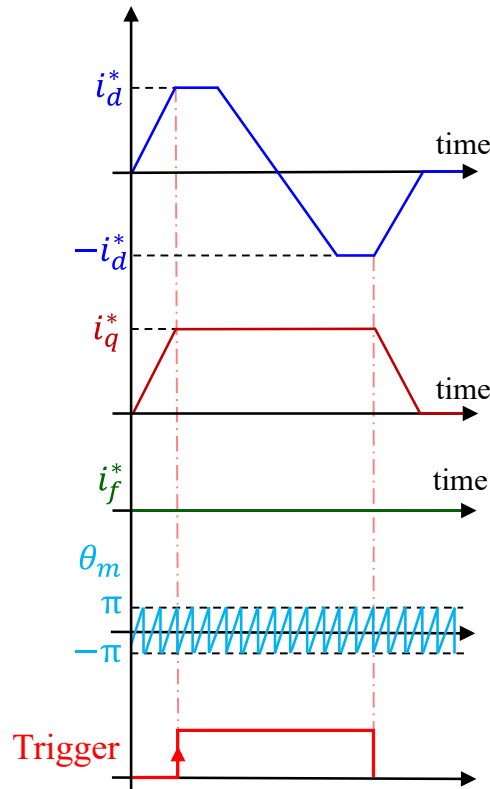


Fig. 2.17 Example of one cycle of the initial magnetic state identification test.

Integrating the measured rotor terminal voltage over the transition interval $[t_a, t_b]$ yields the total flux variation:

$$\int_{t_a}^{t_b} v_f d\tau = \int_{t_a}^{t_b} \frac{d\lambda_f}{dt} d\tau = \lambda_f(t_b) - \lambda_f(t_a) \quad (2.23)$$

where t_a and t_b denote the time instants immediately preceding and following the d -axis current reversal. Combining the symmetry condition (2.22) with (2.23), the rotor flux linkages at the boundaries are related by:

$$\lambda_f(t_b) = -\lambda_f(t_a) \quad (2.24)$$

Substituting (2.24) into (2.23) isolates the rotor flux linkage at the specific operating point (i_d^*, i_q^*) :

$$\int_{t_a}^{t_b} v_f d\tau = -2 \cdot \lambda_f(t_a) \Rightarrow \lambda_f(i_d^*, i_q^*) = -\frac{1}{2} \int_{t_a}^{t_b} v_f d\tau \quad (2.25)$$

The result of (2.25) defines the integration constant $\lambda_{f,0}(i_d = i_d^*, i_q = i_q^*)$ (see Fig. 2.18). This procedure is iteratively applied across the full stator dq -current range to construct the complete cross-coupling profile. Finally, the full rotor flux linkage map $\lambda_f(i_d, i_q, i_f)$ is derived by combining the corrected sinusoidal characterization results with the open-rotor test data.

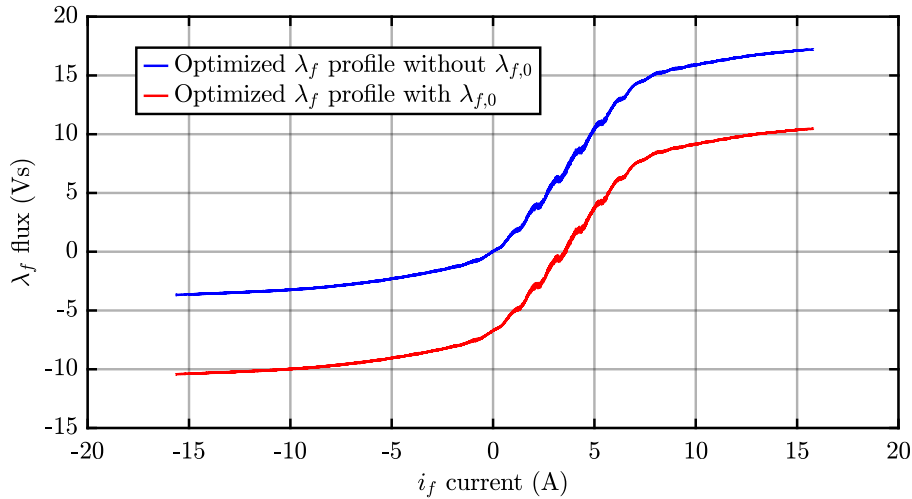


Fig. 2.18 Example result of optimized rotor flux profile with and without considering the initial magnetic state: $i_d^* = -100$ A, $i_q^* = 0$ A.

2.4.3 Experimental results

The proposed experimental identification procedure has been performed on the MUC previously described in Section 2.3, which is considered the machine under test (MUT) in this case.

Experimental setup

Experimental validation was performed on the test-bench setup shown in Fig. 2.19. The MUT has been mechanically coupled to a driving machine (DM) serving as the prime mover, and the mechanical torque is monitored via an in-line HBK T40B high-precision torque transducer.

The power electronics stage comprises a three-phase IGBT-based voltage source inverter (VSI) for the stator terminals and an IGBT H-bridge DC/DC converter for the rotor field winding, both supplied by a bidirectional 350 V DC source. The stator VSI operates at a switching frequency of 5 kHz with a 5 μ s dead time, while the rotor converter operates at 20 kHz with a 500 ns dead time.

Real-time control is executed on two dSPACE MicroLabBox platforms, with the sampling frequencies that are synchronized to the respective converter switching frequencies. Control algorithms were developed in MATLAB/Simulink and deployed via automatic C-code generation.

High-bandwidth instrumentation ensures signal acquisition fidelity. Phase currents are measured using LEM IT 605-S closed-loop Hall-effect transducers, while line-to-line voltages are captured through HBK GN610B high-voltage modules. All signals are logged by an HBK GEN7tA transient recorder at a sampling rate of 2 MS/s. Throughout the identification procedure, the DM maintained a constant mechanical speed of 240 r/min.

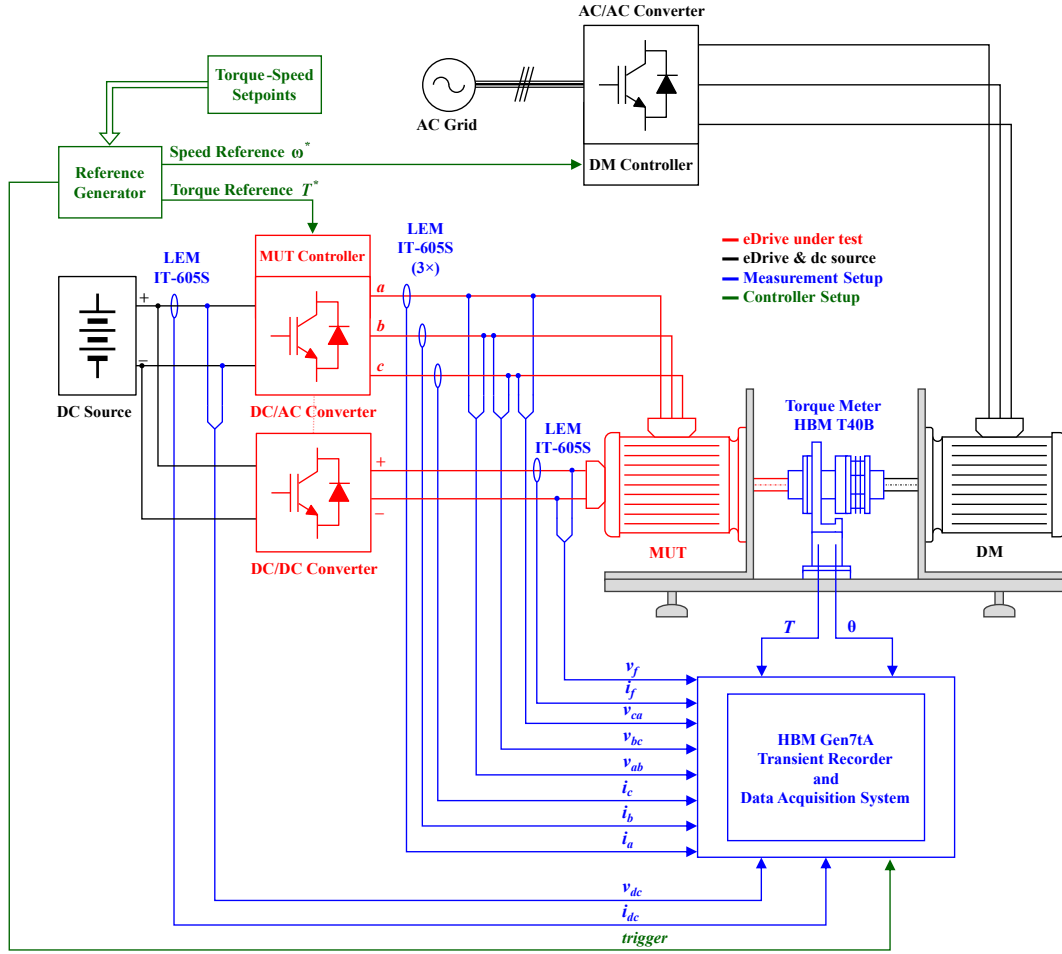


Fig. 2.19 Test and measurement setup.

Comparison with FEA results

The experimental characterization of the stator flux linkage followed the consolidated procedure described in Section 2.4.1, with results summarized in Figs. 2.7 - 2.10. Consequently, the following analysis focuses on the validation of the rotor flux linkage identification methodology, using FEA as the benchmark.

First, the initial magnetic state characterization $\lambda_f(i_d, i_q, i_f = 0)$ was compared with the FEA data. The procedure outlined in Section 2.4.2 was implemented using a current slew rate of 800 A/s. This value provides an optimal trade-off between inducing a sufficient rotor voltage and maintaining an adequate measurement window.

Fig. 2.20 illustrates the open-rotor test results at the maximum stator current ($i_d^* = 400$ A, $i_q^* = 0$ A). The experimental rotor flux λ_f was derived via real-time integration of the terminal voltage v_f . Although the measured v_f exhibits high-frequency noise due to inverter switching dynamics, phenomena not captured in FEA, the moving average of the measured voltage $\bar{v}_f$ aligns closely with the simulation.

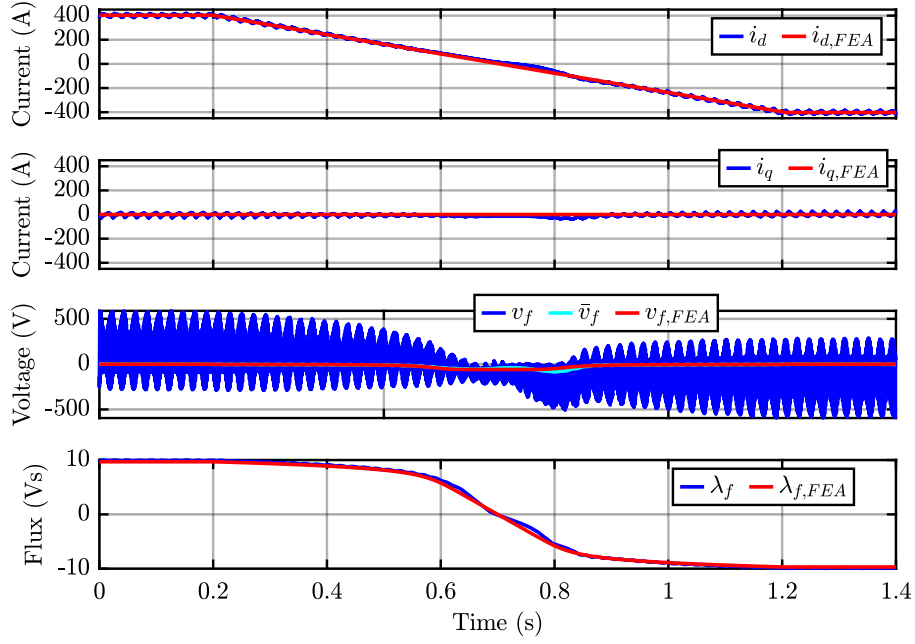


Fig. 2.20 Open rotor test with the maximum stator current $i_d^* = 400$ A: experimental vs FEA results.

The complete experimental profile of the initial magnetic state is compared with FEA results in Fig. 2.21. The results demonstrate nearly identical characteristics across the entire dq operating range of the MUT.

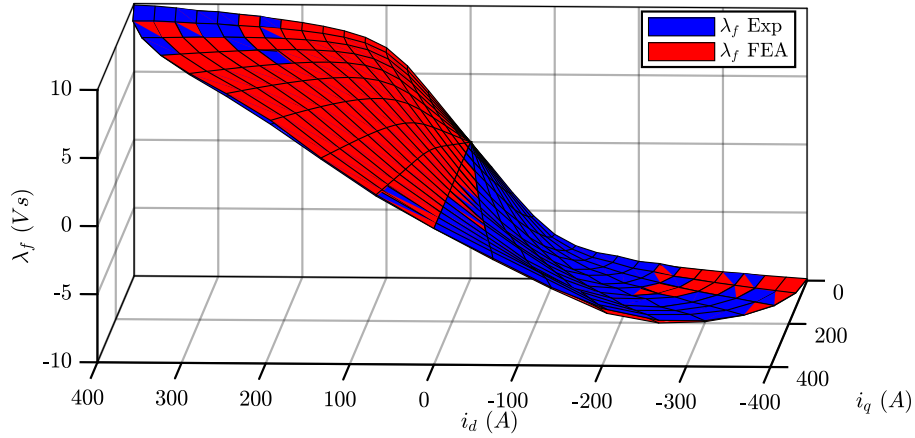


Fig. 2.21 Comparison between the rotor flux linkage at zero rotor current $\lambda_f(i_d, i_q, i_f = 0)$ derived with FEA and with the proposed experimental procedure.

Following the validation of the initial magnetic state, the total rotor flux linkage map was identified using the procedures detailed in Sections 2.4.2. The frequency of the sinusoidal rotor current reference i_f^* was set to 0.5 Hz to ensure stable tracking and minimize electromagnetic transients. Two sinusoidal periods were captured at each operating point to enhance robustness against measurement noise while preventing excessive machine heating (see Fig. 2.22). The dq current grid was discretized with a step of 2.5 A, corresponding to 0.6% of the maximum rated current.

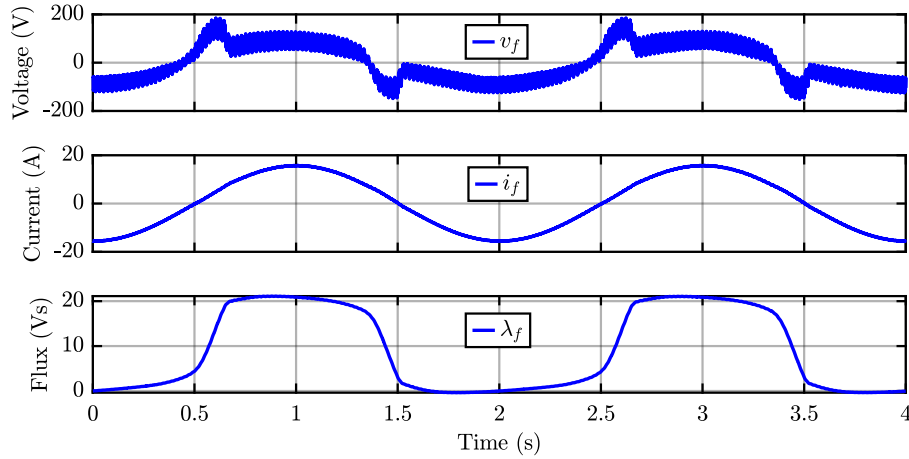


Fig. 2.22 Experimental signals of the rotor flux characterization test: $i_d^* = -100$ A, $i_q^* = 0$ A.

By integrating these results with the initial magnetic state data, the complete rotor flux linkage map $\lambda_f(i_d, i_q, i_f)$ was derived. As shown in Fig. 2.23, the experimentally identified surfaces for each rotor current level show a high degree of correlation with

the FEA benchmark. The absolute percentage error remains below 5%, validating the high fidelity of the proposed experimental identification approach.

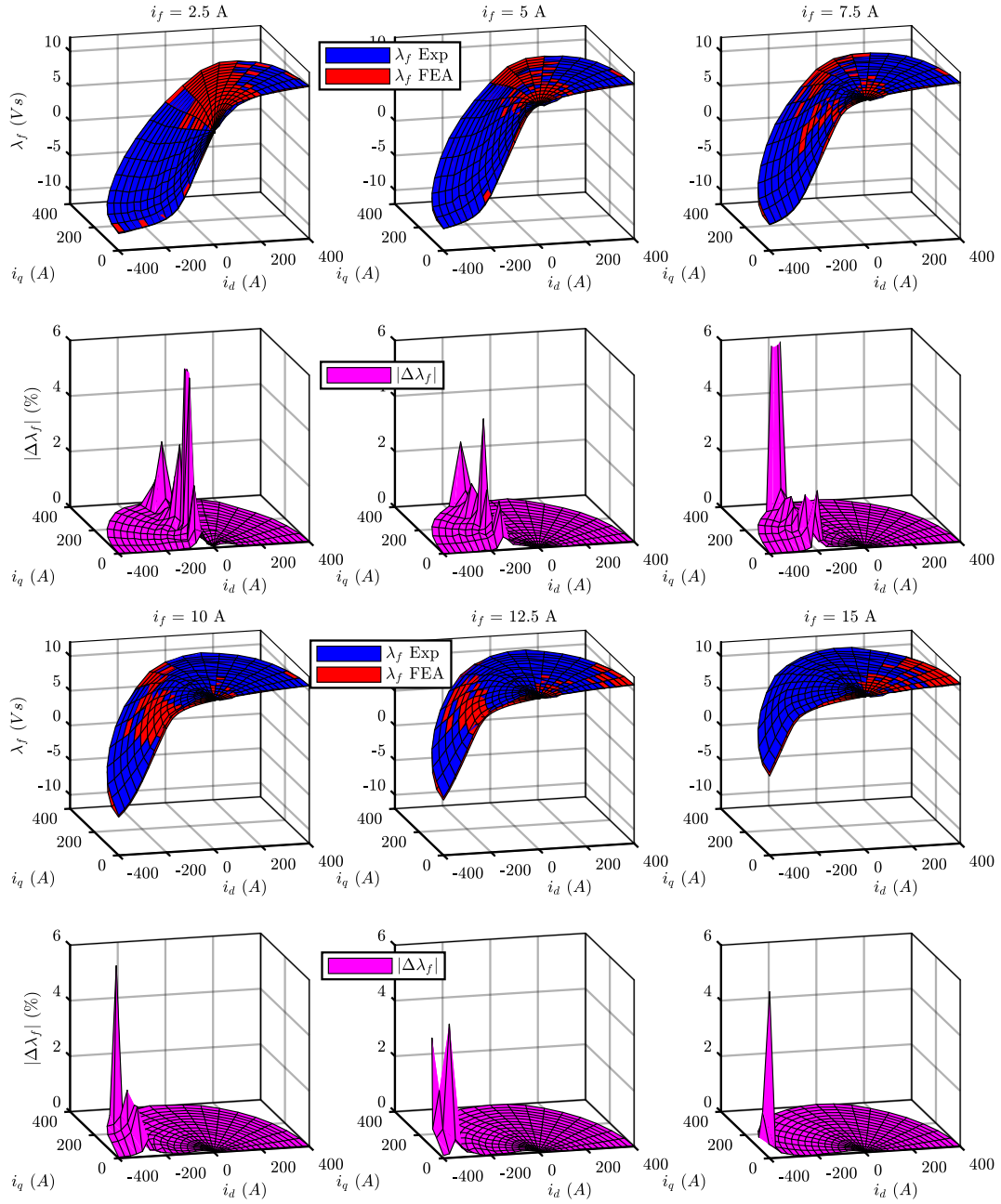


Fig. 2.23 Comparison between the rotor flux linkage $\lambda_f(i_d, i_q, i_f)$ derived with FEA and with the proposed experimental procedure.

Experimental validation

As demonstrated by the identification results in Fig. 2.23, the EESM exhibits highly nonlinear magnetic saturation with a multivariable dependency on the stator and rotor current vectors. These non-linear effects can be directly accounted for within the rotor control law by utilizing the identified flux linkage $\lambda_f(i_d, i_q, i_f)$ as the state variable [116]. While this flux-linkage-based formulation simplifies regulator tuning by linearizing the plant, the steady-state regulation accuracy is fundamentally contingent upon the fidelity of the current-to-flux map. To ensure high-precision tracking and provide an experimental metric for map verification, an error-correction mechanism utilizing rotor current feedback is implemented.

Accordingly, a DB rotor flux control scheme with an integral correction loop is employed, as shown in Fig. 2.24. The regulator modifies the rotor current reference i_f^* based on the deviation from the measured current i_f , yielding an adjusted reference $i_{f,adj}^*$ for the current-to-flux map. The integral gain is defined as the reciprocal of the target convergence time; specifically, it is set to 5 s^{-1} to ensure the elimination of residual current errors within 0.2 s.

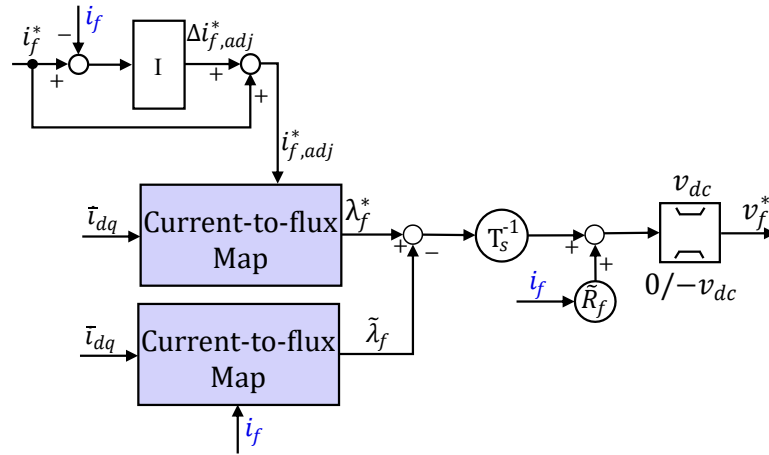


Fig. 2.24 Schematic overview of the rotor current DB control with the integral adjustment method.

The described rotor current regulation scheme is integrated into a torque control framework optimized to minimize global Joule losses in both the stator and rotor windings [167]. The rotor current reference is generated coherently with this efficiency criterion and the instantaneous torque demand.

Experimental validation was conducted at a mechanical speed of 4000 r/min, where the MUT performed a torque reversal from -150 Nm to 150 Nm at a slew rate of 50 Nm/s. The results in Fig. 2.25 show the comparison between the nominal reference i_f^* and the adjusted reference $i_{f,adj}^*$. The correction term $\Delta i_{f,adj}^*$ remained consistently below 3% throughout the transient. This minimal correction requirement experimentally confirms the high fidelity of the proposed rotor flux linkage identification procedure.

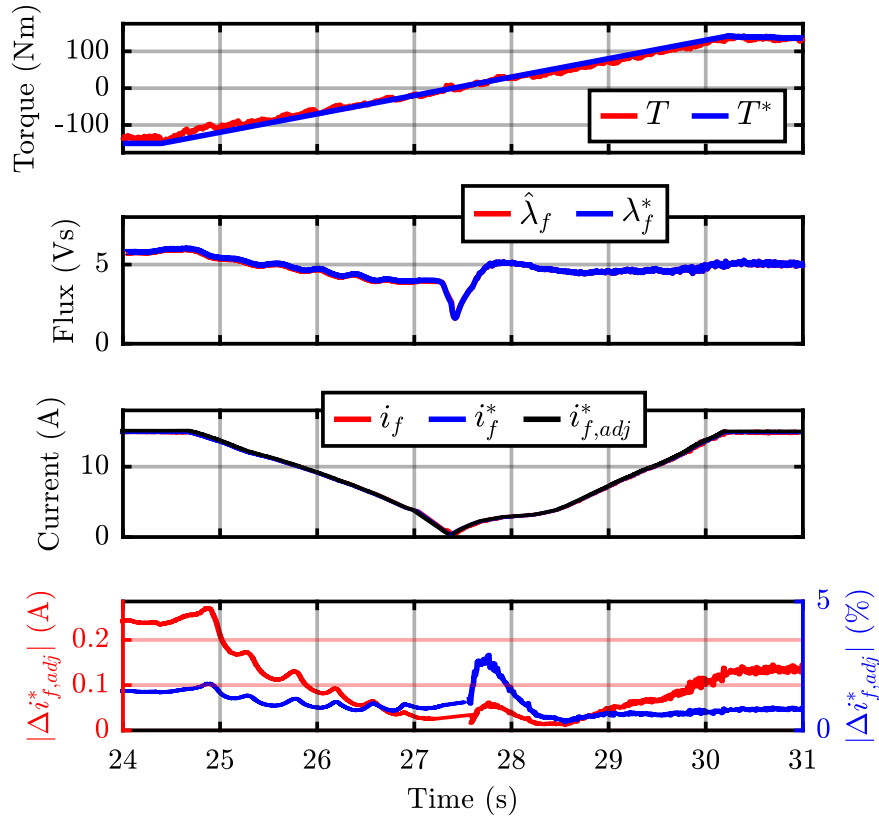


Fig. 2.25 Experimental results for torque reversal at 4000 r/min, 50 Nm/s.

To rigorously validate the fidelity of the identified rotor flux map across the complete torque-speed range of the MUT, the rotational speed was swept in 500 r/min increments from zero up to 10000 r/min. At each discrete speed level, ten equally spaced torque references were applied, ranging from zero to a maximum of 150 Nm. As illustrated in Fig. 2.26, the rotor current regulation error Δi_f is consistently negligible, while the rotor current adjustment $\Delta i_{f,adj}^*$ remains below 3%. This

marginal compensation requirement provides experimental evidence of the high accuracy achieved by the proposed rotor flux linkage identification procedure.

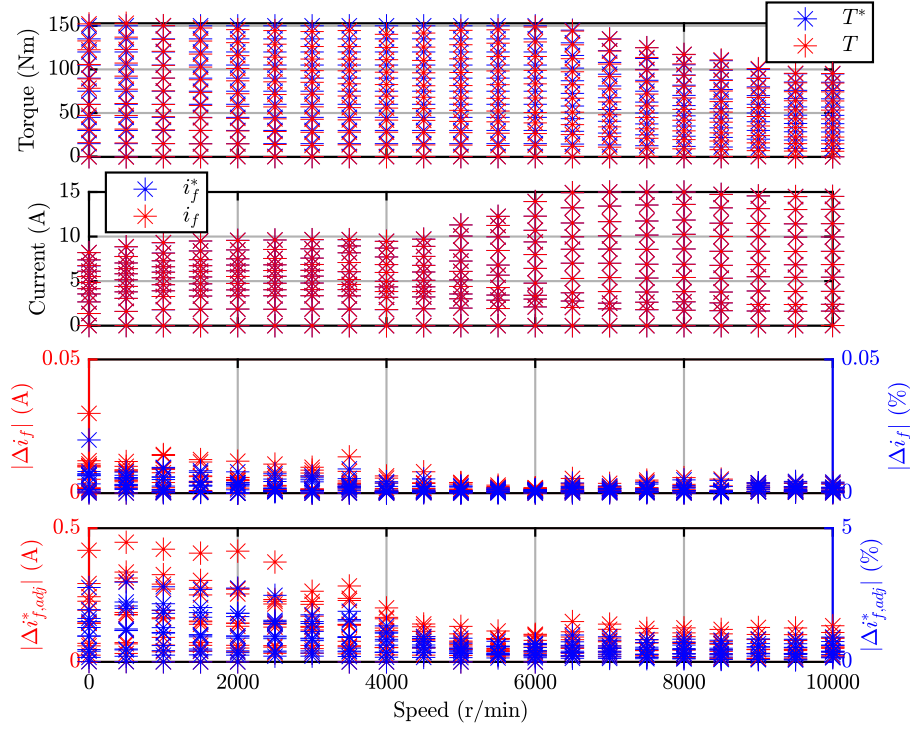


Fig. 2.26 Experimental results of steady-state torque and rotor current regulation accuracy.

2.5 MUC inductance maps

From the identified stator and rotor flux maps, it is possible to retrieve the differential inductance maps representing the current-to-inductance relationships. According to (2.3), the inductance maps are retrieved by applying the discrete differentiation of direct flux maps as:

$$l_{xy} = \frac{\Delta \lambda_x}{\Delta i_y} \quad (2.26)$$

where x and y can be any of the reference axis. Such elaborated maps are shown in Figs. 2.27 - 2.35.

As previously described in Section 2.1, the differential inductances are necessary to define the state-space equation linearizing the current dynamics around a given

operating point. According to that definition, the inductance maps can be employed in TCAs that require the regulation of a current (CVC, DFVC) to adaptively tune the current regulators coherently with the working point.

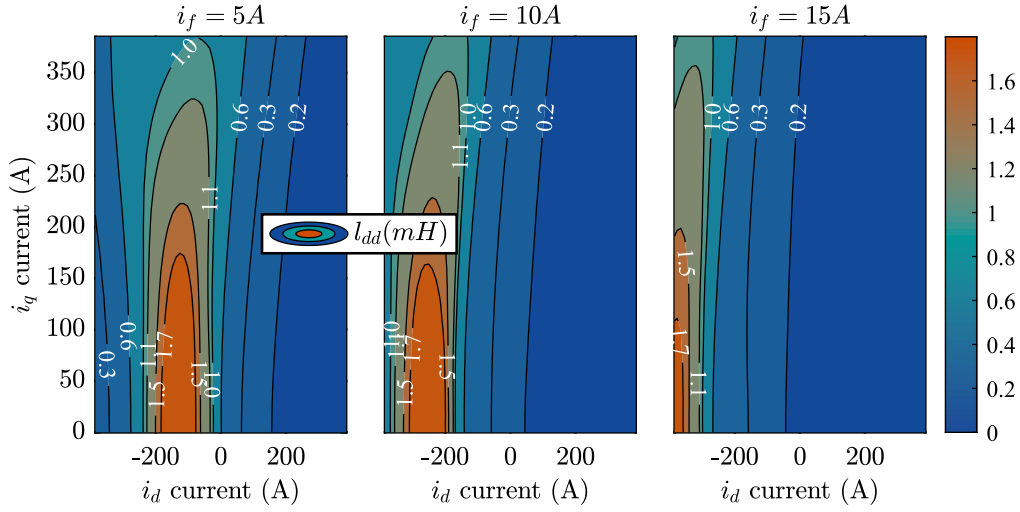


Fig. 2.27 Current-to-inductance l_{dd} map for the EESM under consideration.

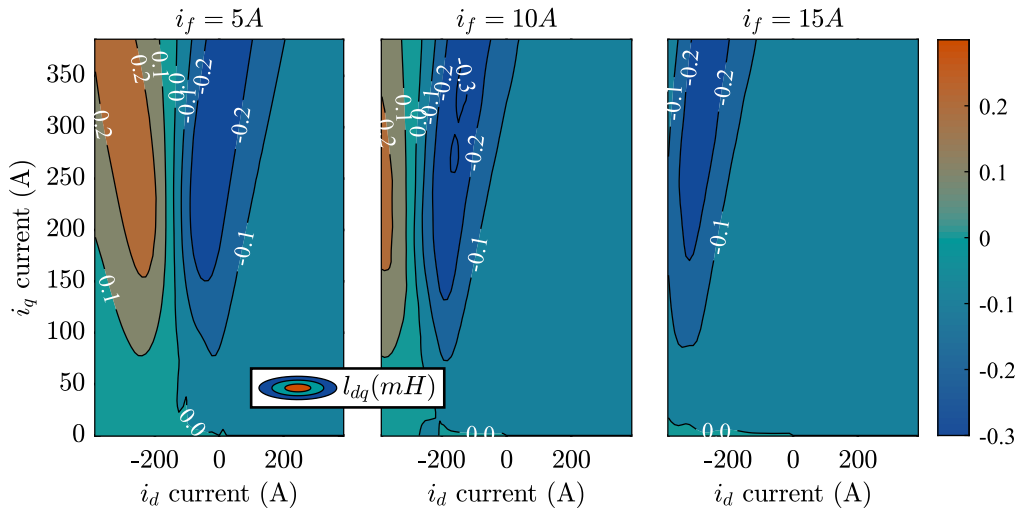


Fig. 2.28 Current-to-inductance l_{dq} map for the EESM under consideration.

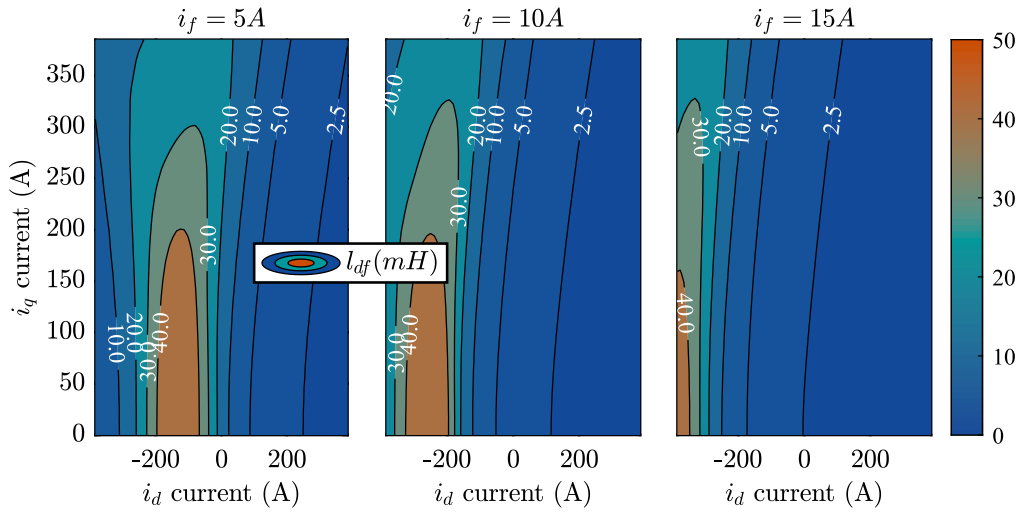


Fig. 2.29 Current-to-inductance l_{df} map for the EESM under consideration.

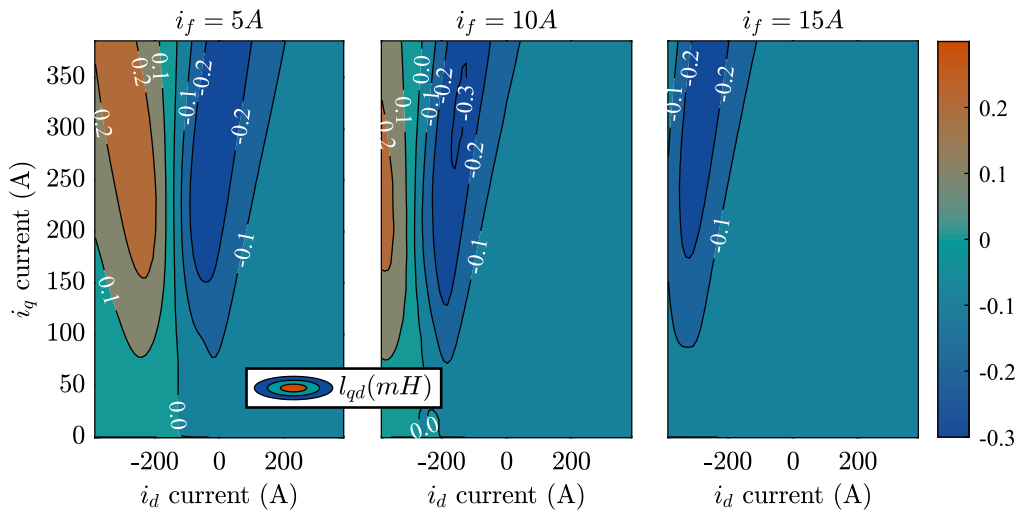


Fig. 2.30 Current-to-inductance l_{qd} map for the EESM under consideration.

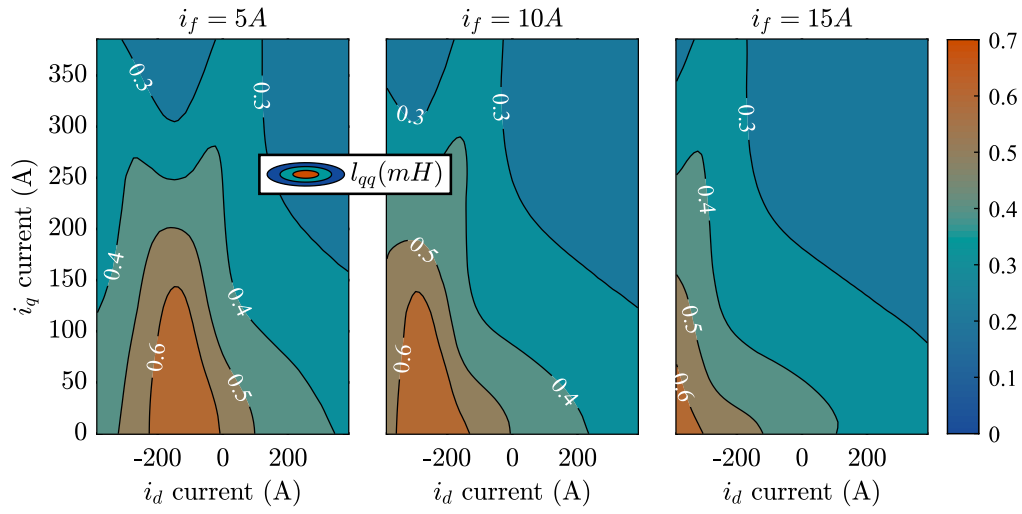


Fig. 2.31 Current-to-inductance l_{qq} map for the EESM under consideration.

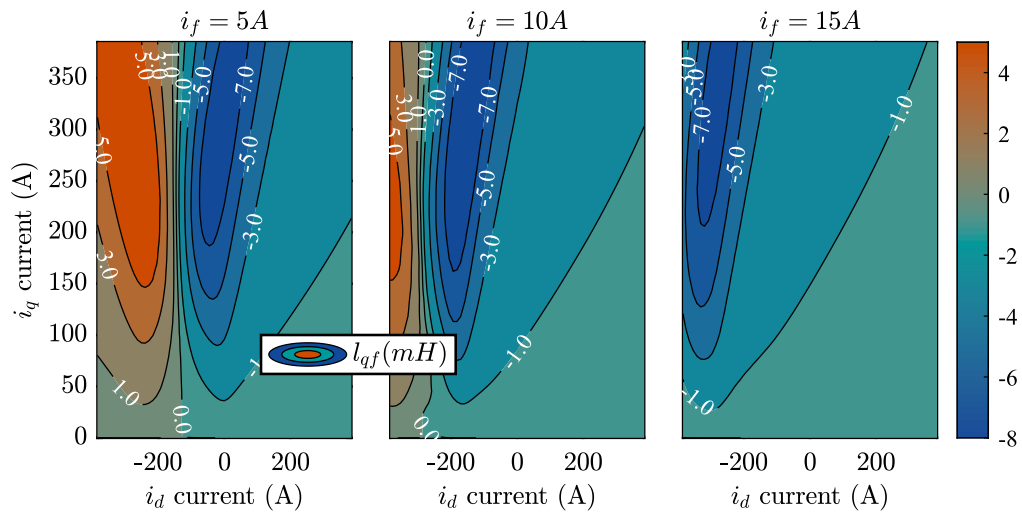


Fig. 2.32 Current-to-inductance l_{qf} map for the EESM under consideration.

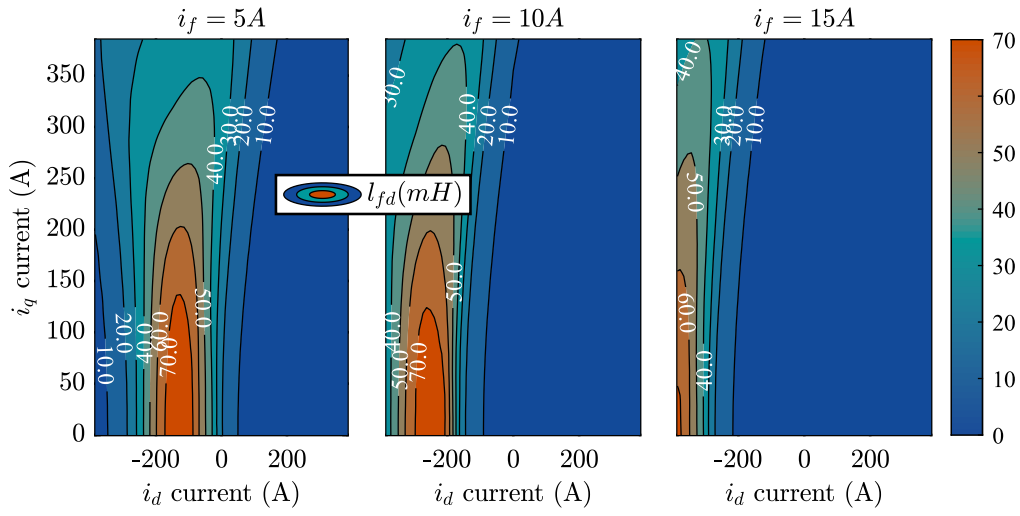


Fig. 2.33 Current-to-inductance l_{fd} map for the EESM under consideration.

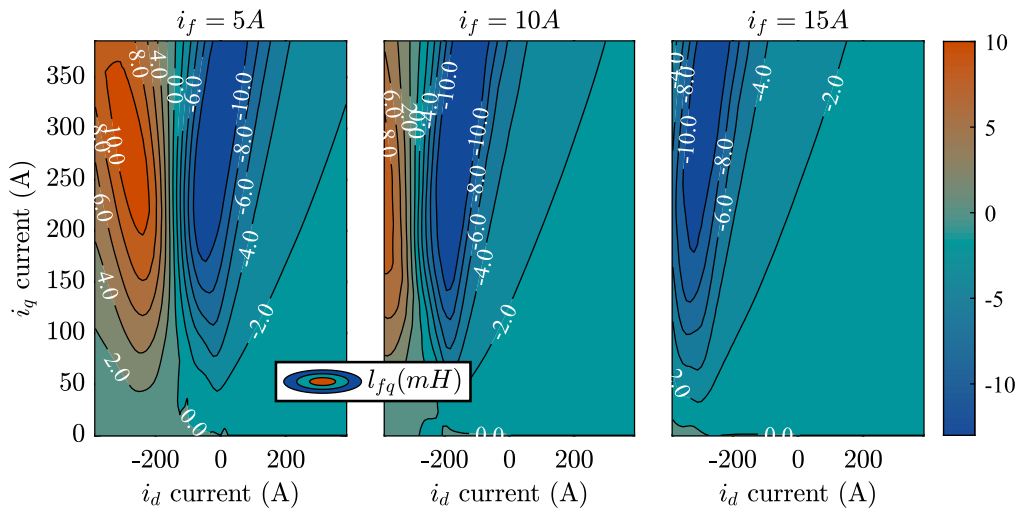


Fig. 2.34 Current-to-inductance l_{fq} map for the EESM under consideration.

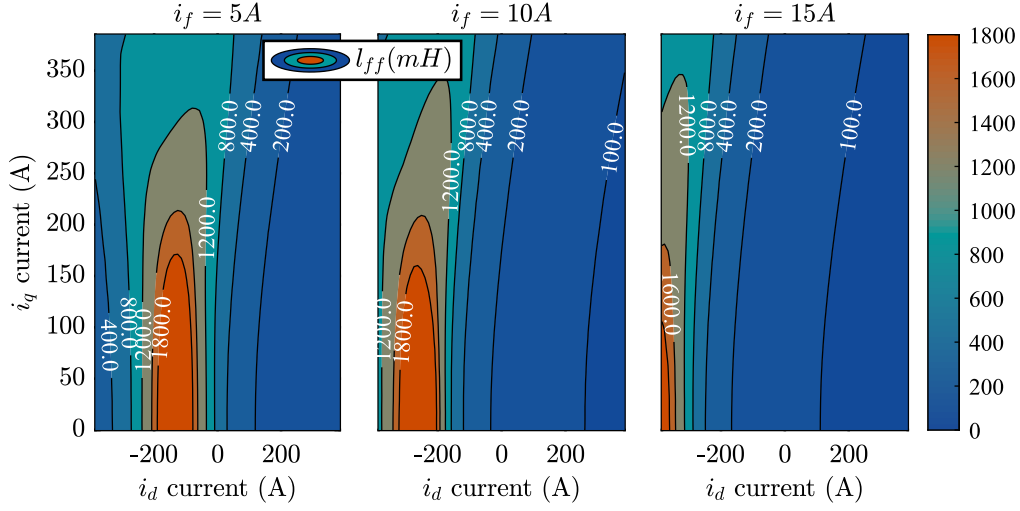


Fig. 2.35 Current-to-inductance l_{ff} map for the EESM under consideration.

In Fig. 2.36, the equivalent differential inductance on the rotor axis accounting for the cross-coupling terms $l_{eq,f}$ is shown. Comparing it with the rotor self-inductance l_{ff} (see Fig. 2.35), it is evident that they have considerably different values. Therefore, it is confirmed that no approximation can be done for the rotor current state-space equation, as previously stated in Chapter 2.1.1.

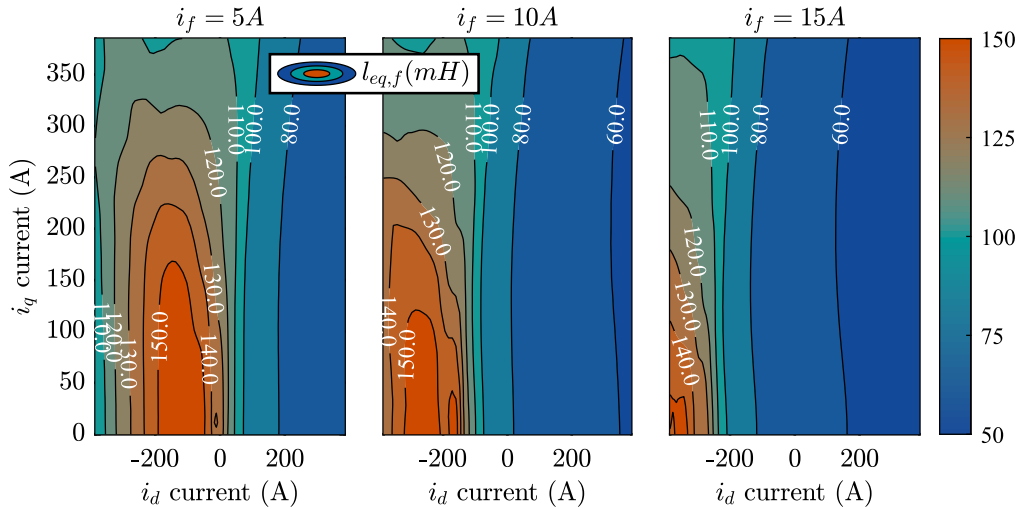


Fig. 2.36 Current-to-inductance $l_{eq,f}$ map for the EESM under consideration.

If a TCA in the dqs reference frame has to be applied, it is also helpful to retrieve the l_{qs} inductance (2.14) (see Fig. 2.37).

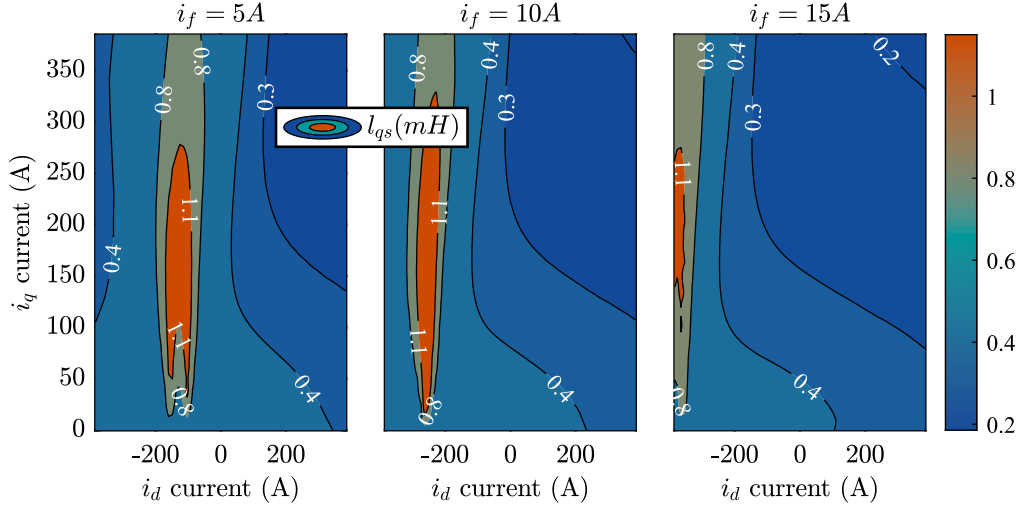


Fig. 2.37 Current-to-inductance l_{qs} map for the EESM under consideration.

2.6 Summary

This chapter establishes the fundamental mathematical modeling formulation for EESMs, serving as the theoretical basis for the advanced control strategies developed in this thesis. The analysis classifies the machine dynamics across three distinct coordinate systems: the stationary $\alpha\beta$ frame, the rotor synchronous dq frame, and the stator flux-oriented dqs frame, evaluating the suitability of each for specific control topologies such as CVC, FVC, DFVC, DTC, FPC.

A rigorous derivation of the electromagnetic voltage and torque equations is presented, with particular emphasis on the state-space formulations required to define the control laws. The modeling approach explicitly incorporates the machine's nonlinear magnetic behavior through the definition of the full differential inductance matrix $[l_{dq,f}]$. This formulation captures the significant cross-coupling effects between the stator d -axis and the rotor excitation winding, identifying the distinct time scale disparity between stator and rotor dynamics as a critical constraint in traction drive performance.

The theoretical models are applied to a commercial Renault ZOE R135 traction motor. The chapter details the experimental identification of high-fidelity multidimensional direct (current-to-flux) magnetic maps. For the stator flux identification, a standard technique reported in the literature for PMSMs is adapted to accommodate

the unique EESM topology. However, the concurrent identification of the rotor flux linkage is indispensable for modeling the dynamic behavior of traction EESMs. The proposed methodology addresses the critical limitations of conventional locked-rotor and square-wave excitation tests, which are unsuitable for traction designs characterized by low stator inductance and significant armature reaction.

The proposed approach utilizes a three-stage rotor flux identification sequence:

1. Rotor flux linkages are derived via time-domain integration of the rotor back-EMF during continuous sinusoidal excitation.
2. A PSO routine is implemented to eliminate fictitious hysteresis loops caused by thermal resistance drift and inverter non-linearities.
3. The integration constant $\lambda_{f,0}$ is identified by leveraging magnetic symmetry during open-rotor stator current reversals.

Validation against FEA benchmarks confirms the high fidelity of the 3D maps across the entire (i_d, i_q, i_f) operating range. Furthermore, experimental results with a DB rotor flux control scheme demonstrate that the identified maps enable high-precision tracking, with a residual rotor current error of less than 3%.

Finally, these maps are utilized to extract variable differential inductance profiles (e.g., $l_{dd}, l_{qq}, l_{eq,f}$), providing the necessary parameter set for the adaptive tuning of regulators and the implementation of robust, saturation-aware torque controllers.

Chapter 3

Digital Twin of EESMs

The accurate characterization of nonlinear magnetic behavior in modern AC machines relies on multidimensional direct and inverse flux maps to capture saturation and cross-saturation effects [45, 81, 46, 47]. For traction EESMs, the complexity increases due to the additional degree of freedom provided by the rotor field current i_f , requiring flux linkages to be defined as functions of the full current vector (i_d, i_q, i_f) [48–51]. To support digital twin approaches, the Voltage Behind Reactance (VBR) formulation offers a modular modeling approach, partitioning the machine into stator and rotor subsystems [52]. While historical EESM models for industrial generation relied on simplified one-dimensional saturation curves [59–61], high-performance traction applications require VBR-based models that incorporate full 3D flux-linkage surfaces to account for complex armature reaction and cross-coupling phenomena [62–64].

To address these issues, state-of-the-art modeling strategies employ high-resolution multidimensional flux maps derived from FEA [168, 169]. In the automotive field, FEA-based characterization is the standard approach for reproducing the machine's magnetic response under wide-ranging operating conditions. These maps, implemented via fitting functions or LUTs, replace simplified saturation curves and enable the digital twin to accurately simulate nonlinear behavior during realistic drive cycles.

A remaining challenge in developing robust digital twins is accurately identifying stator parameters, particularly the stator leakage inductance. Conventional analytical or empirical methods often fail to provide reliable stator leakage inductance estimates under deep saturation conditions [170, 171]. This issue represents an open

problem in both academic research and industry, highlighting the need for improved identification algorithms and validation procedures.

3.1 Literature analysis

Several advanced modeling paradigms for synchronous machines have been proposed in the literature to meet power system simulation requirements [172–175]. This section presents a comparative analysis of established EESM modeling strategies, highlighting their respective advantages and limitations as summarized in Table 3.1.

A popular approach approximates the EESM stator magnetic model using a structure analogous to that of PMSMs [176–178]. In this formulation, the constant permanent magnet flux linkage is replaced by a variable flux component directly dependent on the rotor field current i_f [179]. Accordingly, the magnetic behavior is expressed as a three-dimensional mapping:

$$f_\lambda : (i_d, i_q, i_f) \rightarrow (\lambda_d, \lambda_q) \quad (3.1)$$

In this configuration, the rotor field current i_f is treated as an external input variable rather than a state variable governed by the machine's electromechanical dynamics. Consequently, this model neglects the dynamic behavior of the rotor excitation circuit, as illustrated in Figure 3.1.

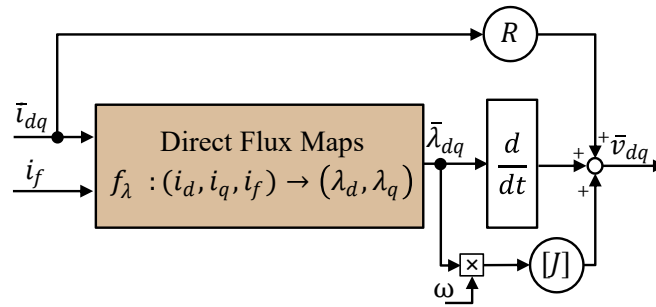


Fig. 3.1 dq model with imposed rotor current i_f for EESM.

For PMSMs, non-linear magnetic effects have been rigorously captured through inverse flux mapping techniques represented as:

$$f_\lambda^{-1} : (\lambda_d, \lambda_q) \rightarrow (i_d, i_q) \quad (3.2)$$

This procedure has been extended to EESMs [51], although with increased methodological complexity due to the additional rotor field variable. A three-dimensional mapping describes the resulting system:

$$f_{\lambda}^{-1} : (\lambda_d, \lambda_q, \lambda_f) \rightarrow (i_d, i_q, i_f) \quad (3.3)$$

Comprehensive inversion algorithms for generic n -dimensional mappings have been reported in [48, 51]. When rotor flux maps are integrated, both stator and rotor behavior are jointly represented, yielding a holistic magnetic characterization (see Figure 3.2). The principal drawback is that such modeling approaches are inherently not suitable for translation into circuit-oriented forms, thereby impeding direct network integration.

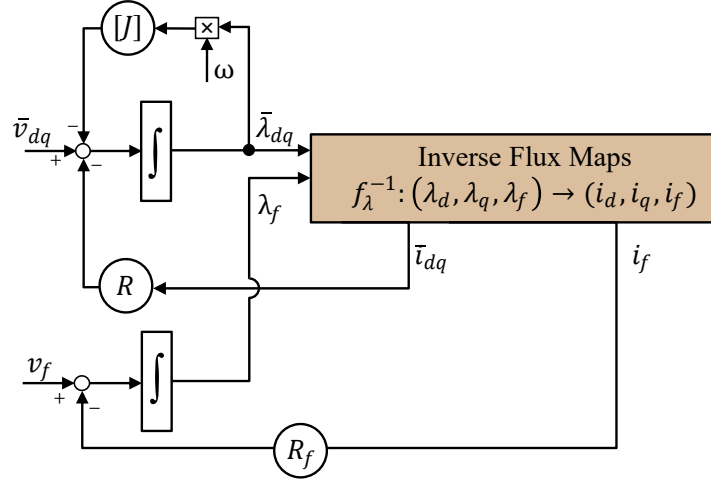


Fig. 3.2 Inverse flux maps-based model for EESM.

An alternative and widely adopted methodology leverages the Voltage-Behind-Reactance (VBR) formulation. This approach is associated with enhanced computational efficiency, flexibility in stator winding configuration, and seamless compatibility with both circuit-based and differential equation-based simulation platforms. The classical implementation of VBR for EESMs, frequently referred to as Stator-VBR, represents only the stator subsystem in a circuitual form, thus facilitating direct connection to network models at the stator terminals [56, 52, 54, 55, 61, 175, 180]. The rotor subsystem, conversely, is interfaced via a controlled current source (see Figure 3.3). Despite its utility, this methodology is subject to restrictive assumptions, precluding scenarios such as an open-circuit condition on the rotor side.

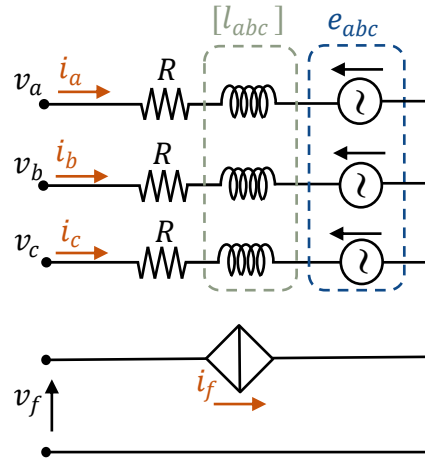


Fig. 3.3 Stator-VBR model for EESM.

A further extension of the VBR paradigm, denoted in [57] as the Full-VBR model, represents both stator and rotor winding systems with circuitual fidelity. In this formalism, the complete machine is characterized through a 4×4 inductance matrix, whose off-diagonal elements are explicitly dependent on the rotor position and facilitate accurate coupling between stator and rotor field windings (see Fig. 3.4) [57, 174, 181].

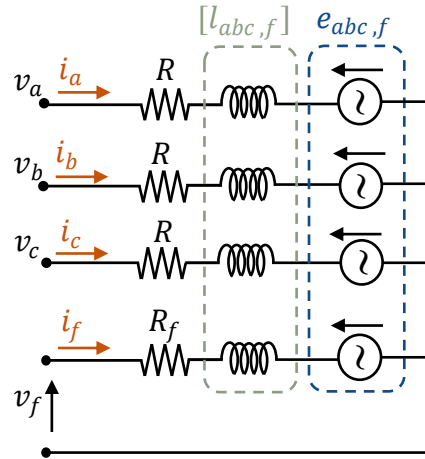


Fig. 3.4 Full-VBR model for EESM.

While numerous works have contributed variants of Stator- and Full-VBR representations, incorporating both constant and variable parameter methodologies [52, 54, 55, 61, 175, 180, 174, 181], these approaches are conventionally limited by their use of magnetic saturation curves to approximate magnetic nonlinearities [182, 58]. This simplification undermines their accuracy for modeling strong

cross-coupling effects between stator and rotor windings under conditions of deep magnetic saturation. Consequently, the existing models fall short of providing a rigorous account of the underlying magnetic relationships, particularly for EESMs designed for traction applications that feature a highly saturated magnetic behavior. The emerging prominence of EESMs as viable alternatives to PMSMs in traction applications has accentuated the necessity for increasingly accurate modeling of stator-rotor electromagnetic coupling.

Table 3.1 Comparison of EESM model formulations.

■ = Direct, □ = Indirect | ● = Accurate, . = Inaccurate

Model Formulation	Circuit Interface		Cross-coupling Accuracy
	Stator	Rotor	
Flux maps inversion [48, 51]	□	□	●
Stator-VBR [56, 52, 54, 55]	■	□	.
Full-VBR [57, 174, 181]	■	■	.
Proposed VBR	■	■	●

3.2 Proposed VBR model

Synthesizing the strengths of established literature formulations, this work proposes an advanced VBR model characterized by an interfacing circuit governed by a full 4×4 incremental inductance matrix, encompassing both stator and rotor quantities. These inductance parameters exhibit functional dependence on the current state vector (i_d, i_q, i_f) and, consequently, on the rotor electrical position, thereby explicitly capturing the magnetic cross-coupling between the stator and the rotor windings. As a result, the parameters of the interfacing circuit are inherently state-dependent and time-variant. This formulation is enabled by a novel inductance computation methodology derived directly from multidimensional flux maps, which is described in the next subsections.

3.2.1 Stator VBR formulation with flux maps

For the stator VBR formulation, some manipulation to the machine equations is required. Starting from (2.1), the flux derivative is written as:

$$\frac{d\bar{\lambda}_{dq}}{dt} = \frac{\partial \bar{\lambda}_{dq}}{\partial \bar{i}_{dq}} \cdot \frac{d\bar{i}_{dq}}{dt} \quad (3.4)$$

where the partial derivative of flux with respect to current is defined as incremental inductance, thus generating the incremental inductance matrix:

$$l_{dd} = \frac{\partial \lambda_d}{\partial i_d} \quad l_{dq} = \frac{\partial \lambda_d}{\partial i_q} \quad l_{qd} = \frac{\partial \lambda_q}{\partial i_d} \quad l_{qq} = \frac{\partial \lambda_q}{\partial i_q} \quad (3.5)$$

$$[l_{dq}] = \begin{bmatrix} l_{dd} & l_{dq} \\ l_{qd} & l_{qq} \end{bmatrix} \quad (3.6)$$

It is pointed out that $[l_{dq}]$ is the matrix of the incremental inductances in the dq frame, while l_{dq} is one of the elements of the matrix $[l_{dq}]$ representing the cross-coupling incremental inductance between stator axis d and q . These incremental inductances vary with the working point according to the currents. Therefore, (2.1) is rewritten as:

$$\bar{v}_{dq} = R \cdot \bar{i}_{dq} + [l_{dq}] \cdot \frac{d\bar{i}_{dq}}{dt} + [J] \cdot \omega \cdot \bar{\lambda}_{dq} \quad (3.7)$$

where $[J] = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$ is the complex operator in matrix form.

The rotational transform (3.8) is introduced as:

$$A(\theta) = \begin{bmatrix} \cos(\theta) & \sin(\theta) \\ -\sin(\theta) & \cos(\theta) \end{bmatrix} \quad (3.8)$$

By applying (3.8) to (3.7), the stationary $\alpha\beta$ voltage equations using the rotor electric position θ (see Fig. 2.1) are computed as:

$$\begin{aligned} A(-\theta) \cdot \bar{v}_{dq} &= A(-\theta) \cdot R \cdot \bar{i}_{dq} + A(-\theta) \cdot [l_{dq}] \cdot \frac{d\bar{i}_{dq}}{dt} + A(-\theta) \cdot [J] \cdot \omega \cdot \bar{\lambda}_{dq} \Rightarrow \\ \bar{v}_{\alpha\beta} &= R \cdot \bar{i}_{\alpha\beta} + A(-\theta) \cdot [l_{dq}] \cdot \frac{d\bar{i}_{dq}}{dt} + A(-\theta) \cdot [J] \cdot \omega \cdot \bar{\lambda}_{dq} \end{aligned} \quad (3.9)$$

Considering only the second term, characterized by the current derivative, (3.9) is transformed in:

$$\begin{aligned} A(-\theta) \cdot [l_{dq}] \cdot \frac{d}{dt} \bar{i}_{dq} &= A(-\theta) \cdot [l_{dq}] \cdot \frac{d}{dt} A(\theta) \cdot \bar{i}_{\alpha\beta} = \\ A(-\theta) \cdot [l_{dq}] \cdot A(\theta) \cdot \frac{d}{dt} \bar{i}_{\alpha\beta} &+ A(-\theta) \cdot [l_{dq}] \cdot \frac{d}{dt} A(\theta) \cdot \bar{i}_{\alpha\beta} \end{aligned} \quad (3.10)$$

The first part is equivalent to the matrix of incremental inductance in $\alpha\beta$ -axes as shown in (3.11).

$$[l_{\alpha\beta}] = A(-\theta) \cdot [l_{dq}] \cdot A(\theta) \quad (3.11)$$

According to the literature, the cross-coupling inductance between the stator dq -axes can be assumed equal, $l_{dq} = l_{qd}$ [117], thereby simplifying subsequent analysis. Therefore, each component of the incremental inductance matrix in $\alpha\beta$ -axes is calculated as:

$$[l_{\alpha\beta}] = \begin{bmatrix} l_{\alpha\alpha} & l_{\alpha\beta} \\ l_{\beta\alpha} & l_{\beta\beta} \end{bmatrix} = \begin{bmatrix} \cos(\theta) & -\sin(\theta) \\ \sin(\theta) & \cos(\theta) \end{bmatrix} \cdot \begin{bmatrix} l_{dd} & l_{dq} \\ l_{dq} & l_{qq} \end{bmatrix} \cdot \begin{bmatrix} \cos(\theta) & \sin(\theta) \\ -\sin(\theta) & \cos(\theta) \end{bmatrix} \quad (3.12)$$

where:

$$l_{\alpha\alpha} = l_{dd} \cdot \cos^2(\theta) + l_{qq} \cdot \sin^2(\theta) - 2 \cdot l_{dq} \cdot \cos(\theta) \cdot \sin(\theta) \quad (3.13)$$

$$l_{\alpha\beta} = l_{\beta\alpha} = l_{dq} \cdot (\cos^2(\theta) - \sin^2(\theta)) + l_{dd} \cdot \cos(\theta) \cdot \sin(\theta) - l_{qq} \cdot \cos(\theta) \cdot \sin(\theta) \quad (3.14)$$

$$l_{\beta\beta} = l_{qq} \cdot \cos^2(\theta) + l_{dd} \cdot \sin^2(\theta) + 2 \cdot l_{dq} \cdot \cos(\theta) \cdot \sin(\theta) \quad (3.15)$$

It is pointed out that $[l_{\alpha\beta}]$ is the matrix of the incremental inductances in the $\alpha\beta$ frame while $l_{\alpha\beta}$ is one of the elements of the matrix $[l_{\alpha\beta}]$ representing the cross-coupling incremental inductance between $\alpha\beta$ axes.

Recalling (3.10), the third term of the equation is evaluated by calculating the derivative of the rotational transform matrix:

$$\frac{d}{dt}(A(\theta)) = -\omega \cdot \begin{bmatrix} \sin(\theta) & -\cos(\theta) \\ \cos(\theta) & \sin(\theta) \end{bmatrix} \quad (3.16)$$

where ω is the electrical speed.

Therefore, the following is obtained.

$$A(-\theta) \cdot [l_{dq}] \cdot \frac{d}{dt}(A(\theta)) \cdot \bar{i}_{\alpha\beta} = A(-\theta) \cdot [l_{dq}] \cdot (-\omega) \cdot \begin{bmatrix} \sin(\theta) & -\cos(\theta) \\ \cos(\theta) & \sin(\theta) \end{bmatrix} \cdot \bar{i}_{\alpha\beta} \quad (3.17)$$

Considering that $\bar{i}_{\alpha\beta} = A(-\theta) \cdot \bar{i}_{dq}$, (3.17) becomes:

$$A(-\theta) \cdot [l_{dq}] \cdot (-\omega) \cdot \begin{bmatrix} \sin(\theta) & -\cos(\theta) \\ \cos(\theta) & \sin(\theta) \end{bmatrix} \cdot \begin{bmatrix} \cos(\theta) & -\sin(\theta) \\ \sin(\theta) & \cos(\theta) \end{bmatrix} \cdot \bar{i}_{dq} = A(-\theta) \cdot [l_{dq}] \cdot (-\omega) \cdot [J] \cdot \bar{i}_{dq} \quad (3.18)$$

Therefore, the stator voltage equation in $\alpha\beta$ coordinates is written as:

$$\begin{aligned} \bar{v}_{\alpha\beta} &= R \cdot \bar{i}_{\alpha\beta} + [l_{\alpha\beta}] \cdot \frac{d\bar{i}_{\alpha\beta}}{dt} + A(-\theta) \cdot [l_{dq}] \cdot (-\omega) \cdot [J] \cdot \bar{i}_{dq} + A(-\theta) \cdot [J] \cdot \omega \cdot \bar{\lambda}_{dq} = \\ &R \cdot \bar{i}_{\alpha\beta} + [l_{\alpha\beta}] \cdot \frac{d\bar{i}_{\alpha\beta}}{dt} + A(-\theta) \cdot \{ [l_{dq}] \cdot (-\omega) \cdot [J] \cdot \bar{i}_{dq} + [J] \cdot \omega \cdot \bar{\lambda}_{dq} \} \end{aligned} \quad (3.19)$$

The last term in brackets is equivalent to the back-electromotive force in dq -axes. By multiplying the latter with the rotational transformation matrix $A(-\theta)$, this term becomes the back-electromotive force in $\alpha\beta$ -axes:

$$\bar{e}_{\alpha\beta} = A(-\theta) \cdot \overbrace{\{ [l_{dq}] \cdot (-\omega) \cdot [J] \cdot \bar{i}_{dq} + [J] \cdot \omega \cdot \bar{\lambda}_{dq} \}}^{\bar{e}_{dq}} \quad (3.20)$$

$$\bar{v}_{\alpha\beta} = R \cdot \bar{i}_{\alpha\beta} + [l_{\alpha\beta}] \cdot \frac{d\bar{i}_{\alpha\beta}}{dt} + \bar{e}_{\alpha\beta} \quad (3.21)$$

To transform the voltage equation (3.21) in abc coordinates, it is necessary to add the zero-sequence component as:

$$\bar{v}_{\alpha\beta 0} = \begin{bmatrix} v_\alpha \\ v_\beta \\ v_0 \end{bmatrix} \quad \bar{i}_{\alpha\beta 0} = \begin{bmatrix} i_\alpha \\ i_\beta \\ i_0 \end{bmatrix} \quad \bar{e}_{\alpha\beta 0} = \begin{bmatrix} e_\alpha \\ e_\beta \\ e_0 \end{bmatrix} \quad (3.22)$$

The contribution of the leakage inductance L_{ls} can be added in the zero sequence of the inductance matrix $[l_{\alpha\beta 0}]$. The proposed elaboration neglects L_{ls} in this step to have a separate term representing the leakage flux for clarity of explanation. So the 3×3 $[l_{\alpha\beta 0}]$ matrix is:

$$[l_{\alpha\beta 0}] = \begin{bmatrix} l_{\alpha\alpha} & l_{\alpha\beta} & 0 \\ l_{\alpha\beta} & l_{\beta\beta} & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad (3.23)$$

Then, the inverse Clarke transform $[T]^{-1}$ is applied considering also the zero sequence component as:

$$[T]^{-1} \cdot \bar{v}_{\alpha\beta 0} = R \cdot [T]^{-1} \cdot \bar{i}_{\alpha\beta 0} + [T]^{-1} \cdot [l_{\alpha\beta 0}] \cdot \frac{d\bar{i}_{\alpha\beta 0}}{dt} + [T]^{-1} \cdot \bar{e}_{\alpha\beta 0} \quad (3.24)$$

where the Clarke transform $[T]$ is:

$$[T] = \frac{2}{3} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \\ 1/2 & 1/2 & 1/2 \end{bmatrix} \quad (3.25)$$

Consequently, (3.24) is equal to:

$$v_{abc} = R \cdot i_{abc} + [T]^{-1} \cdot [l_{\alpha\beta 0}] \cdot \frac{di_{abc}}{dt} + e_{abc} \quad (3.26)$$

Taking in account that $\bar{i}_{\alpha\beta 0} = [T] \cdot i_{abc}$ and the Clarke transform is time-independent, (3.26) can be rewritten as:

$$v_{abc} = R \cdot i_{abc} + [T]^{-1} \cdot [l_{\alpha\beta 0}] \cdot [T] \cdot \frac{di_{abc}}{dt} + e_{abc} \quad (3.27)$$

where:

$$l_{abc} = \begin{bmatrix} l_{aa} & l_{ab} & l_{ac} \\ l_{ba} & l_{bb} & l_{bc} \\ l_{ca} & l_{cb} & l_{cc} \end{bmatrix} = [T]^{-1} \cdot [l_{\alpha\beta 0}] \cdot [T] \quad (3.28)$$

Finally, the terms of the matrix of incremental inductances in abc are:

$$\begin{aligned} l_{aa} &= \frac{2}{3} \cdot l_{\alpha\alpha} \\ l_{ab} &= l_{ba} = \frac{1}{3} \cdot \left[-l_{\alpha\alpha} + \sqrt{3} \cdot l_{\alpha\beta} \right] \\ l_{ac} &= l_{ca} = \frac{1}{3} \cdot \left[-l_{\alpha\alpha} - \sqrt{3} \cdot l_{\alpha\beta} \right] \\ l_{bb} &= \frac{1}{3} \cdot \left[\frac{l_{\alpha\alpha}}{2} + \frac{3}{2} \cdot l_{\beta\beta} - \sqrt{3} \cdot l_{\alpha\beta} \right] \\ l_{bc} &= l_{cb} = \frac{1}{3} \cdot \left[\frac{l_{\alpha\alpha}}{2} - \frac{3}{2} \cdot l_{\beta\beta} \right] \\ l_{cc} &= \frac{1}{3} \cdot \left[\frac{l_{\alpha\alpha}}{2} + \frac{3}{2} \cdot l_{\beta\beta} + \sqrt{3} \cdot l_{\alpha\beta} \right] \end{aligned} \quad (3.29)$$

The contribution of leakage inductance is now added to the (3.27), obtaining:

$$v_{abc} = R \cdot i_{abc} + L_{ls} \cdot \frac{di_{abc}}{dt} + [l_{abc}] \cdot \frac{di_{abc}}{dt} + e_{abc} \quad (3.30)$$

The (3.30) represents the Stator-VBR formulation exploiting the flux maps elaboration.

3.2.2 Proposed Full-VBR formulation

The standard VBR formulation described above addresses only stator dynamics under constant excitation. To fully integrate the rotor circuit dynamics, the flux derivative expressed in (3.4) is expanded to include the rotor current derivative. This yields a generalized expression within the augmented $(dq0, f)$ reference frame.

$$\bar{\lambda}_{dq0,f} = \begin{bmatrix} \lambda_d \\ \lambda_q \\ \lambda_0 \\ \lambda_f \end{bmatrix} \quad \bar{i}_{dq0,f} = \begin{bmatrix} i_d \\ i_q \\ i_0 \\ i_f \end{bmatrix} \quad (3.31)$$

$$\frac{d\bar{\lambda}_{dq0,f}}{dt} = \frac{\partial \bar{\lambda}_{dq0,f}}{\partial \bar{i}_{dq0,f}} \cdot \frac{d\bar{i}_{dq0,f}}{dt} \quad (3.32)$$

The cross-coupling incremental inductances between the stator $dq0$ axes and the rotor field circuit and the rotor self inductance (3.33) appear and need to be added to the inductance matrix (3.6).

$$l_{df} = \frac{\partial \lambda_d}{\partial i_f} \quad l_{qf} = \frac{\partial \lambda_q}{\partial i_f} l_{fd} = \frac{\partial \lambda_f}{\partial i_d} \quad l_{fq} = \frac{\partial \lambda_f}{\partial i_q} \quad l_{ff} = \frac{\partial \lambda_f}{\partial i_f} \quad (3.33)$$

The complete incremental inductance matrix in a $dq0, f$ frame, including also the leakage inductance L_{ls} to have a more compact form, becomes:

$$\frac{d}{dt} \begin{bmatrix} \lambda_d \\ \lambda_q \\ \lambda_0 \\ \lambda_f \end{bmatrix} = \begin{bmatrix} l_{dd} & l_{dq} & 0 & l_{df} \\ l_{qd} & l_{qq} & 0 & l_{qf} \\ 0 & 0 & L_{ls} & 0 \\ l_{fd} & l_{fq} & 0 & l_{ff} \end{bmatrix} \cdot \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \\ i_0 \\ i_f \end{bmatrix} = [l_{dq0,f}] \cdot \frac{d\bar{i}_{dq0,f}}{dt} \quad (3.34)$$

The partial derivatives of (3.5) and (3.33) can be elaborated from the flux maps and stored in inductance maps as described in Chapter 2.5. The incremental inductances of the EESM motor under consideration are shown in Figs. 2.27-2.35.

To derive the new VBR model in (abc, f) frame, it is necessary to apply to (3.34) an extended version of the Park transformation matrix $[K]$ that takes into account also the rotor terms (3.35):

$$[K] = \frac{2}{3} \begin{bmatrix} \cos(\theta) & \cos(\theta - \frac{2}{3}\pi) & \cos(\theta + \frac{2}{3}\pi) & 0 \\ -\sin(\theta) & -\sin(\theta - \frac{2}{3}\pi) & -\sin(\theta + \frac{2}{3}\pi) & 0 \\ 1/2 & 1/2 & 1/2 & 0 \\ 0 & 0 & 0 & 3/2 \end{bmatrix} \quad (3.35)$$

The rotor axis, as evidenced by (3.35), does not experience any rotational transformation. Consequently, the new formulation of the incremental inductance matrix in

the (abc, f) form becomes:

$$l_{abc,f} = \begin{bmatrix} l_{aa} & l_{ab} & l_{ac} & l_{af} \\ l_{ba} & l_{bb} & l_{bc} & l_{bf} \\ l_{ca} & l_{cb} & l_{cc} & l_{cf} \\ l_{fa} & l_{fb} & l_{fc} & l_{ff} \end{bmatrix} = [K]^{-1} \cdot [l_{dq0,f}] \cdot [K] \quad (3.36)$$

From (3.20), the dq back-emf is:

$$\bar{e}_{dq} = [l_{dq}] \cdot (-\omega) \cdot [J] \cdot \bar{i}_{dq} + [J] \cdot \omega \cdot \bar{\lambda}_{dq} \quad (3.37)$$

To complete the VBR model, the back-emf is also elaborated considering the frame $dq0, f$ as:

$$\bar{e}_{dq0,f} = [l_{dq0,f}] \cdot (-\omega) \cdot [J_{4 \times 4}] \cdot \bar{i}_{dq0,f} + [J_{4 \times 4}] \cdot \omega \cdot \bar{\lambda}_{dq0,f} \quad (3.38)$$

Recalling from (3.18) that $[J] = \frac{dA(\theta)}{dt} \cdot A(-\theta)$ and that the rotor axis does not experience any rotation from $dq0, f \rightarrow abc, f$, the extended $[J]$ becomes:

$$\begin{bmatrix} -\sin(\theta) & \cos(\theta) & 0 & 0 \\ -\cos(\theta) & -\sin(\theta) & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} \cos(\theta) & -\sin(\theta) & 0 & 0 \\ \sin(\theta) & \cos(\theta) & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} = [J_{4 \times 4}] = \begin{bmatrix} 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \quad (3.39)$$

According to (3.38), a voltage component, behaving as an equivalent back-emf (3.38), appears also in the rotor equivalent circuit and is defined accordingly as:

$$e_f = (l_{fd} \cdot i_q - l_{fq} \cdot i_d) \cdot \omega \quad (3.40)$$

This contribution appears in the VBR formalization to depict the armature reaction, i.e., the effect of the stator currents induced voltage on the rotor circuit. Applying to (3.38) the transformation matrix $[K]$ (3.35), it is obtained $e_{abc,f}$. The final voltage equation in abc, f form is:

$$v_{abc,f} = R \cdot i_{abc,f} + [l_{abc,f}] \cdot \frac{di_{abc,f}}{dt} + e_{abc,f} \quad (3.41)$$

3.3 Stator leakage inductance computation

The proposed VBR formulation imposes no intrinsic constraints on the methodology used to derive the stator leakage inductance, rendering it compatible with various strategies documented in the literature [170]. However, precise quantification of the stator leakage inductance is critical in specific applications, particularly those involving non-isolated neutral points or dual-bridge inverter topologies [183]. In such configurations, the fidelity of the zero-sequence current simulation depends on a reliable estimate of this parameter. Consequently, this work introduces a rigorous

analytical extraction method obtained by directly manipulating flux maps. The identification procedure is based on the equivalent circuit illustrated in Fig. 3.6, wherein the superscripts ' and '' denote quantities referred to the stator and rotor magnetic domains, respectively.

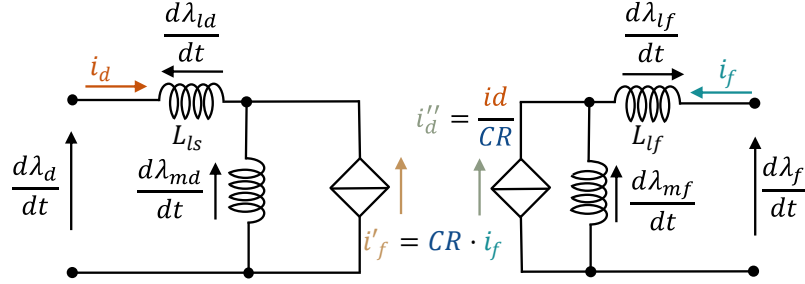


Fig. 3.6 Equivalent circuit of the EESM considering the cross-coupling with the rotor circuit (stator and rotor resistances are not reported for simplicity).

The computational procedure for determining the stator leakage inductance is detailed in the following sequential steps.

1. **Flux evaluation at null stator d -axis current:** The rotor and stator d -axis flux linkages are first evaluated under the condition of zero stator d -axis current:

$$\begin{cases} \lambda_f |_{i_d=0} = \lambda_f(0, i_q, i_f) \\ \lambda_d |_{i_d=0} = \lambda_d(0, i_q, i_f) \end{cases} \quad (3.42)$$

In this state, the stator-side d -axis magnetizing flux λ_{md} corresponds directly to the total stator flux, as the leakage component is nullified:

$$\lambda_{md} |_{i_d=0} = \lambda_d |_{i_d=0} - L_{ls} \cdot \overset{0}{i_d} = \lambda_d |_{i_d=0} \quad (3.43)$$

Simultaneously, leveraging the rotor leakage inductance L_{lf} , which can be easily determined via FEA or experimentally derived as described in [151], the magnetizing flux referred to the rotor side λ_{mf} is calculated as:

$$\lambda_{mf} |_{i_d=0} = \lambda_f |_{i_d=0} - \lambda_{lf} |_{i_d=0} = \lambda_f |_{i_d=0} - L_{lf} \cdot i_f \quad (3.44)$$

2. **Flux evaluation at null rotor current:** Subsequently, the flux linkages are evaluated under the condition of zero rotor current:

$$\begin{cases} \lambda_f |_{i_f=0} = \lambda_f(i_d, i_q, 0) \\ \lambda_d |_{i_f=0} = \lambda_d(i_d, i_q, 0) \end{cases} \quad (3.45)$$

Consequently, the magnetizing flux referred to the rotor side is straightforwardly obtained, as the rotor leakage drop is zero:

$$\lambda_{mf} |_{i_f=0} = \lambda_f |_{i_f=0} - L_{lf} \cdot \overset{0}{i_f} = \lambda_f |_{i_f=0} \quad (3.46)$$

3. **Determination of the equivalent current:** Referring to the equivalent circuit in Fig. 3.6, two distinct magnetization scenarios are analyzed:

- Under zero field current ($i_f = 0$), the rotor magnetizing flux ($\lambda_{mf} |_{i_f=0}$) is sustained solely by the stator d -axis current referred to the rotor side, denoted as i_d'' .
- Under zero stator d -axis current ($i_d = 0$), the rotor magnetizing flux ($\lambda_{mf} |_{i_d=0}$) is sustained solely by the rotor current i_f .

This comparison allows for the numerical determination of a fictitious stator current referred to the rotor side $i_d''^*$, which produces an equivalent magnetizing effect to that of the rotor current i_f :

$$i_d''^* : \lambda_{mf}(i_f) |_{i_d=0} = \lambda_{mf}(i_d''^*) |_{i_f=0} \quad (3.47)$$

While the numerical implementation of (3.47) is methodologically consistent, practical challenges arise due to the domain limits of the flux maps. Specifically, satisfying (3.47) requires calculating $i_d''^*$ values that may correspond to stator currents significantly exceeding the machine's rated limits. This disparity stems from the typically high turns ratio characteristic of EESMs. Consequently, the armature reaction current i_d'' required to replicate the flux generated by the rotor winding is scaled by the current ratio ($CR = i_d/i_d''$), often necessitating the extrapolation of the direct flux maps into deep saturation regions.

4. **Computation of stator magnetizing flux:** For each derived combination of $i_d''^*$ and i_q , the stator d -axis flux is computed under the hypothetical condition where $i_d = 0$ and the rotor is excited by $i_f = i_d''^*$. In this state, the total stator flux is equal to the magnetizing flux:

$$\lambda_d(0, i_q, i_f = i_d''^*) = \lambda_{md}(0, i_q, i_f = i_d''^*) \quad (3.48)$$

By the definition of the equivalent current $i_d''^*$, the following equality holds:

$$\lambda_{md}(0, i_q, i_f = i_d''^*) = \lambda_{md}(i_d, i_q, 0) = \lambda_{md} |_{i_f=0} \quad (3.49)$$

5. **Stator leakage inductance calculation:** Combining (3.45) and (3.49), the stator leakage flux component is isolated as:

$$\lambda_{ld} = \lambda_d |_{i_f=0} - \lambda_{md} |_{i_f=0} \quad (3.50)$$

Finally, the stator leakage inductance L_{ls} is computed as:

$$L_{ls}(i_d, i_q) = \frac{\lambda_{ld}(i_d, i_q)}{i_d} \quad (3.51)$$

By applying this procedure across the full operating range of (i_d, i_q, i_f) , the stator leakage inductance map $L_{ls}(i_d, i_q)$ is derived, as illustrated in Fig. 3.7 . It is observed that L_{ls} exhibits negligible dependence on i_f , a characteristic attributable to the fact that leakage flux is primarily governed by local saturation phenomena within the stator slot leakage paths, which are largely decoupled from the main air-gap flux.

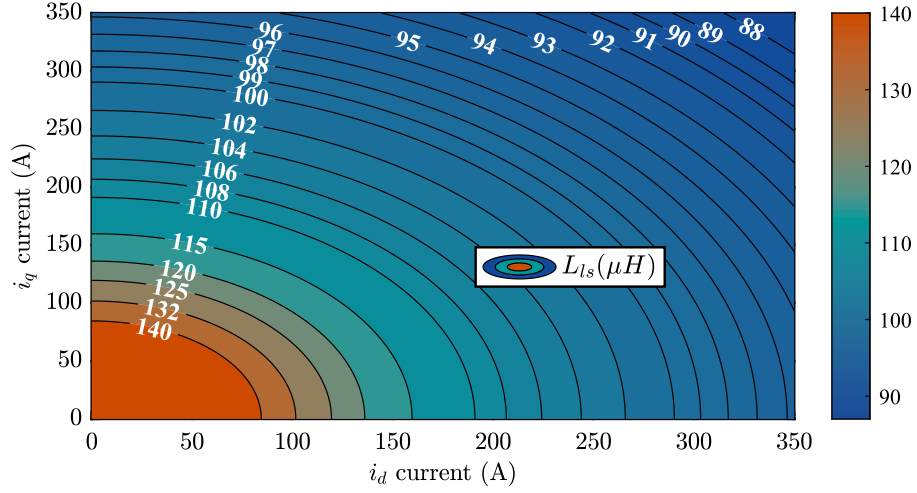


Fig. 3.7 View of the stator leakage inductance for the EESM under consideration.

Furthermore, the proposed identification methodology facilitates the extraction of the effective current ratio (CR) between the stator and rotor circuits. Utilizing (3.52), the CR is defined as:

$$CR(i_d, i_q, i_f) = \frac{i_d}{i_d^*} \quad (3.52)$$

For the EESM under investigation, the computed $CR(i_d, i_q, i_f)$ surfaces for various rotor current levels are depicted in Fig. 3.8 . Notably, the CR is strictly dependent on the operating point, exhibiting significant variation due to nonlinear magnetic saturation. It is emphasized that while the accurate computation of the variable CR provides valuable conceptual insight into the machine's cross-coupling behavior, this parameter is not strictly required for the implementation of the proposed VBR model.

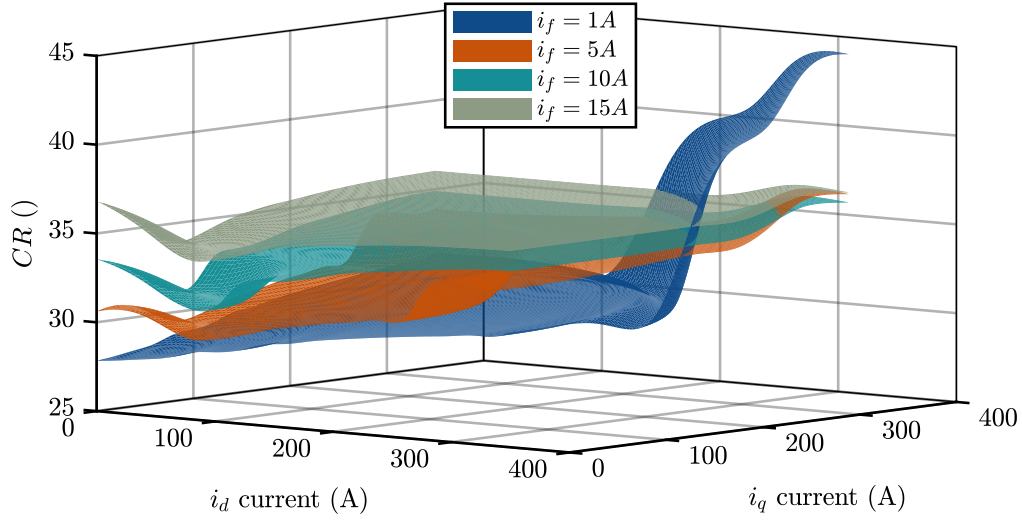


Fig. 3.8 View of the current ratio for the EESM under consideration.

3.4 Simulation validation

The proposed VBR architecture was implemented in the PLECS simulation environment, selected for its optimized solvers capable of delivering high-fidelity circuit simulations. The validation has been pursued considering the EESM described in Chapter 2.3. The key specifications of this machine are detailed in Table 2.1, while the characteristic flux linkage maps are illustrated in Fig. 2.7 and Fig. 2.23.

To benchmark the performance, the proposed formulation is compared against two established modeling paradigms: the inverse flux model [51] and the standard Full-VBR model documented in the literature [57]. The latter is readily available in the PLECS component library as the “*Synchronous Machine (Salient Pole, Saturable)*” block. The selection of PLECS facilitates a direct, consistent comparison between the proposed method and the literature implementation.

Magnetic saturation characteristics are implemented via three-dimensional LUTs. A significant computational advantage of the proposed VBR model is its reduced memory footprint: the direct flux maps require a grid of only $21 \times 21 \times 21$ points (approx. 0.254 MB). In contrast, the inverse flux maps utilized in [51] necessitate a much denser discretization ($101 \times 101 \times 101$ points, approx. 50 MB) to maintain convergence stability and accuracy during the flux-to-current inversion process.

All simulation scenarios were executed using the variable-step *RADAU* solver within PLECS, with a maximum step size constrained to $10 \mu s$ and a relative tolerance of 1×10^{-3} . Computations were performed on a workstation equipped with a 3.8-GHz Intel processor running the Windows 11 operating system.

The validation protocol entails a comprehensive sweep of the operational domain by testing permutations of the current vector components (i_d^*, i_q^*, i_f^*) . These values serve as the reference targets for the models under test, ensuring that validation covers the full operating range of the machine.

The validation methodology is schematically illustrated in Fig. 3.9. Initially, the reference flux linkages $(\lambda_d^*, \lambda_q^*, \lambda_f^*)$ are derived from the commanded current vector (i_d^*, i_q^*, i_f^*) via direct interpolation of the machine's magnetic flux maps. These reference values are compared against the instantaneous flux estimates $(\lambda_d, \lambda_q, \lambda_f)$ produced by the EESM model under test. The resulting flux error vector is regulated to zero utilizing three ideal PI controllers, with infinite bandwidth. These regulators operate without output saturation or bandwidth constraints to ensure the rigorous nullification of the flux error.

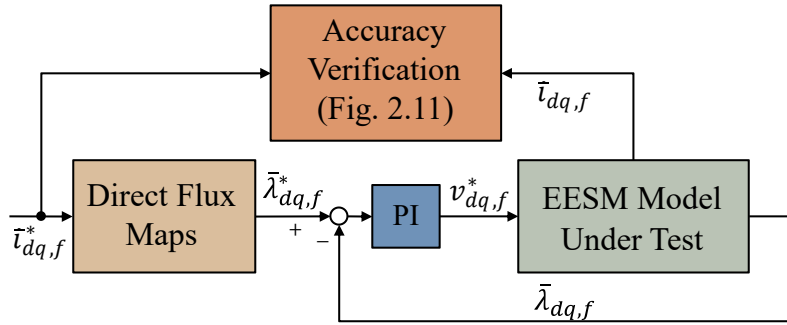


Fig. 3.9 Block diagram of the validation scheme used in the simulation environment.

The theoretical premise of this validation strategy is as follows: if the closed-loop control successfully eliminates the flux error, the model's instantaneous flux state aligns perfectly with the reference trajectory (i.e., $\lambda_{dq,f} = \lambda_{dq,f}^*$). According to the unique constitutive relationship defined by the direct flux maps, this specific flux state must physically correspond to the reference current triplet (i_d^*, i_q^*, i_f^*) . Consequently, the fidelity of the model is quantified by the discrepancy between the simulated effective currents (i_d, i_q, i_f) and the reference currents. A minimal deviation between these quantities indicates high modeling accuracy. It is noted that the output of the

PI flux regulators corresponds physically to the machine's stator and rotor voltages, a topology consistent with FVC architectures documented in the literature [116].

Fig. 3.10 presents the comparative analysis of the current tracking performance for the three considered modeling paradigms: the inverse flux map approach [51], the standard Full-VBR component available in PLECS [57], and the proposed VBR formulation. To ensure a rigorous validation spanning the entire operational range of the machine, the reference current trajectories are synthesized to sweep the complete (i_d, i_q, i_f) space. The reference signals are defined as follows:

$$\begin{cases} i_d^* = i^* \cdot \cos(\gamma^*) \\ i_q^* = i^* \cdot \sin(\gamma^*) \\ i_f^* = i_{f,max} \cdot \sin(2\pi f_{if} t) \end{cases} \quad (3.53)$$

where the amplitude (i^*) and phase angle (γ^*) of the stator current vector are modulated as:

$$\begin{cases} i^* = i_{max} \cdot \sin(2\pi f_i t) \\ \gamma^* = \pi \cdot \sin(2\pi f_\gamma t) \end{cases} \quad (3.54)$$

The current limits are established at $i_{max} = 385$ A and $i_{f,max} = 15$ A, consistent with the primary machine parameters listed in Table 2.1. The frequencies of the sinusoidal variations are selected as $f_i = 1$ Hz, $f_\gamma = 10$ Hz, and $f_{if} = 0.1$ Hz to sweep the complete (i_d, i_q, i_f) space in a single test.

The experimental results presented in Fig. 3.10 demonstrate that the proposed Full-VBR model achieves precise reference tracking, even under highly dynamic operating conditions. Conversely, the conventional VBR formulation [57] exhibits substantial deviations, particularly along the d -axis. This discrepancy, which is also observable in the phase-domain coordinates, is consistent with the theoretical limitations discussed in Chapter 3.1; specifically, the conventional model fails to accurately resolve the nonlinear magnetic coupling between the stator d -axis and the rotor current under conditions of deep saturation. It is observed, however, that the conventional formulation retains its validity within linear magnetic operating regions where saturation effects are negligible. Notably, the inverse flux map model yields performance equivalent to the proposed VBR formulation, despite lacking a physical circuit topology.

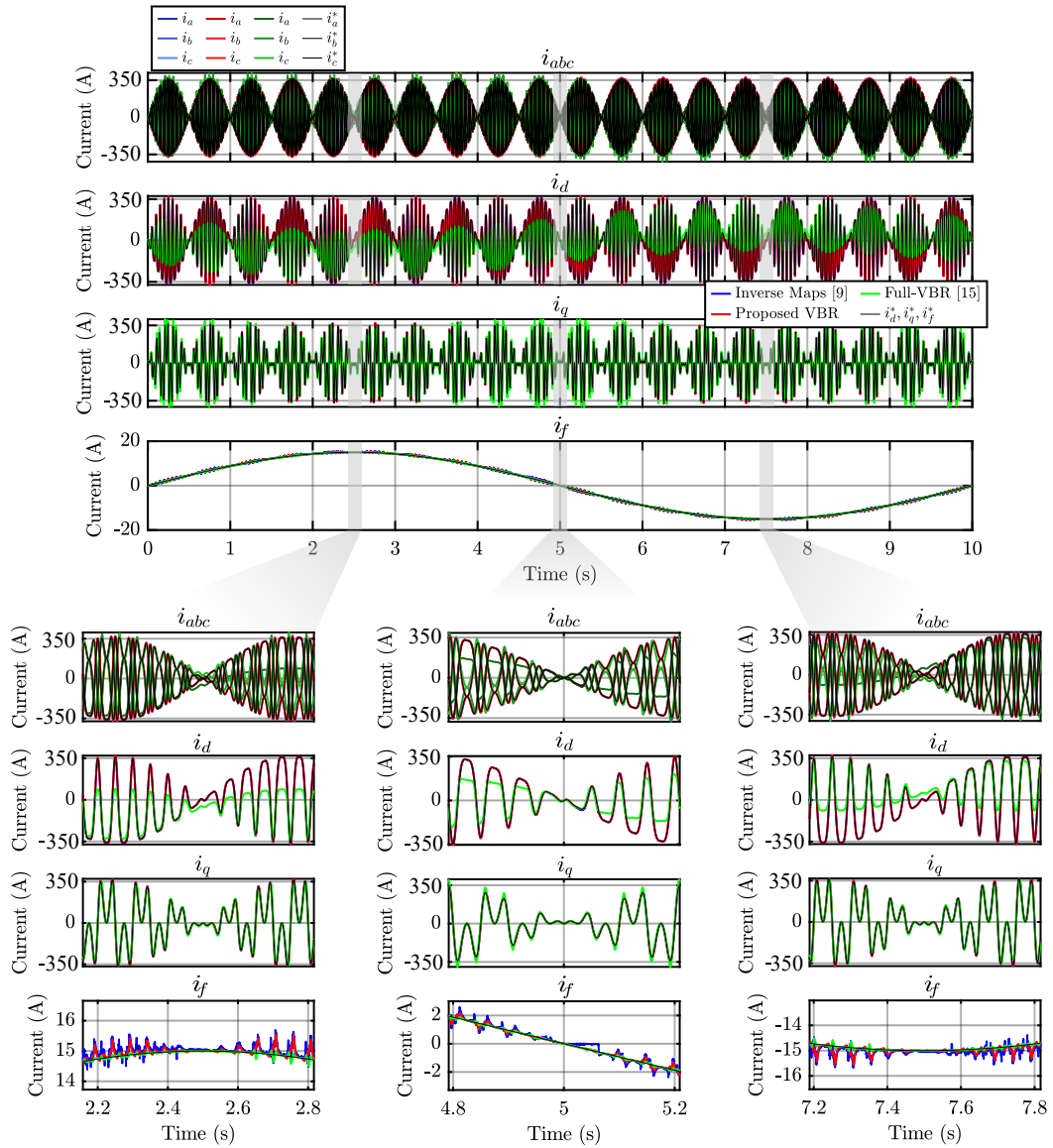


Fig. 3.10 Accuracy verification of EESM models in simulation environment. From left to right: maximum positive rotor current; zero rotor current; maximum negative rotor current.

The inverse flux map model also serves as a robust benchmark for validating the proposed stator leakage inductance extraction method, as it inherently take into account the total flux linkage, including the leakage component, within its definition. Therefore, to validate the stator leakage inductance profile identified in Chapter 3.3, a comparative analysis was conducted between the inverse flux map model and the proposed VBR formulation under two distinct configurations: with and without the identified stator leakage inductance profile applied. The validation utilizes the same

closed-loop current control procedure detailed in Fig. 3.9. The results are presented in Fig. 3.11, which provides a magnified view of the d -axis stator current at peak amplitude. It is observed that incorporating the identified stator leakage profile enables the proposed VBR model to closely replicate the response of the inverse flux map benchmark. Furthermore, the proposed method yields a smoother current trajectory, mitigating numerical chattering. This convergence serves as a robust validation of the analytical procedure proposed for leakage inductance computation.

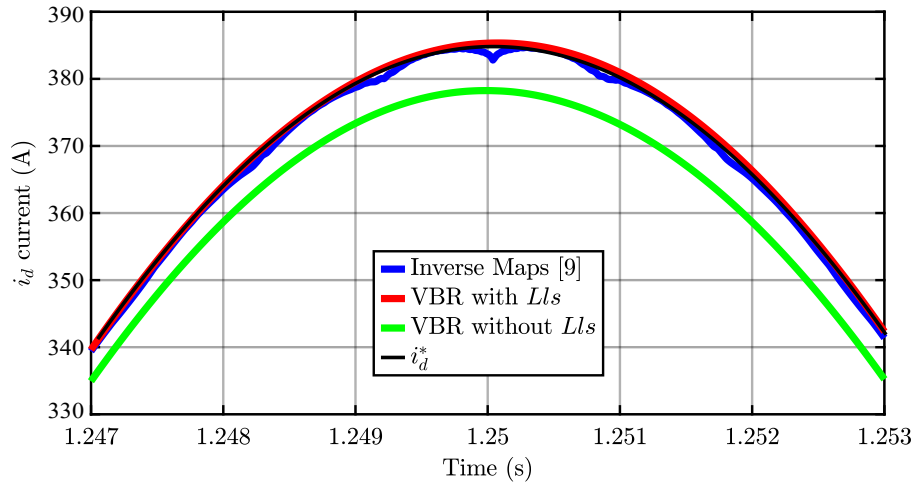


Fig. 3.11 Accuracy verification of EESM models in simulation environment.

A comparative assessment of computational efficiency was also performed for the simulation scenarios described above. The inverse flux map model [51] demonstrated the highest efficiency, completing the simulation in 12 s. This performance is expected, as the model relies exclusively on LUT interpolation and algebraic operations without solving differential equations for circuit dynamics. In contrast, the circuit-based VBR formulations incurred higher computational costs: the standard Full-VBR model [57] required 1 min 32 s, while the proposed VBR model concluded in 1 min 53 s. The slight increase in execution time for the proposed method is attributable to the calculation of state-dependent variable parameters at each time step, as opposed to the fixed-parameter structure of the conventional model. However, this marginal increase in computational burden is fully justified by the superior accuracy achieved, particularly under conditions of significant magnetic saturation.

3.5 Experimental validation

The experimental validation was conducted on a test rig where the MUT is mechanically coupled to a DM operating as a speed-controlled prime mover, as illustrated in Fig. 3.12. This configuration facilitates characterization under both current- and torque-controlled scenarios. To validate the torque estimates provided by the VBR model, the mechanical shaft torque is measured via a high-precision HBM T40B torque sensor installed between the machines.

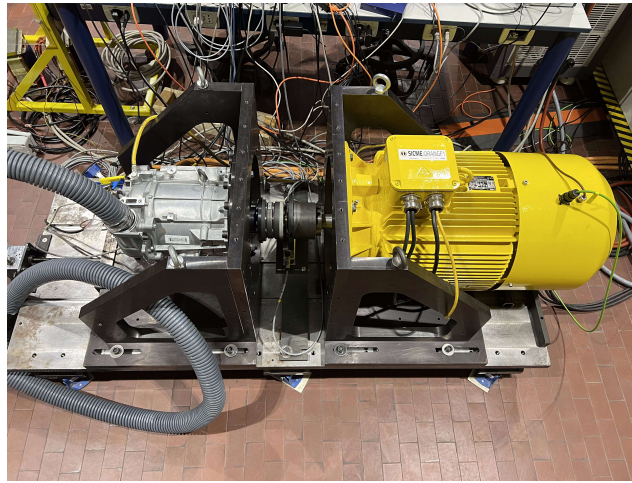


Fig. 3.12 View of the test rig: EESM under test (left), torque sensor (center), driving machine (right).

The stator winding is powered by a three-phase IGBT VSI supplied by a bidirectional 350 V DC power supply. The inverter operates at a switching frequency of 5 kHz with an inserted dead time of 5 μ s. Rotor excitation is provided by a dedicated IGBT full-bridge DC/DC converter sharing the common DC link, operating at 20 kHz with a 500 ns dead time. Control execution is handled by two synchronized dSPACE MicroLabBox platforms (one for the stator and one for the rotor control), with sampling rates synchronized to the respective switching frequencies (single-edge PWM modulation). The control algorithms were implemented in MATLAB/Simulink and deployed via automatic C-code generation.

High-fidelity instrumentation was employed for electrical measurements. Phase currents are acquired using high-bandwidth LEM IT 605-S ULTRASTAB closed-loop current sensors. Line-to-line PWM voltages are measured via shielded cables connected to the high-voltage HBM GmbH GN610B input cards. All analog signals

are sampled and stored by a calibrated HBM GEN7tA transient recorder at a sampling rate of 2 MS/s.

3.5.1 Experimental results

The experimental validation of the proposed VBR framework was conducted through two distinct digital twin test scenarios. The classification of this approach as a "*digital twin*" is supported by the testing methodology, in which the simulation model is driven by the exact physical inputs recorded from the MUT, specifically the instantaneous line-to-line PWM voltages and the mechanical speed profile. This configuration enables a rigorous, time-domain comparison of the model's predictive outputs, namely stator phase currents (i_a, i_b, i_c), rotor excitation current i_f , and mechanical torque T , against experimental measurements. The specific testing protocols are detailed below.

Armature reaction test

A distinct advantage of the proposed VBR formulation is its ability to rigorously reproduce the electromagnetic cross-coupling effects between the stator and rotor windings. To validate this characteristic, a specific transient test was performed.

The rotor winding was operated in an open-circuit configuration to enforce a null rotor current. Simultaneously, the stator currents were regulated using a standard CVC scheme. To maximize the armature reaction effect on the rotor circuit, the q -axis current was regulated to zero, while the d -axis current was driven from a steady state of 400 A to -400 A. The transition was governed by a slew rate of 800 A/s, a value selected as a trade-off to induce a measurable electromotive force on the rotor while maintaining a temporal window sufficient for accurate acquisition. Theoretically, the induced voltage measured at the rotor terminals should align with the predictions of the VBR model.

However, direct comparison of the instantaneous voltage waveforms presents significant experimental challenges. Since the stator currents are regulated via a PWM inverter, high-frequency switching harmonics induce substantial noise on the rotor back-EMF. While this issue is easily mitigated in simulation environments (PLECS or FEA) by employing averaged voltage sources, experimental validation

requires a robust post-processing strategy to filter the PWM ripple. To address this, the validation relies on the integral relationship of the magnetic quantities. As already discussed in Chapter 2.4.2, based on the physical symmetry of the EESM, the initial flux linkage can be derived from the integral of the measured induced voltage (2.25). The experimental application of (2.25) is robust, as the numerical integration inherently acts as a low-pass filter, effectively suppressing the high-frequency PWM noise associated with the stator switching.

The validation results are presented in Fig. 3.13. Despite the smooth trajectory of the d -axis current, the raw measured rotor voltage remains highly noisy. However, the underlying profile of the induced voltage and the derived flux linkage aligns closely with the trajectories computed via FEA and the proposed VBR simulation.

Quantitative analysis confirms this correlation: the simulation predicts a flux linkage value of 9.698 Vs. Conversely, applying the integral method (2.25) to the experimental data yields a value of 9.883 Vs. The resulting discrepancy is approximately 2%, a deviation considered negligible given the inherent uncertainties in voltage sensing, numerical integration, and finite element discretization.

Further corroboration is provided by the auxiliary waveforms in Fig. 3.13: the moving average of the measured rotor voltage $\bar{V}_f$ and the real-time reconstruction of the flux linkage λ_f exhibit excellent agreement with the digital twin predictions. Consequently, this test unequivocally demonstrates the proposed VBR model's ability to replicate the dynamic magnetic coupling between the stator and rotor domains.

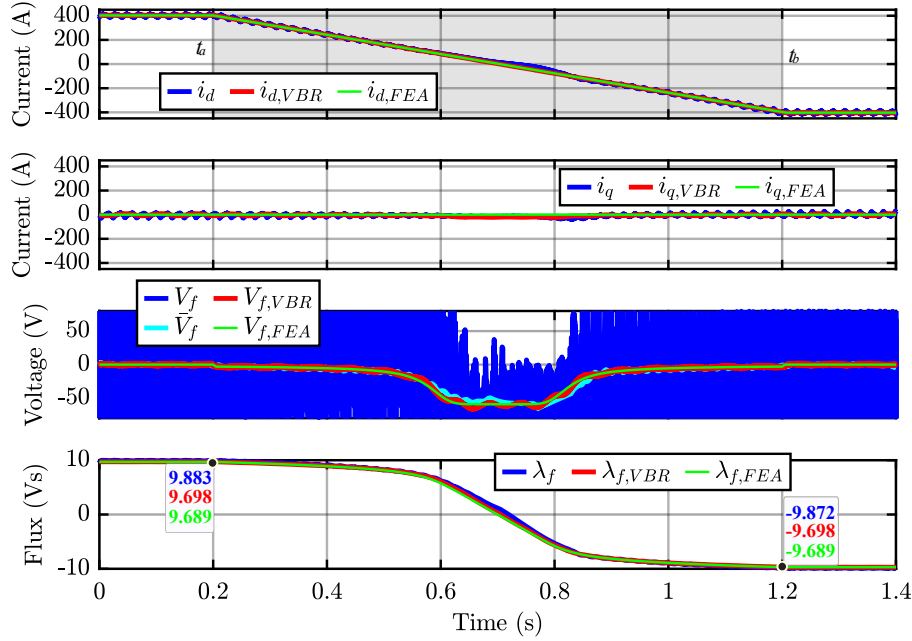


Fig. 3.13 Armature reaction test: experimental results.

Fast torque transient

The second validation experiment involved the proposed VBR model concurrently with the MUT's torque control. In this configuration, the instantaneous line-to-line voltages applied by the converters and the mechanical speed imposed by the DM were fed in real-time to the simulation model. The validation was therefore based on a direct comparison between the simulated state variables (currents and torque) and the corresponding experimental measurements acquired from the test rig.

To ensure a rigorous assessment of the electromagnetic model, the test was conducted at a low rotational speed (500 r/min), thereby rendering negligible the effects of high-frequency phenomena such as skin effect, iron losses, and mechanical friction. The integration of these parasitic effects into the VBR framework remains a subject for future research. The torque regulation was implemented using a DFVC scheme; detailed analysis of this control strategy can be found in [63] and in the next chapter of this thesis.

The test procedure involved a rapid torque transient, commanding a step from 0 Nm to the peak torque of 245 Nm at a slew rate of 1000 Nm/s. The experimental results, presented in Fig. 3.14, demonstrate the high fidelity of the proposed VBR

model, as it accurately reproduces the experimental current and torque trajectories. Notably, the model successfully reproduces high-frequency phenomena, including the PWM current ripple and the voltage distortion induced by inverter dead-time.

Certain discrepancies are observed in the rotor current and torque signals, which are attributable to instrumentation characteristics rather than modeling errors. Specifically, the experimental rotor current exhibits a slight delay relative to the simulation and reference signals, caused by communication latency in the acquisition chain. Regarding the mechanical torque, the VBR model explicitly accounts for the electromagnetic torque ripple generated by the switching harmonics. In contrast, the experimental torque signal appears smooth due to the limited bandwidth of the torque transducer and the low-pass filtering applied during signal conditioning.

In conclusion, this test confirms the capability of the proposed VBR formulation to provide accurate, high-bandwidth predictions in both transient and steady-state operating regimes.

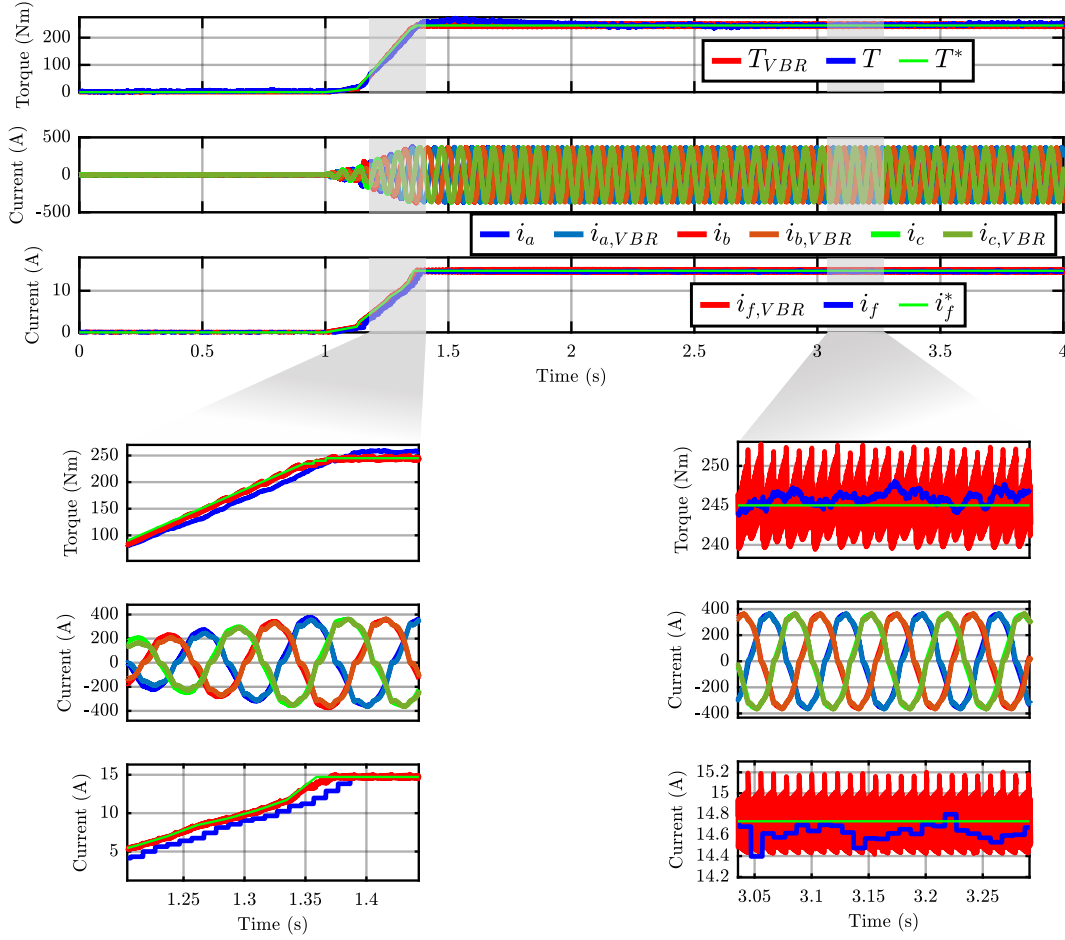


Fig. 3.14 Fast torque transient test: experimental results.

3.6 Summary

This chapter establishes a high-fidelity digital twin platform for EESMs, specifically addressing the modeling challenges imposed by traction applications. The core contribution is the development of a novel Full-VBR formulation. Unlike conventional models that rely on simplified saturation curves, this architecture integrates high-resolution, multidimensional flux maps. This approach enables the explicit calculation of a state-dependent 4×4 incremental inductance matrix, allowing the model to accurately reproduce deep magnetic saturation and dynamic stator-rotor cross-coupling effects within a circuit-oriented environment.

Furthermore, the chapter introduces a rigorous analytical methodology for directly extracting the stator leakage inductance L_{ls} from global flux maps. This procedure isolates the leakage component by evaluating magnetic behaviors under specific zero-current conditions in the stator and rotor domains, providing a critical parameters for accurate zero-sequence modeling.

The proposed framework was validated through a dual-stage process:

- **Simulation validation:** Comparative analysis against inverse flux map models and standard Full-VBR approach demonstrated that the proposed formulation achieves superior accuracy in tracking reference currents under dynamic conditions, while maintaining computational efficiency suitable for system-level simulations.
- **Experimental validation:** A "digital twin" testing protocol was executed on a commercial EESM automotive sample. By feeding real-time voltage inputs from the test rig into the simulation, the model's outputs were cross-referenced against physical measurements. Specific tests, including an open-rotor armature reaction test and fast torque transients, confirmed the model's ability to replicate high-frequency phenomena and complex magnetic coupling with high fidelity.

The description and the validation of the proposed Full-VBR model for EESMs have been published in the journal article (J1).

Chapter 4

EESM torque control

This chapter details the development and implementation of a high-performance torque control architecture for EESMs. The primary objective is to define a Unified Torque Controller (UTC), a versatile control structure capable of reconfiguring itself to execute various Torque Control Approaches (TCAs) within a single framework. This unified topology facilitates the seamless selection of the optimal control strategy dictated by specific application requirements and machine design constraints.

As discussed in Chapter 1.6, the state-of-the-art in AC traction drives encompasses several mature control paradigms, each characterized by a distinct set of regulated state variables (CVC, DTC, DFVC, FPC, FVC).

The proposed UTC framework is designed to accommodate all the aforementioned strategies. Furthermore, the control scheme integrates a loss-minimization algorithm that dynamically optimizes the reference trajectories of the controlled variables to minimize the overall copper losses (stator+rotor). Finally, to comply with the stringent safety protocols inherent in automotive applications, the controller incorporates variable current limits that protect critical powertrain components from thermal overload across all operating conditions.

The schematic view of the EESM drive topology analyzed in this work is shown in Fig. 4.1. The EESM is supplied by the inverter on the stator and by an exciter on the rotor. The torque control scheme generates the reference voltages v_{abc}^* for the inverter duty-cycle generation d_{abc}^* , along with the voltage reference v_f^* for the exciter.

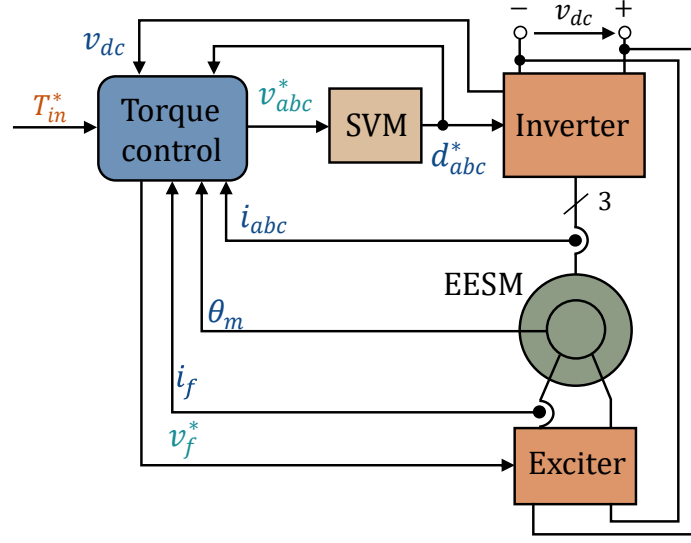


Fig. 4.1 Schematic view of EESM drive topology.

4.1 Torque control scheme

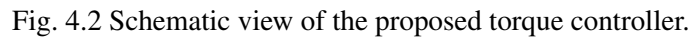
The schematic view of the proposed torque controller is shown in Fig. 4.1 and Fig. 4.2. According to the input torque reference T_{in}^* , the torque controller generates the stator reference phase voltages $v_{abc}^* = [v_a^*, v_b^*, v_c^*]$ and the rotor reference voltage v_f^* according to the following feedback quantities:

- Stator phase currents $i_{abc} = [i_a, i_b, i_c]$.
- Inverter DC-link voltage v_{dc} .
- Rotor mechanical position θ_m .
- Rotor current i_f .

The duty-cycles of the inverter $d_{abc}^* = [d_a^*, d_b^*, d_c^*]$ are computed from the stator reference phase voltages according to the implemented PWM technique. This work considers carrier-based space vector modulation (SVM). As shown in Fig. 4.2, the developed UTC scheme implements the following control blocks:

1. Stator flux observer
2. Flux and currents limits computation

- ## 6. Rotor current regulator



A detailed description of the above control modules is provided in the following subsections, along with a discussion on the implementation of LUTs necessary to deal with the strong non-linear dependence of EESM's magnetic behavior with the rotor current i_f . In the description of the control modules, the following superscripts “ $\sim$ ”, “ $\wedge$ ”, and “ $*$ ” denote an estimated, observed, and reference variable/parameter, respectively. Estimated quantities arise from open-loop estimation schemes, while observed quantities are obtained from closed-loop observers incorporating measurement feedback.

4.1.1 Stator flux observer

The stator flux vector is estimated by adopting a consolidated flux observer structure [109] that combines two model-based estimators, as shown in Fig. 4.3.

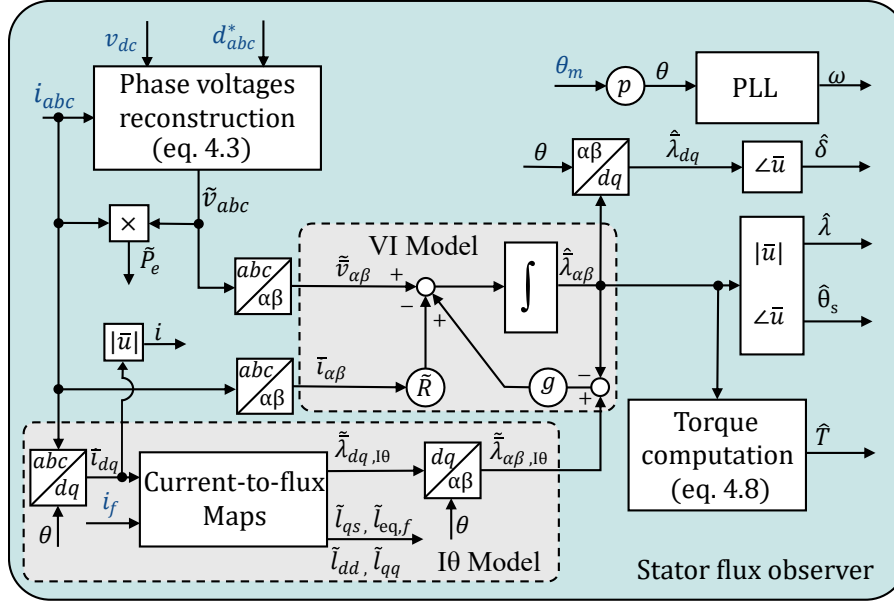


Fig. 4.3 Schematic view of the adopted stator flux observer.

The two model-based estimators are defined as follows.

Voltage model (VI)

The voltage (VI) model provides the flux estimate as the integral of the stator back-EMF voltage vector:

$$\tilde{\lambda}_{\alpha\beta,VI} = \int (\tilde{v}_{\alpha\beta} - \tilde{R} \cdot \tilde{i}_{\alpha\beta}) dt \quad (4.1)$$

where $\tilde{v}_{\alpha\beta}$ is the reconstructed voltage vector in the stationary frame, calculated from the measured inverter DC-link voltage and the command duty-cycles d_{abc}^* . Ideally, the reconstructed voltages are computed from the inverter DC-link voltage and the phase duty-cycles.

$$\begin{bmatrix} \tilde{v}_{a,id} \\ \tilde{v}_{b,id} \\ \tilde{v}_{c,id} \end{bmatrix} = v_{dc} \cdot \begin{bmatrix} d_a^* - d_0 \\ d_b^* - d_0 \\ d_c^* - d_0 \end{bmatrix}, \quad d_0 = \frac{d_a^* + d_b^* + d_c^*}{3} \quad (4.2)$$

The compensation for voltage errors introduced by the inverter $\tilde{v}_{dt,abc}$ is mandatory to improve the observer's accuracy. The voltage errors $\tilde{v}_{dt,abc}$ can be calculated by interpolating a LUT whose inputs are the measured currents i_{abc} , as described in [184]. Therefore, the corrected stator voltages $\tilde{v}_{abc}$ are estimated as:

$$\begin{bmatrix} \tilde{v}_a \\ \tilde{v}_b \\ \tilde{v}_c \end{bmatrix} = \begin{bmatrix} \tilde{v}_{a,id} - \tilde{v}_{dt,a} - \tilde{R}_{ON} \cdot i_a \\ \tilde{v}_{b,id} - \tilde{v}_{dt,b} - \tilde{R}_{ON} \cdot i_b \\ \tilde{v}_{c,id} - \tilde{v}_{dt,c} - \tilde{R}_{ON} \cdot i_c \end{bmatrix} \quad (4.3)$$

where $\tilde{R}_{ON}$ is the equivalent resistance accounting for the voltage drops on the inverter power switches.

Current model ($I\theta$)

The current ($I\theta$) model is based on the current-to-flux relationships [109]. According to Fig. 4.3, the input of the magnetic model is the stator current vector in the rotor dq frame and the rotor field current i_f . Therefore, the $I\theta$ model is immune to inverter's voltage errors, which are significant in the low-speed range.

The direct flux maps (see Fig. 2.7) and the inductance maps (see Figs. 2.27 - 2.37) of the machine are implemented with LUTs.

Observer outputs

The stator flux estimate is obtained as a linear combination between the voltage model VI and the current model $I\theta$. The crossover frequency between the two models is equal to the observer gain g , as described in the Laplace domain:

$$\hat{\lambda}_{\alpha\beta}(s) = \tilde{\lambda}_{\alpha\beta,I\theta} \cdot \frac{g}{g+s} + \tilde{\lambda}_{\alpha\beta,VI} \cdot \frac{s}{s+g} \quad (4.4)$$

where $\tilde{\lambda}_{\alpha\beta,I\theta}$ is the output of the $I\theta$ estimator, as shown in Fig. 4.3.

With reference to Fig. 4.3, the observer outputs the following estimated quantities:

- Flux amplitude

$$\hat{\lambda} = \sqrt{\hat{\lambda}_\alpha^2 + \hat{\lambda}_\beta^2} \quad (4.5)$$

- Stator flux vector position

$$\hat{\theta}_s = \text{atan2}\left(\frac{\hat{\lambda}_\beta}{\hat{\lambda}_\alpha}\right) \quad (4.6)$$

- Load angle

$$\hat{\delta} = \text{atan2}\left(\frac{\hat{\lambda}_q}{\hat{\lambda}_d}\right) \quad (4.7)$$

- Observed torque

$$\hat{T} = \frac{3}{2} \cdot p \cdot (\hat{\lambda}_\alpha \cdot i_\beta - \hat{\lambda}_\beta \cdot i_\alpha) \quad (4.8)$$

Although not directly obtained by the flux observer, the following quantities are computed for the proposed UTC:

- Estimated electric power

$$\tilde{P}_e = (\tilde{v}_a \cdot i_a + \tilde{v}_b \cdot i_b + \tilde{v}_c \cdot i_c). \quad (4.9)$$

- Current amplitude

$$i = \sqrt{i_d^2 + i_q^2}. \quad (4.10)$$

Finally, the electrical position θ and the electrical speed ω are obtained by a phase-locked loop (PLL) scheme [185] whose input is the measured rotor mechanical position θ_m .

4.1.2 Flux and currents limits computation

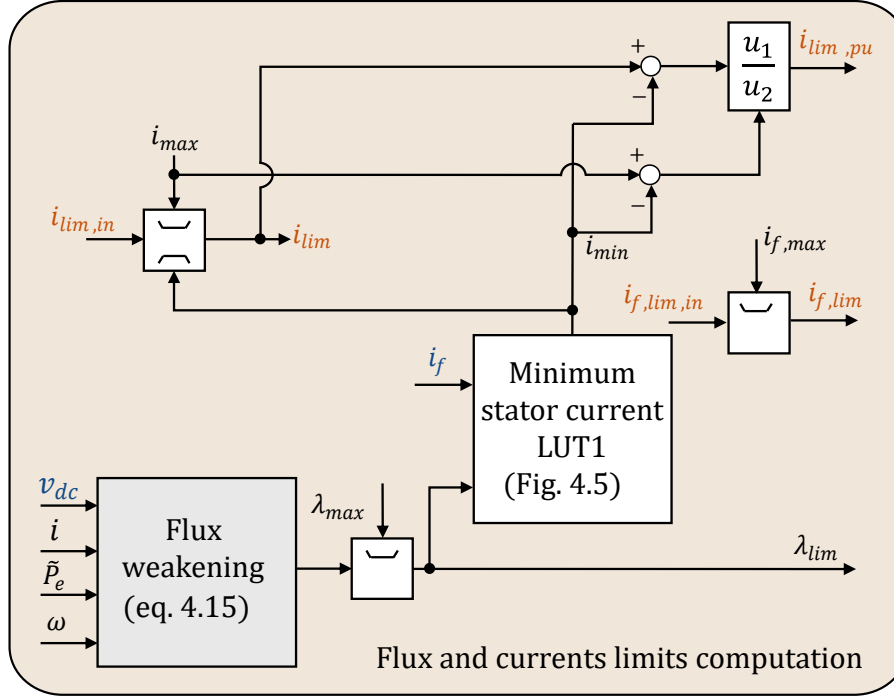


Fig. 4.4 Schematic view of flux and currents limits computation.

The torque control algorithm must comply with the constraints imposed by the considered machine and the converters (inverter and exciter) feeding it. As previously highlighted, EESMs introduce a new control variable, the rotor field current i_f . Consequently, together with the limit of stator maximum current i_{max} , the limit of maximum rotor current $i_{f,max}$ must be considered (see Fig. 4.4).

In the FW region, the flux amplitude λ must always be limited to prevent exceeding the voltage limit. As the speed increases, the flux limit λ_{lim} is determined by the inverter's maximum voltage, accounting for resistive voltage drops. The FW law can be thus retrieved recalling (2.1) and computing the square of the dq flux components as:

$$\begin{cases} \lambda_d^2 = \frac{v_d^2 + R^2 \cdot i_d^2 - 2 \cdot R \cdot i_d \cdot v_d}{\omega^2} \\ \lambda_q^2 = \frac{v_q^2 + R^2 \cdot i_q^2 - 2 \cdot R \cdot i_q \cdot v_q}{\omega^2} \end{cases} \quad (4.11)$$

Straightforwardly, the sum of the squares of the dq flux components is:

$$(\lambda_d^2 + \lambda_q^2) = \frac{(v_d^2 + v_q^2) + R^2 \cdot (i_d^2 + i_q^2) - 2 \cdot R \cdot (v_d \cdot i_d + v_q \cdot i_q)}{\omega^2} \quad (4.12)$$

Recognizing that:

$$\begin{cases} (\lambda_d^2 + \lambda_q^2) = \lambda^2 \\ (v_d^2 + v_q^2) = v^2 \\ (i_d^2 + i_q^2) = i^2 \\ \left(\frac{4}{3} \cdot \frac{3}{4}\right) \cdot 2 \cdot R \cdot (v_d \cdot i_d + v_q \cdot i_q) = \frac{4}{3} \cdot R \cdot \left[\frac{3}{2} \cdot (v_d \cdot i_d + v_q \cdot i_q)\right] = \frac{4}{3} \cdot R \cdot P_e \end{cases} \quad (4.13)$$

where P_e is the electric power, it is possible to write the formulation of the flux amplitude as:

$$\lambda = \sqrt{\frac{v^2 + (R \cdot i)^2 - (4/3) \cdot R \cdot P_e}{\omega^2}} \quad (4.14)$$

Considering that the flux limit is derived according to the inverter's maximum limit the FW law is written as follows:

$$\lambda_{lim} = k_{fw} \cdot \frac{\sqrt{v_{max}^2 + (\tilde{R} \cdot i)^2 - (4/3) \cdot \tilde{R} \cdot \tilde{P}_e}}{|\omega|} \quad (4.15)$$

In (4.15), v_{max} is the inverter voltage limit at the end of the linear PWM operation, i.e., $v_{dc}/\sqrt{3}$ since SVM-based PWM is adopted. The parameter k_{fw} is a scaling factor that compensates for inverter voltage errors and provides an adequate voltage margin to maintain high dynamic torque regulation performance, even under deep FW conditions. Typical values for k_{fw} lie within the range of 0.9 to 0.95. Obviously, λ_{lim} can never exceed the maximum possible flux value λ_{max} defined by the machine's current-to-flux relationships.

Since the rotor winding typically exhibits a significant inductance, the regulation of the rotor field current i_f might not be fast enough to comply with the flux limit λ_{lim} . This is especially true if a significant rotor acceleration is requested. Therefore, in these scenarios, the flux amplitude λ has to be reduced by exploiting the faster regulation of stator quantities. Consequently, a minimum stator current amplitude

i_{min} is necessary to counteract the flux-producing effect of the rotor field current i_f has to be defined accordingly. In the proposed torque controller, i_{min} is provided by interpolating the 2-dimensional LUT1 (see Fig. 4.4), whose inputs are the measured rotor current i_f and the flux limit λ_{lim} . To facilitate the understanding, a view of LUT1 is shown in Fig. 4.5.

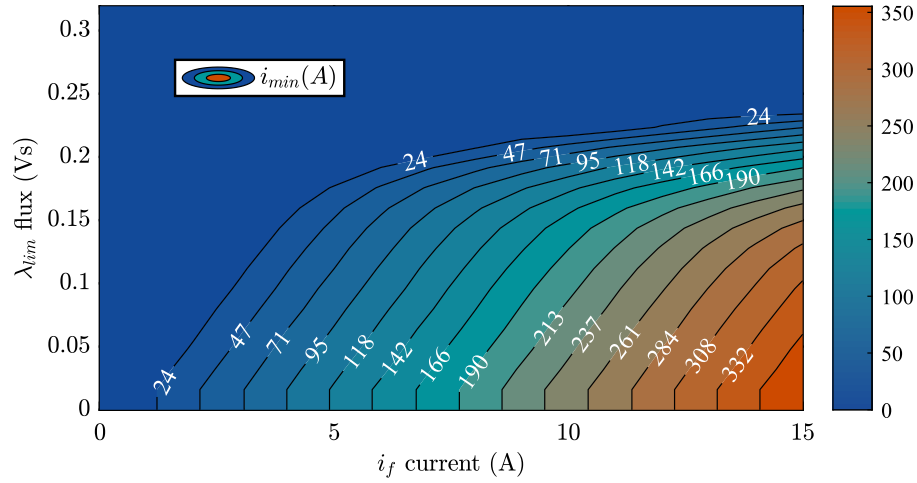


Fig. 4.5 Minimum stator current amplitude map (LUT1) for the considered EESM motor.

In this work, the maximum limits of stator current amplitude i_{max} and rotor current $i_{f,max}$ correspond to the most critical overload conditions of the EESM under consideration. However, these limits cannot always be considered as the effective ones during the machine's operation. Indeed, according to the literature [146], adaptive limitation of the stator and rotor current limits is mandatory to protect windings from excessive temperature levels. Therefore, an efficient derating strategy of the current limits ($i_{lim}, i_{f,lim}$) is integrated into the proposed torque controller to allow the implementation of thermal strategies reported in the literature [146–148]. According to Fig. 4.4, the effective stator current amplitude limit i_{lim} is thus obtained from the input target $i_{lim,in}$ after applying the constraints of minimum i_{min} and maximum i_{max} current amplitude. Indeed, regardless of the thermal strategy to implement, the lower limit of stator current amplitude i_{min} must be respected to guarantee the safe operability of the EESM. At the same time, the effective rotor current limit $i_{f,lim}$ is obtained from the input target $i_{f,lim,in}$ after applying the obvious constraint of maximum rotor current $i_{f,max}$. In summary, flux and currents limitations performed in Fig. 4.4 corresponds to the following constraints:

$$\begin{cases} i \leq i_{lim} & ; & i_{min} \leq i_{lim} \leq i_{max} \\ i_f \leq i_{f,lim}; & & i_{f,lim} \leq i_{f,max} \\ \lambda \leq \lambda_{lim} \end{cases} \quad (4.16)$$

4.1.3 Torque limits and torque references computation

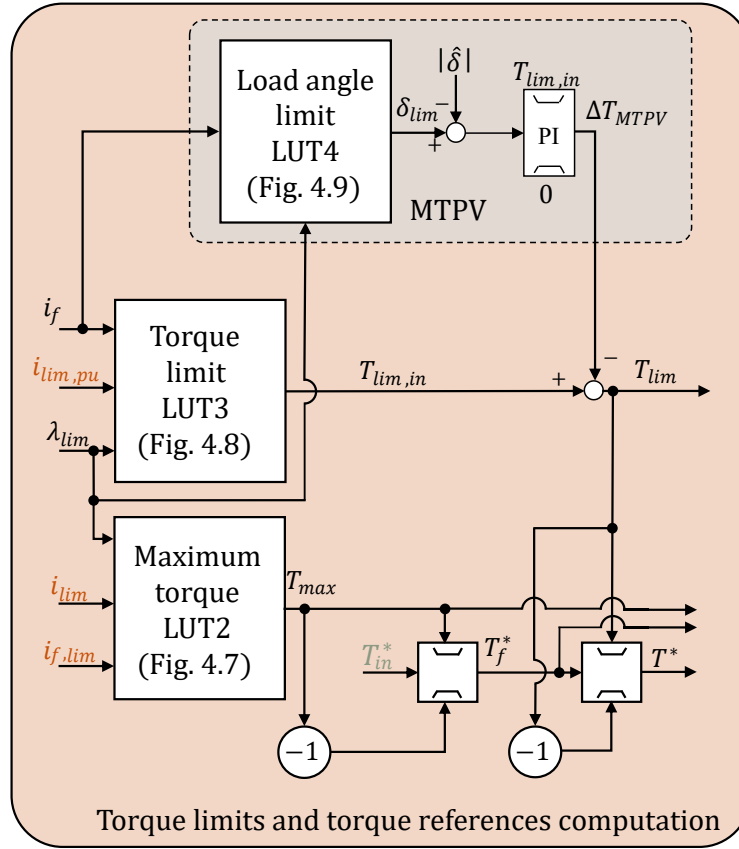


Fig. 4.6 Schematic view of the torque limits and torque references computation.

As previously pointed out, the primary drawback regarding control of traction EESMs is the exciter dynamics. Indeed, it is well known that the dynamics of the rotor current are relented by the large rotor inductance compared to the stator dq ones. Even in the high-speed region, where the stator current control dynamic is limited by the back-emf, with the same DC-link voltage, the rotor current dynamic is still way lower than the stator dq ones. This difference between the dynamics of stator and rotor currents hinders the dynamic performance of torque regulation. Therefore, one of

the main contributions of the proposed torque controller is the definition of a control structure that achieves high dynamic performance in torque regulation despite of the exciter's slow dynamic response. For this purpose, two different torque references, one for the stator regulation T^* and one for the rotor regulation T_f^* are defined (see Fig. 4.6).

While the implementation of a dual-torque reference increases the architectural complexity of the proposed controller, this approach is justified by the significant bandwidth separation and time-constant mismatch between the stator and rotor circuits. In standard traction cycles characterized by moderate torque gradients, a unified reference may suffice. However, the dual-reference scheme is essential for EESM drives requiring high-dynamic performance, such as anti-panic braking or high-end automotive applications, to ensure transient stability. The implications of utilizing a unified torque reference are analyzed through two critical scenarios:

- If T^* is utilized to derive both stator and rotor references, a recursive dependency occurs. Since the feasible stator torque reference T^* is inherently constrained by the instantaneous rotor current i_f , the rotor current reference i_f^* becomes limited by its own state. This "algebraic loop" significantly derates the rotor current dynamics, resulting in a sluggish torque response.
- Alternatively, if the rotor torque reference T_f^* is utilized as a unified reference, the rotor loop operates normally, but significant damping issues arise in the stator control path whenever the actual rotor current i_f deviates from its reference i_f^* . During fast transients, the stator references (derived from the target rather than the actual state) fail to compensate for the armature reaction, leading to underdamped torque oscillations.

The criticality of these scenarios is governed by the rotor time constant and the requested torque slew rate. To mitigate these effects, the proposed dual-torque-reference mechanism decouples the regulation paths.

Torque reference for rotor current regulation

Starting from the generation of the torque reference for the rotor regulation T_f^* , the maximum torque that the EESM under consideration can deliver is first computed

according to the limits of flux amplitude λ_{lim} , rotor current $i_{f,lim}$, and stator current amplitude i_{lim} . The maximum torque T_{max} is thus obtained by interpolating the 3-dimensional LUT2, having the above limits as inputs. To facilitate the understanding, a view of LUT2 for different values of rotor current limit $i_{f,lim}$ is shown in Fig. 4.7. Therefore, the torque for the rotor regulation is obtained by limiting the input target torque T_{in}^* to T_{max} as:

$$|T_f^*| = \min(|T_{in}^*|, |T_{max}|). \quad (4.17)$$

It is highlighted that this torque limitation primarily reflects the limitations of the stator-side by including λ_{lim} , and does not explicitly incorporate the dynamics of the rotor current i_f .

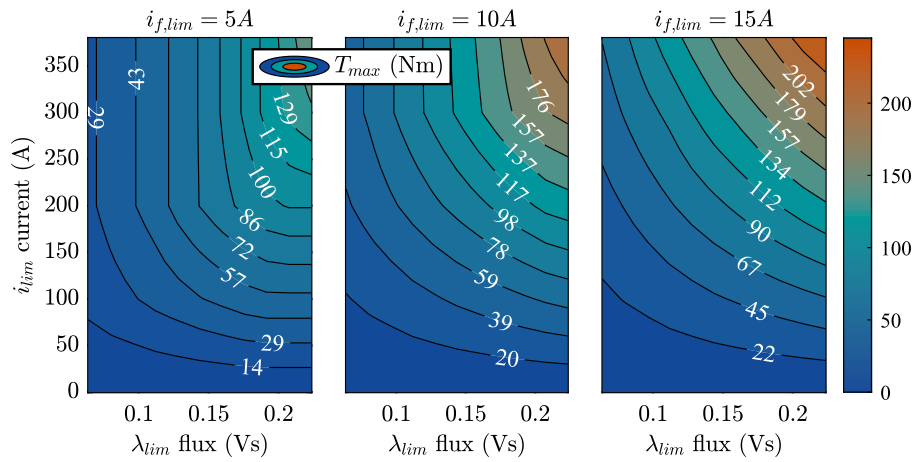


Fig. 4.7 Maximum torque map (LUT2) for the considered EESM motor.

Torque reference for regulation of stator quantities

The generation of the torque reference for the stator regulation T^* considers the limitations introduced by the measured value of the rotor current i_f and the MTPV limit, i.e., the load angle limitation δ_{lim} , avoiding the EESM pull-out.

Torque limitation due to rotor current measurement: According to Fig. 4.6, the maximum achievable torque $T_{lim,in}$ is first computed based on the stator current and flux amplitude limits (i_{lim}, λ_{lim}) together with the measurement of the rotor current i_f . This torque limit is obtained by interpolating the 3-dimensional LUT3 having the aforementioned variables as inputs. However, the stator current limit i_{lim} is, in

turn, correlated with the values of the measured rotor current i_f and flux amplitude limit λ_{lim} , according to the minimum stator current i_{min} , as already discussed and shown in Fig. 4.4. This mutual correlation between the input variables of LUT3 results in a highly irregular map domain, which poses challenges for implementing and interpolating these LUT on digital controllers.

To efficiently address this issue, the method proposed in [80] is adopted. This technique enables the construction of multi-dimensional LUTs by systematically managing the interdependencies among multiple input variables, which is essential when dealing with strongly coupled nonlinear systems such as EESMs. A brief overview of the employed method is provided below. In the case where one or more input directions of the LUT are dependent on others, a normalization procedure is applied. Specifically, the dependent input dimensions are transformed into a per-unit (pu) scale ranging from 0 to 1. In this normalized domain:

- A value of 0 corresponds to the minimum physically admissible value of the variable for the given values of the remaining inputs.
- A value of 1 corresponds to its maximum admissible value of the variable for the given values of the remaining inputs.

This normalization ensures that the LUT can be interpolated with linear methods while preserving the integrity of the original nonlinear relationships. Therefore, for the implementation of LUT3, i_{lim} is normalized coherently with the value of i_{max} and $i_{min}(i_f, \lambda_{lim})$ as (see Fig. 4.4):

$$i_{lim,pu} = \frac{i_{lim} - i_{min}}{i_{max} - i_{min}} \quad (4.18)$$

This normalization enables the use of linear multi-dimensional interpolation methods, facilitating efficient implementation on digital controllers. The graphical representation of the map's regularization process is shown in Fig. 4.8.

Implementing LUT3 allows for obtaining a torque limit $T_{lim,in}$ that can always be satisfied by the torque controller. Besides, this torque limit accounts for the possibility of temporarily overloading the stator windings up to i_{lim} while waiting for the rotor current to achieve its reference. This way, significant variations of the torque request can be quickly satisfied regardless of the potential slow dynamic of the rotor current.

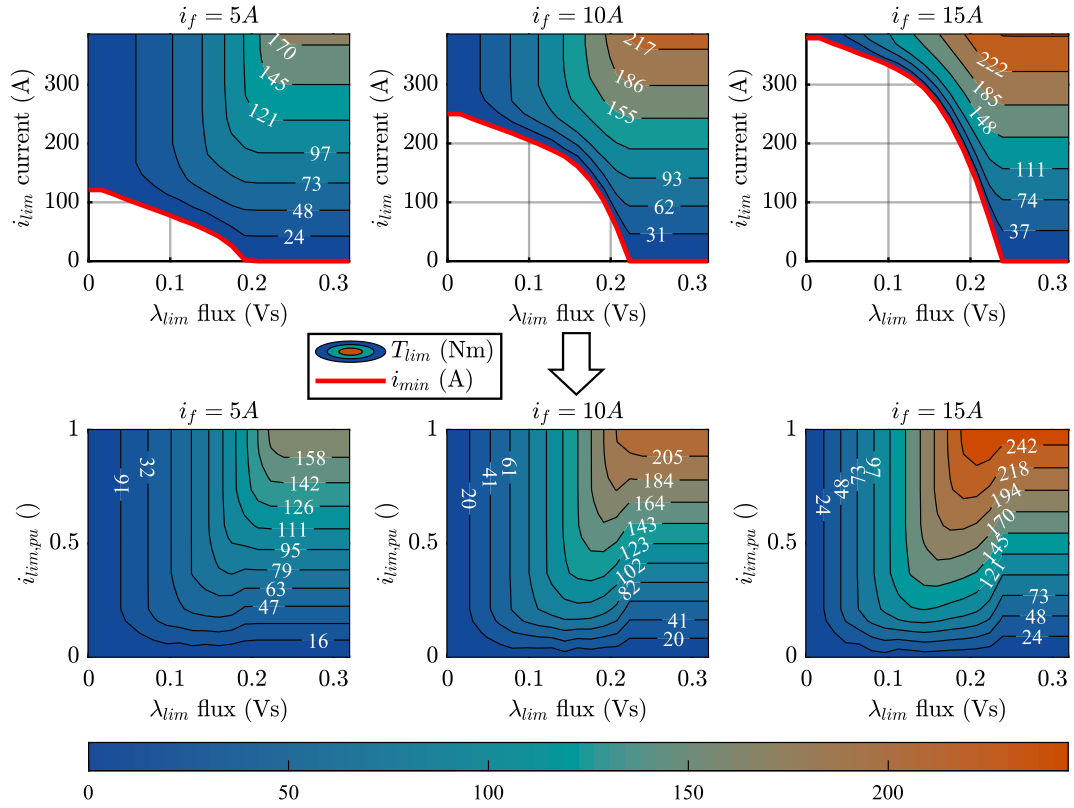


Fig. 4.8 Graphical view of the torque limit map (LUT3) transformation process for the considered EESM motor.

Torque limitation due to MTPV: The MTPV operation occurs in deep FW operation if the stator current amplitude limit i_{lim} overcomes the EESM's characteristic current. Using a DFVC or DTC scheme, applying the MTPV limitation corresponds to appropriately limiting the machine's load angle δ [109]. Besides, this limitation is pivotal in DFVC and DTC schemes since overcoming it may lead to instability of torque regulation due to the machine's pull-out. Although the previously computed torque limit $T_{lim,in}$ automatically accounts for the MTPV operation, the DFVC and DTC schemes need an active limitation of the machine's load angle by using an outer regulator that properly limits the torque reference [109], justifying what is shown in Fig. 4.6.

It is noted that an outer PI regulator is implemented to ensure that the estimated load angle $\hat{\delta}$ follows its limit value δ_{lim} when torque regulation would otherwise result in $\hat{\delta} > \delta_{lim}$, leading to instability. Both the limit δ_{lim} and the estimated $\hat{\delta}$ load angle must be assumed as absolute values to manage negative torque operations

straightforwardly. This critical load angle δ_{lim} (see Fig. 4.9) can be determined by computing the MTPV limit from the current-to-flux maps.

Regarding the calibration of this PI regulator, heuristic procedures are typically adopted starting from approximated mathematical approaches [109]. However, regardless of the adopted calibration method, the output of this regulator ΔT_{MTPV} cannot overcome the torque limit $T_{lim,in}$ previously computed. Indeed, by summing $T_{lim,in}$ and ΔT_{MTPV} , the absolute torque limit T_{lim} is computed:

$$T_{lim} = T_{lim,in} - \Delta T_{MTPV}, \quad \text{with} \quad \Delta T_{MTPV} \leq T_{lim,in}. \quad (4.19)$$

This way, the torque reference T^* for regulating the stator quantities is obtained, as shown in Fig. 4.4 as:

$$|T^*| = \min(|T_{in}^*|, |T_{lim}|). \quad (4.20)$$

In the proposed UTC scheme, the load angle limit δ_{lim} is obtained by interpolating the 2-dimensional LUT4, whose inputs are the flux amplitude limit λ_{lim} and the measured rotor current i_f . A view of the load angle limit map is shown in Fig. 4.9.

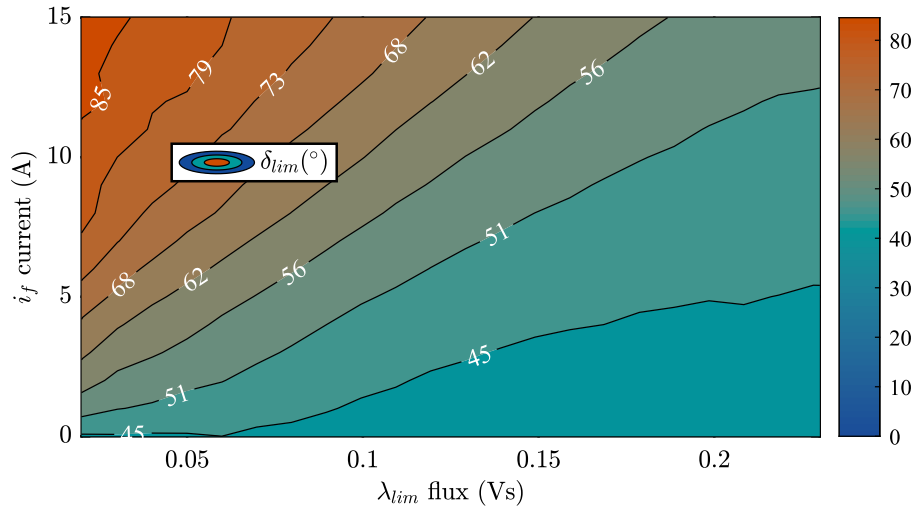


Fig. 4.9 Load angle limit map (LUT4) for the considered EESM motor.

4.1.4 Control references computation

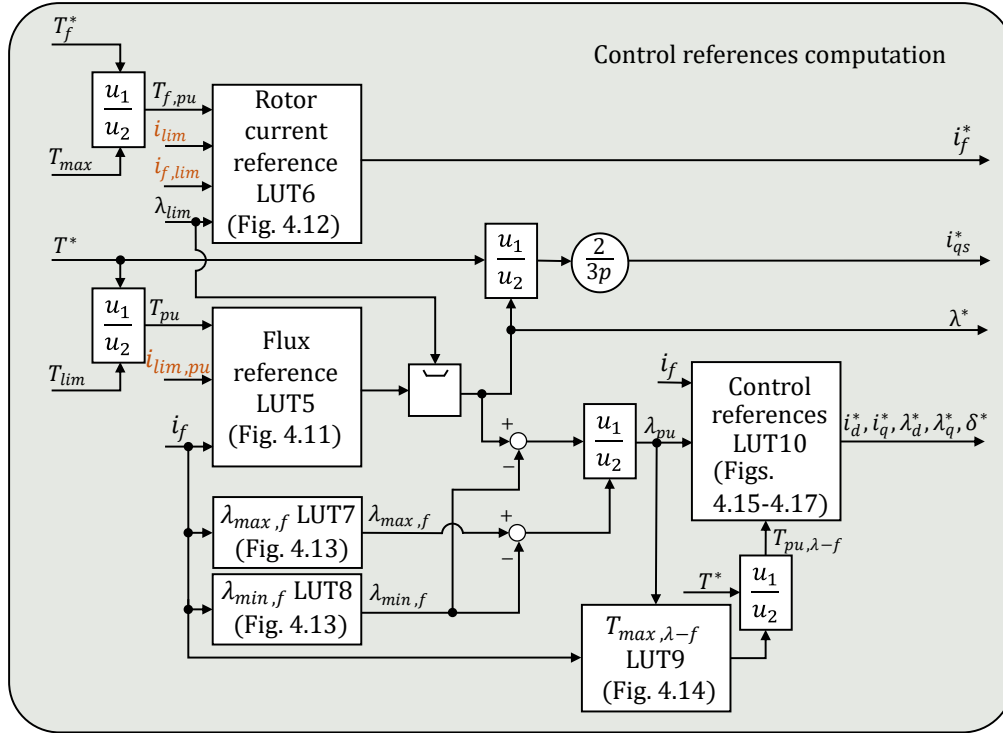


Fig. 4.10 Schematic view of the control references computation.

Based on the stator torque reference T^* , the complete set of control references required for each TCA is systematically derived, as illustrated in Fig. 4.10. These quantities include the optimal stator flux amplitude λ^* , the stator current components (i_d^*, i_q^*) for CVC, the stator flux components (λ_d^*, λ_q^*) for FVC, the load angle δ^* for FPC, and the torque-producing current i_{qs}^* for DFVC. Concurrently, the rotor current reference i_f^* is computed based on the defined rotor torque reference T_f^* .

It is noted that the EESM is characterized by an intrinsic redundancy, wherein a specific torque output can be synthesized through multiple combinations of stator $(\bar{\lambda}, \bar{i})$ and rotor (i_f) state variables. This degree of freedom arises because torque generation results from the interaction between stator and rotor quantities. Consequently, rigorous torque regulation requires accounting for the combined flux contribution of both windings. Furthermore, to enhance dynamic performance and operational flexibility, it is advantageous to decouple the flux and torque control loops. This structural separation allows the independent optimization of the flux amplitude regardless of the instantaneous torque demand.

Stator flux amplitude and rotor current references computation

Regardless of the specific employed TCA, the optimal stator flux amplitude λ^* is computed to strictly satisfy the voltage and current constraints defined in (4.16). This ensures that the generated control references remain within the machine's safe thermal and electrical operating range. While the control structure can accommodate various efficiency strategies, this work derives the optimal references for stator flux amplitude λ^* and rotor current i_f^* by solving a constrained optimization problem minimizing total Joule losses:

$$\min_{\lambda, i_f} \left\{ \frac{3}{2} \cdot R \cdot i^2 + R_f \cdot i_f^2 \right\} \quad (4.21)$$

subject to the constraints in (4.16).

The real-time analytical resolution of (4.21) is computationally heavy due to the nonlinear magnetic topology of the EESM, which is characterized by deep saturation and significant stator-rotor cross-coupling. To address this issue, the optimization is performed offline, elaborating the direct flux maps (Fig. 2.7). This process yields two pre-computed LUTs:

- **LUT5:** Outputs the optimal stator flux amplitude reference λ^* .
- **LUT6:** Outputs the optimal rotor current reference i_f^* .

A critical challenge in the computation of these control LUTs is the dependence of torque capability on variable system constraints. The rotor torque reference T_f^* is a function of the flux limit and current constraints $(\lambda_{lim}, i_{lim}, i_{f,lim})$, while the stator torque reference T^* depends on the instantaneous rotor current i_f and the stator current limit i_{lim} . Notably, the stator current limit is coupled to the rotor current via the minimum current trajectory i_{min} , as illustrated in Fig. 4.5.

To manage these interdependencies, the previously introduced linearization procedure that generalizes the normalization technique proposed in [80] for PMSMs is applied. The stator current limit is normalized in pu form relative to the instantaneous rotor current value according to the minimum stator current i_{min} profile, as depicted in (4.18) (see Fig. 4.4).

Similarly, the linear interpolation of LUT5 and LUT6 utilizes torque references normalized with respect to their instantaneous maximum capabilities (see Fig. 4.6):

$$T_{pu} = \frac{T^*}{T_{lim}}; \quad T_{f,pu} = \frac{T_f^*}{T_{max}} \quad (4.22)$$

The resulting map for the flux amplitude reference LUT5, plotted considering varying rotor current and stator current limit, is shown in Fig. 4.11. Correspondingly, the rotor current reference map LUT6, considering different stator flux and rotor and stator current limits, is presented in Fig. 4.12.

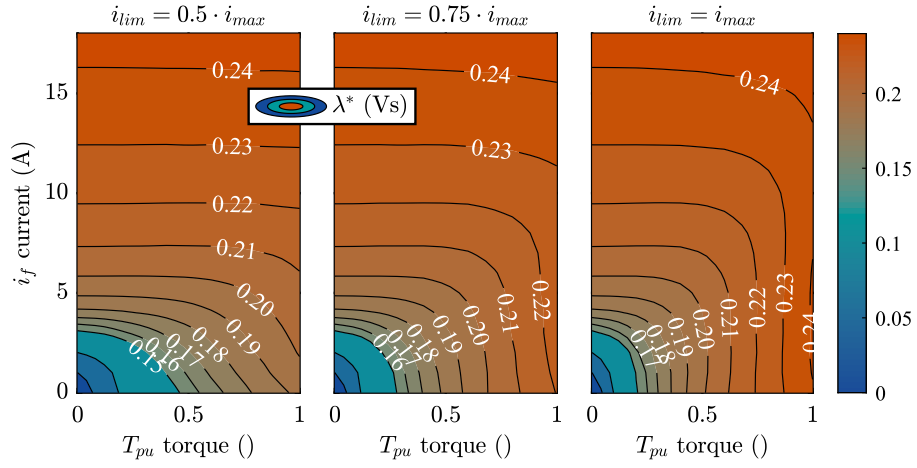


Fig. 4.11 Flux amplitude reference (λ^*) map (LUT5) for different stator current limits and rotor current for the considered EESM motor.

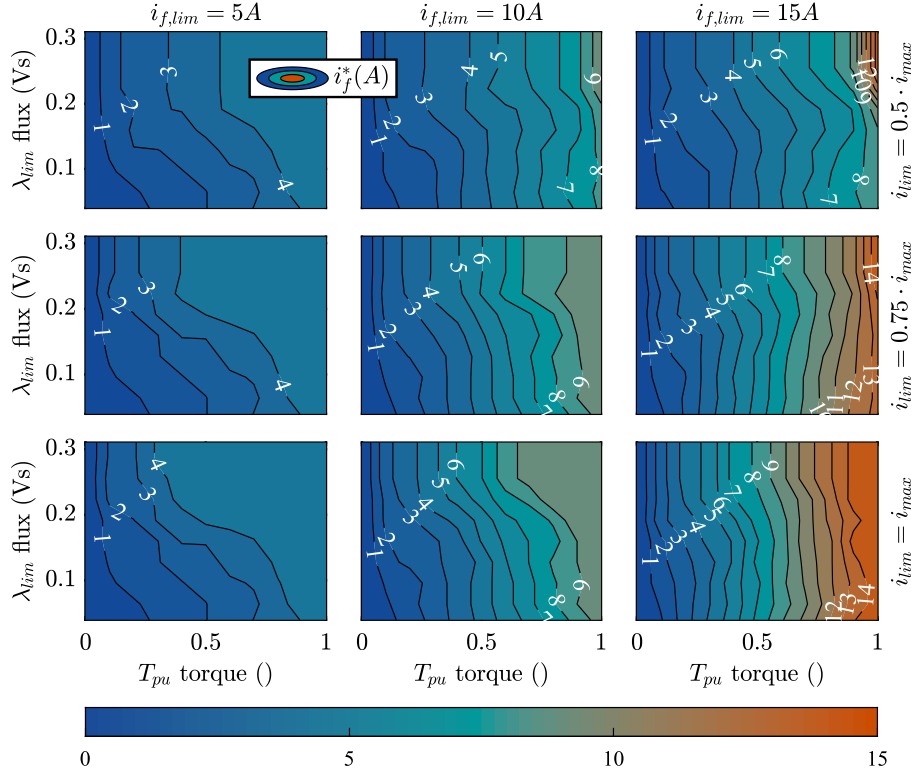


Fig. 4.12 Rotor current reference (i_f^*) map (LUT6) for different stator and rotor current limits for the considered EESM motor.

For simplicity, the optimization formulation in (4.21) currently employs nominal resistance values of stator and rotor resistances. However, the framework is intrinsically extensible; thermal dependencies can be rigorously incorporated by augmenting the dimension of the optimization LUTs (LUT5 and LUT6) to include stator and rotor winding temperatures as interpolation variables. Furthermore, the proposed control architecture (see Fig. 4.2) exhibits structural invariance regarding the specific optimization objective. Consequently, the LUT5 and LUT6 can be reformulated to satisfy alternative cost functions such as the minimization of combined Joule and iron losses without necessitating topological modifications to the control scheme.

Computation of TCA references

As illustrated in Fig. 4.10, the reference variables for any TCA (i_d^* , i_q^* , λ_d^* , λ_q^* , δ^*) are derived from the instantaneous rotor current i_f and the reference torque and flux commands (T^* , λ^*). With the exception of the torque-producing current i_{qs}^* , which is

algebraically determined via (2.11), these quantities are governed by nonlinear magnetic relationships, necessitating the utilization of LUT10. The direct implementation of LUT10 on digital platforms is non-trivial due to the complex, coupled dependency of both torque capability and flux magnitude on the rotor current i_f . Consequently, a primary contribution of this work is the development of a linearization methodology for EESM torque regulation. This approach is agnostic to the selected TCA and generalizes the normalization procedure originally proposed for PMSMs in [80] as described before.

The linearization process begins with the normalization of the stator flux amplitude inpu form, relative to the instantaneous rotor current. Specifically, the feasible flux range defined by lower and upper bounds ($\lambda_{\min,f}$ and $\lambda_{\max,f}$) is extracted from LUT7 and LUT8, respectively (see Fig. 4.13). The normalized flux is subsequently computed as:

$$\lambda_{\text{pu}} = \frac{\lambda^* - \lambda_{\min,f}}{\lambda_{\max,f} - \lambda_{\min,f}} \quad (4.23)$$

Notably, while $\lambda_{\min,f}$ is negligible for the specific EESM under test, the formulation retains this term to ensure the generalizability of the control structure to machine topologies exhibiting significant minimum flux characteristics.

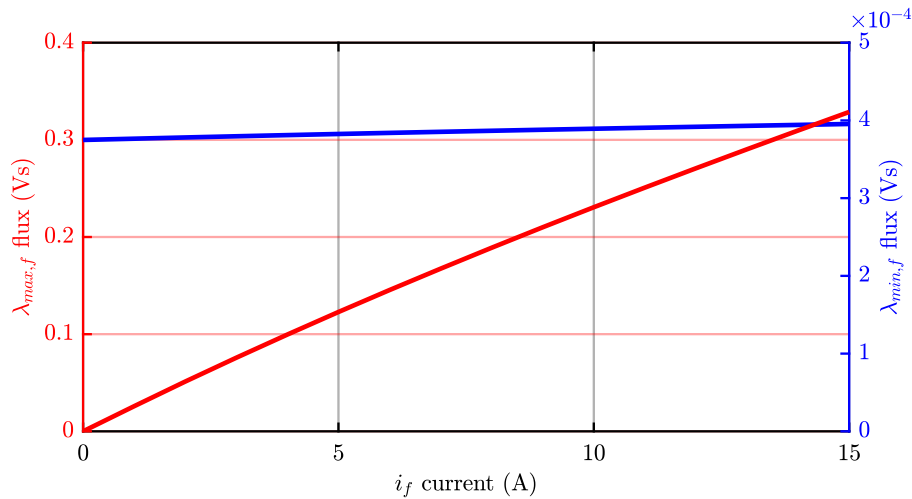


Fig. 4.13 Maximum and minimum flux as a function of the rotor current (LUT7 and LUT8) for the EESM under consideration.

Subsequently, a maximum torque $T_{\max,\lambda-f}$ function of the rotor current i_f and the stator flux λ_{pu} is defined and stored in LUT9 (see Fig. 4.14). A new pu reference

torque is thus retrieved as:

$$T_{pu,\lambda-f} = \frac{T^*}{T_{max,\lambda-f}}. \quad (4.24)$$

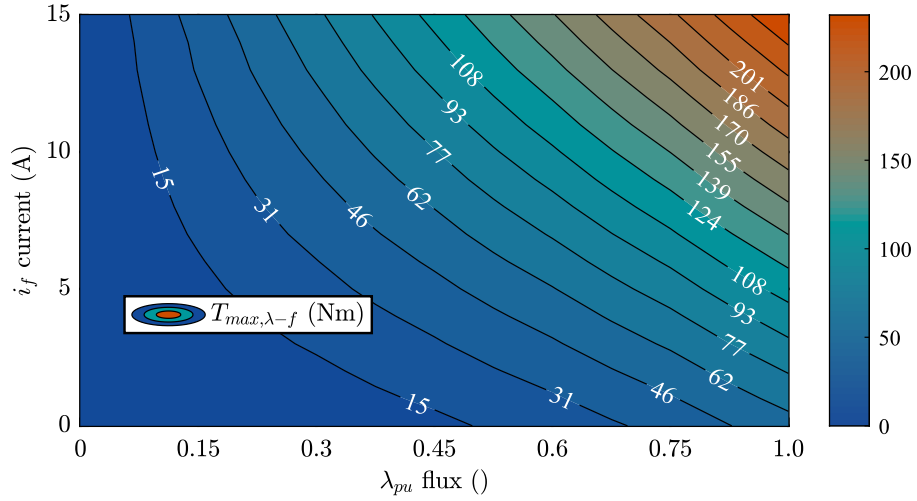


Fig. 4.14 Maximum torque surface as a function of the rotor current and the stator flux amplitude (LUT9) for the EESM under consideration.

Finally, the control references are derived through the linear interpolation of LUT10, which is parametrized by $(\lambda_{pu}, T_{pu,\lambda-f}, i_f)$. It is noteworthy that the LUT10 output configuration is reconfigurable to accommodate the specific requirements of the selected TCA. Conversely, in DFVC and DTC architectures, this LUT is bypassed, as the control variables are derived directly (see Figs. 4.15-4.17).

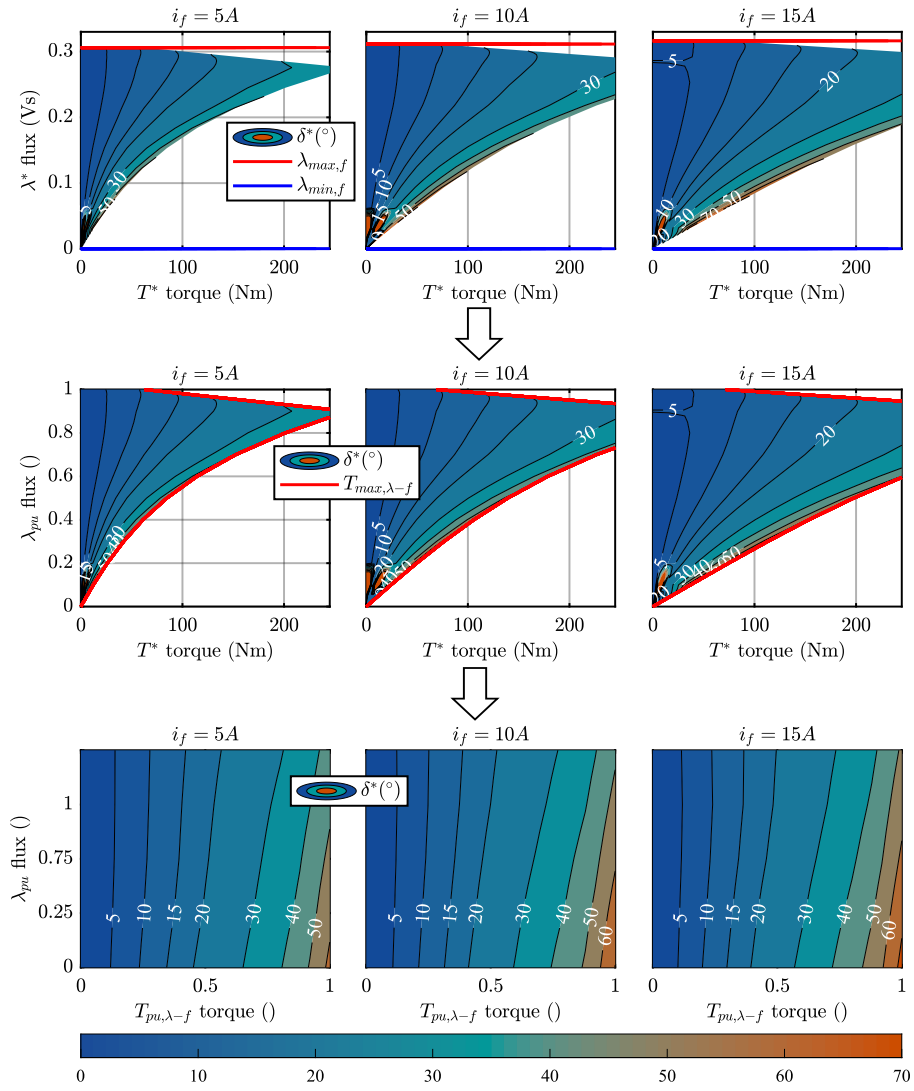


Fig. 4.15 Load angle reference (δ^*) normalization (LUT10) for the EESM under consideration.

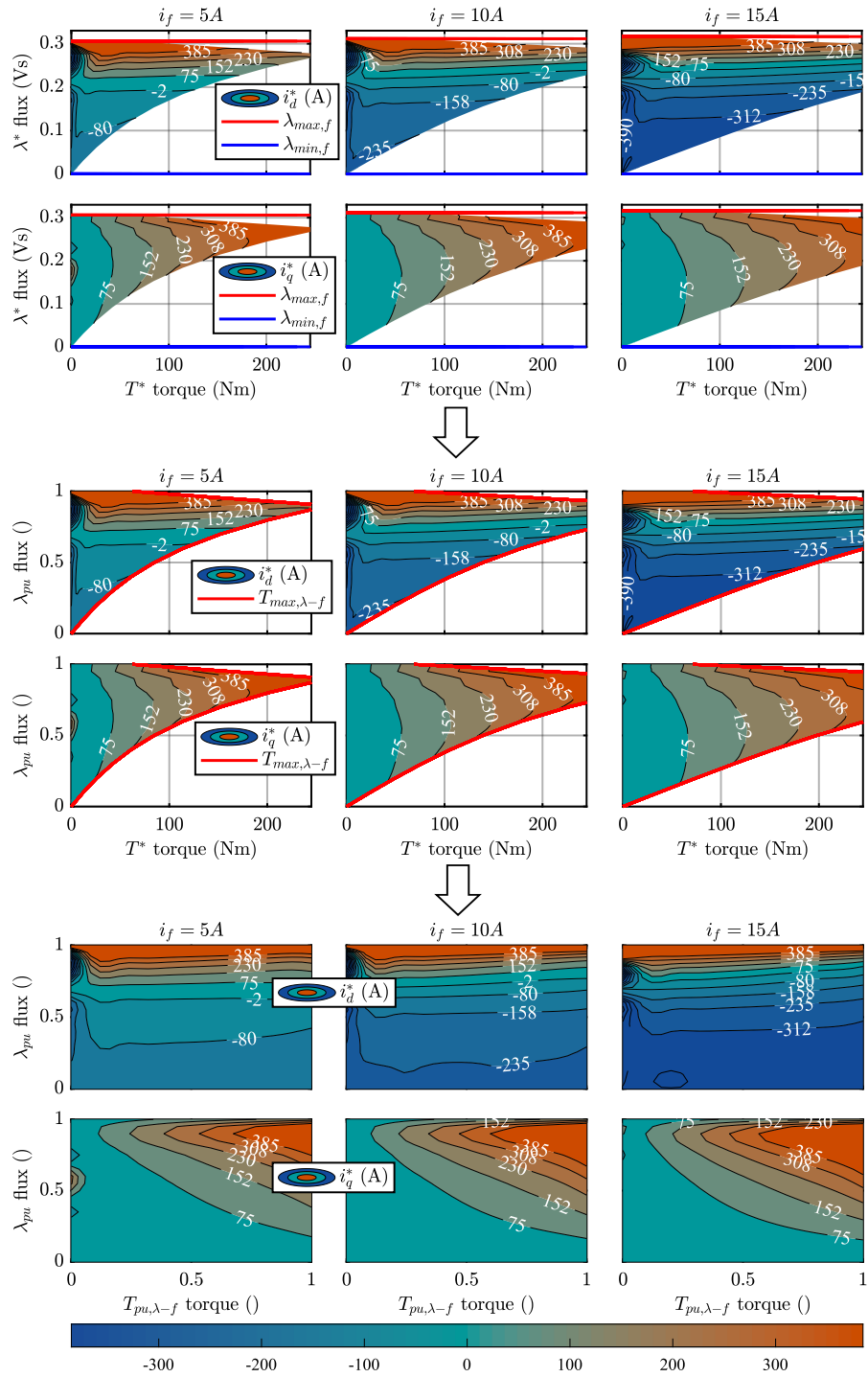


Fig. 4.16 dq currents references (i_d^* , i_q^*) normalization (LUT10) for the EESM under consideration.

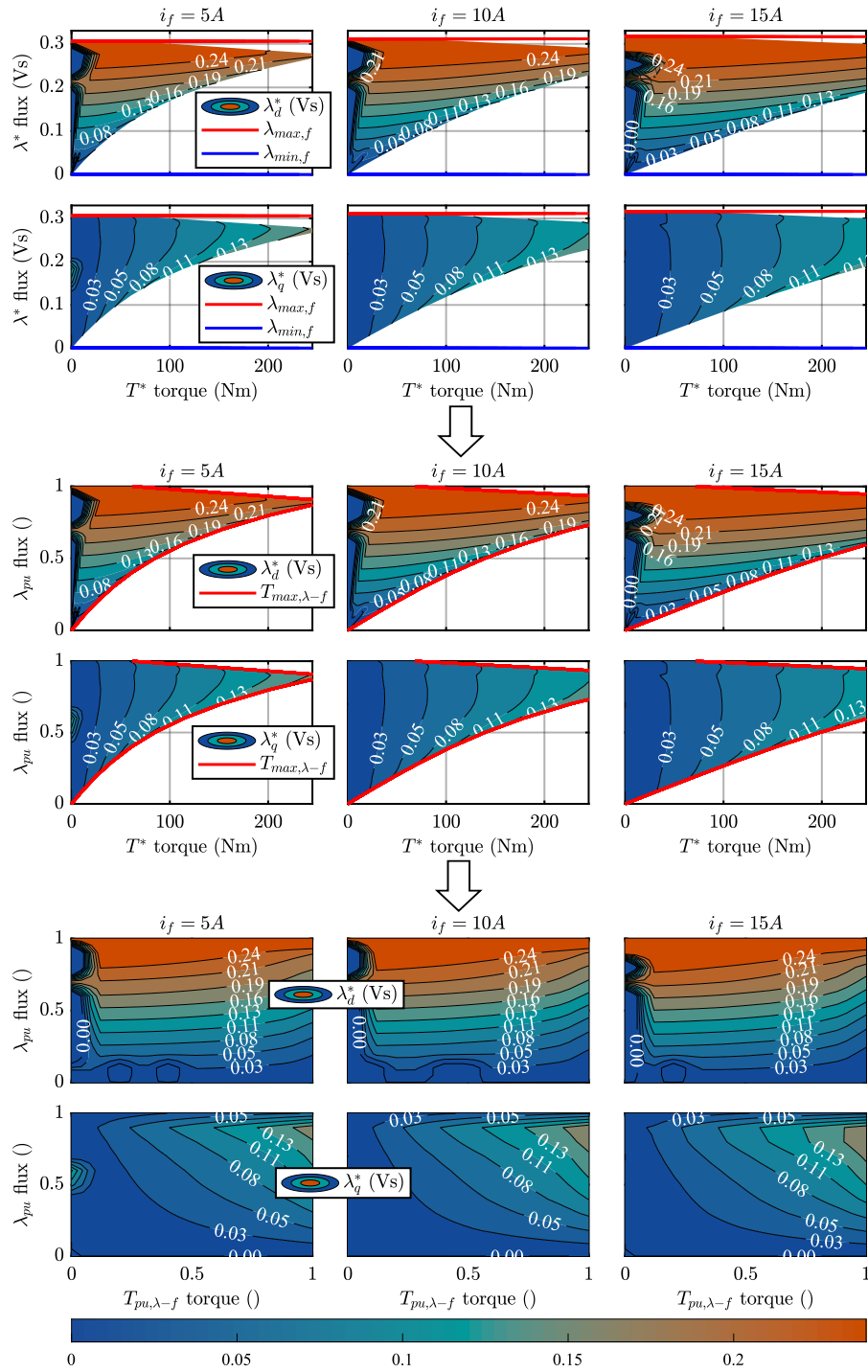


Fig. 4.17 dq stator fluxes references (λ_d^* , λ_q^*) normalization (LUT10) for the EESM under consideration.

In synthesis, the control architecture utilizes LUT5 and LUT6 to determine the optimal steady-state trajectories for the stator flux magnitude λ^* and the rotor current i_f^* according to two different torque references (T^*, T_f^*) , strictly adhering to the selected efficiency criterion (here, copper loss minimization). This topology effectively decouples the stator and rotor regulation loops, thereby ensuring that the global torque bandwidth is constrained solely by the stator electro-dynamics and the operational limits defined in (4.16). Subsequently, LUT10 operates as a mapping layer, translating these optimized setpoints into the specific reference frame required by the active TCA.

It is highlighted that the proposed control architecture exhibits inherent robustness against thermal variations. Since EESMs are magnetless, the current-to-flux and current-to-torque characteristics remain invariant to temperature changes; consequently, LUT1-LUT4 and LUT7-LUT10 are independent of the machine's thermal state. In contrast, LUT5 and LUT6 are derived based on the minimization of global Joule losses, utilizing the rated temperatures of the stator and rotor windings. To maintain the optimal efficiency condition, these two LUTs should be adapted according to measured or estimated winding temperatures. Nevertheless, torque regulation accuracy is strictly preserved regardless of thermal drift, as the torque generation logic relies exclusively on the temperature-invariant control references defined in LUT10.

4.1.5 Torque control approach (TCA)

The proposed UTC can implement multiple TCAs, each of them being chosen based on specific application requirements and the EESM design. Depending on the selected TCA, different regulators' characteristics have been previously highlighted in Chapter 2, and hereafter further analyzed.

Stator dqs -axes TCAs

In DFVC, DTC, and FPC architectures, torque control is implemented using two regulators operating in the stator-flux-oriented dqs reference frame. As established in Chapter 2, the ds -axis flux regulator is primarily responsible for compensating resistive voltage drops and inverter non-ideal effects. Consequently, the voltage saturation limit for the flux loop, $v_{ds,lim}$, requires minimal voltage amplitude and is

empirically defined as:

$$2 \cdot R \cdot i_{\max} \leq v_{ds,\lim} \leq 5 \cdot R \cdot i_{\max} \quad (4.25)$$

This allocation strategy maximizes the voltage available to the qs -axis regulator, which controls the torque-producing variables $(i_{qs}, \hat{T}, \hat{\delta})$. This decision ensures superior dynamic performance, particularly during deep flux-weakening operation where voltage disposal is critical. To further enhance the bandwidth of the torque loop, feed-forward compensation of the back-emf $(\omega \cdot \hat{\lambda})$ is implemented. Analysis of (2.12) indicates that the plant dynamics of the flux control loop are invariant with respect to the operating point. Consequently, if a PI controller is employed, the gains (k_p, k_i) do not require adaptive tuning and can be tuned according to the design criteria proposed in [80]. A similar property of invariance applies to the load angle regulator in FPC, as derived from (2.18).

In contrast, the control loops for torque $\hat{T}$ and torque-producing current i_{qs} are governed by the differential inductance l_{qs} , as defined in (2.16) and (2.17). Since l_{qs} is heavily dependent on magnetic saturation (see Fig. 2.37), the gains of the qs -axis regulator require dynamic adaptation. This gain scheduling relies on the real-time estimation of the differential inductance $\tilde{l}_{qs}$, extracted from the magnetic model within the flux observer (see Fig. 4.3), to guarantee stability and damp oscillations across the operating range.

Direct Flux Vector Control: In DFVC scheme, regulation of the stator flux amplitude λ and the torque-producing current i_{qs} is achieved via two regulators operating within the stator flux-oriented dqs reference frame (Fig. 4.18). The resulting voltage commands (v_{ds}^*, v_{qs}^*) are transformed into the phase domain v_{abc}^* via the Park transformation, utilizing the estimated stator flux position $\hat{\theta}_s$ for orientation.

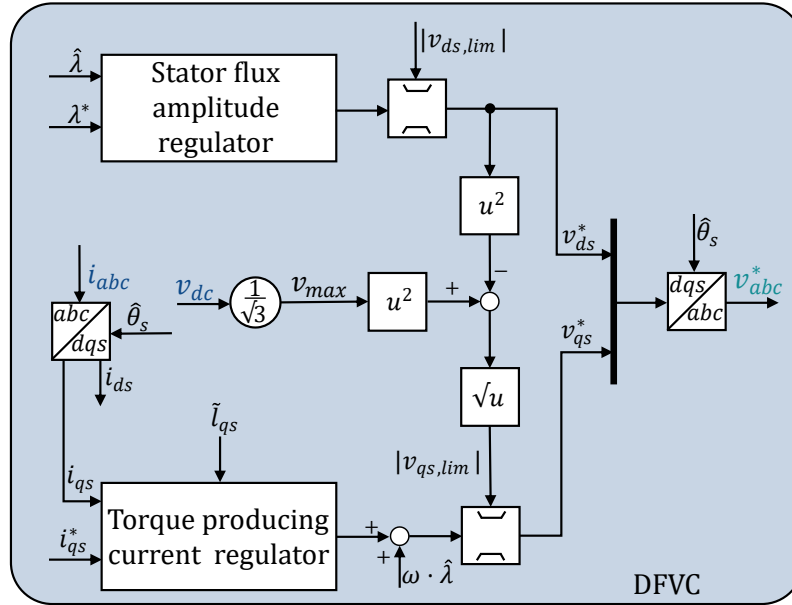


Fig. 4.18 Schematic of the DFVC architecture: closed-loop regulation of stator flux amplitude λ and i_{qs} current.

Direct Torque Control: Analogous to DFVC, the DTC strategy employs two regulators in the dqs frame to control the stator flux amplitude λ and the electromagnetic torque $\hat{T}$. To linearize the control plant, the output of the torque regulator is scaled by the inverse of the observed stator flux amplitude $\hat{\lambda}$, as illustrated in Fig. 4.19. Consequently, consistent with (2.17) and (2.16), the tuning of the torque regulator becomes methodologically equivalent to that of the torque-producing current loop in the DFVC scheme.

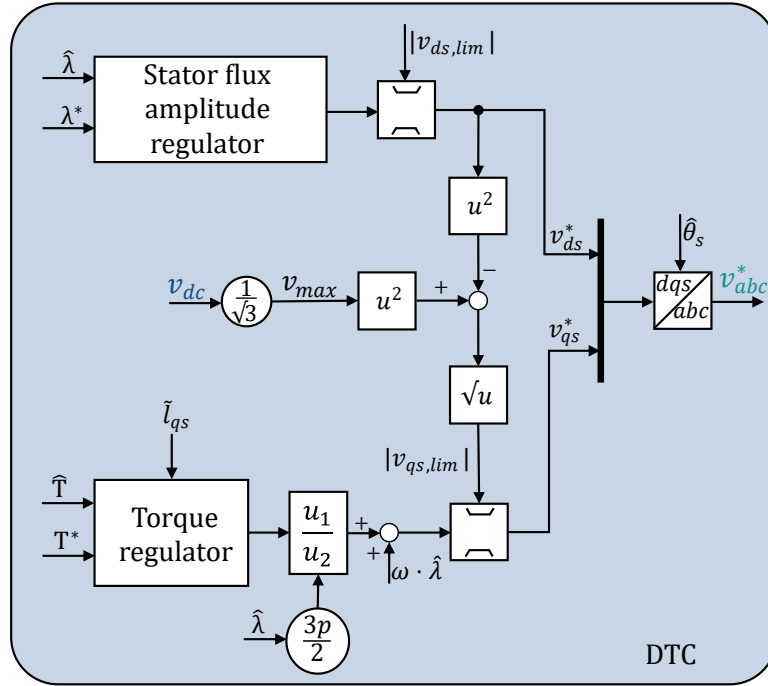


Fig. 4.19 Schematic of the DTC architecture: closed-loop regulation of stator flux amplitude λ and electromagnetic torque T .

Flux Polar Control: FPC regulates the stator flux vector in polar coordinates, controlling the amplitude λ and the load angle $\hat{\delta}$ via the dqs frame. As depicted in Fig. 4.20, the output of the load angle regulator is scaled by the stator flux amplitude. According to (2.18), this normalization renders the load angle dynamics invariant with respect to the machine operating point. Consequently, the controller tuning follows the same linear design criteria employed for the flux amplitude loop described in the following.

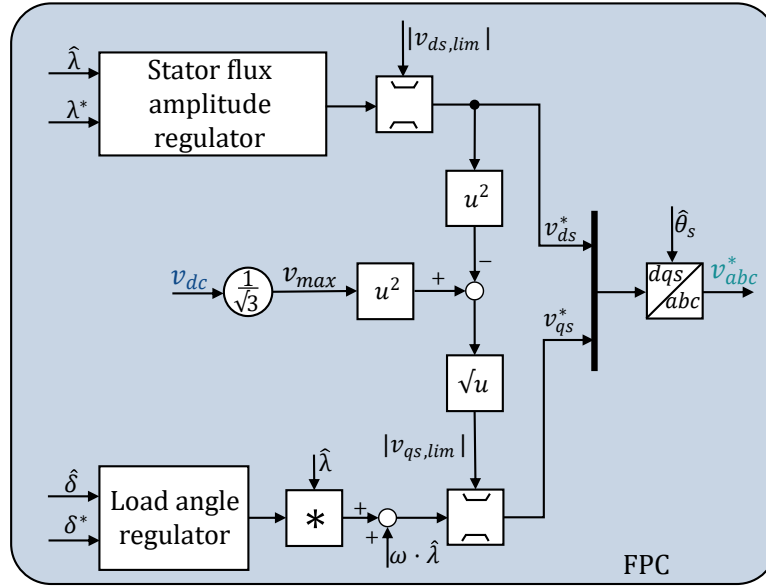


Fig. 4.20 Schematic of the FPC architecture: closed-loop regulation of stator flux amplitude λ and load angle δ .

Rotor dq -axes TCAs

The CVC and FVC schemes operate in the rotor synchronous dq reference frame. Both strategies necessitate the decoupling of the cross-coupling back-emf terms ($-\omega\lambda_q$ and $\omega\lambda_d$) to ensure high-dynamic performance, particularly in the FW region.

Current Vector Control: The CVC strategy regulates the stator current components i_d and i_q (see Fig. 4.21). As indicated by the state-space formulation in (2.8), the regulation dynamic are governed by the differential inductances. Therefore, to maintain consistent bandwidth across the operating range, the current regulators require to be adaptive based on the estimated differential inductances $\tilde{l}_{dd}$ and $\tilde{l}_{qq}$ (see Fig. 2.27 and Fig. 2.31).

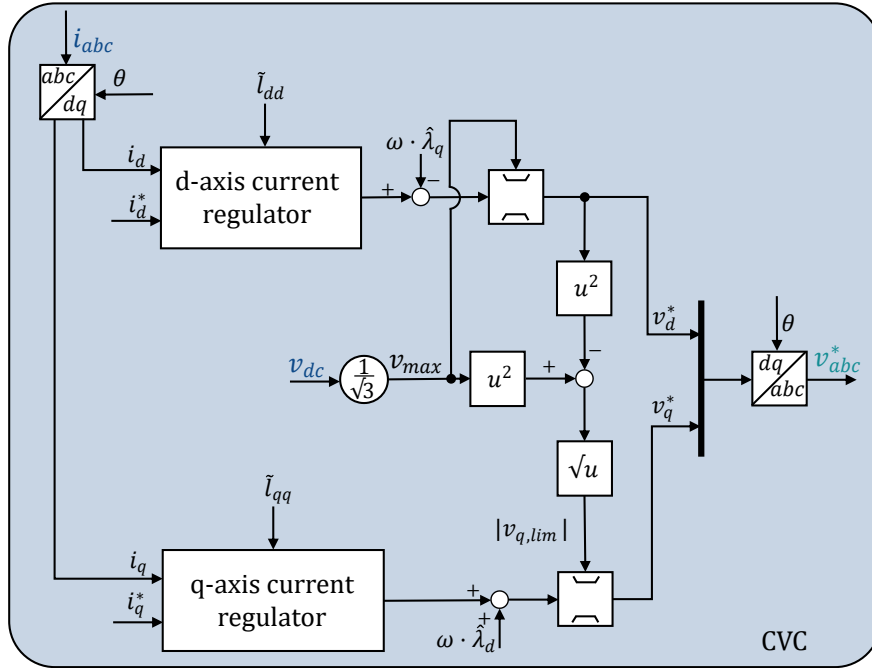


Fig. 4.21 Schematic of the CVC architecture: closed-loop regulation of stator dq currents (i_d, i_q).

Flux Vector Control: In contrast to CVC, the FVC strategy regulates the stator flux linkage components λ_d and λ_q (Fig. 4.22). A key advantage of this scheme, derived from (2.9), is that the state-space equation is independent from magnetic saturation. Consequently, the flux regulators exhibit operating-point invariance, simplifying the tuning process similarly to that of the flux amplitude regulator.

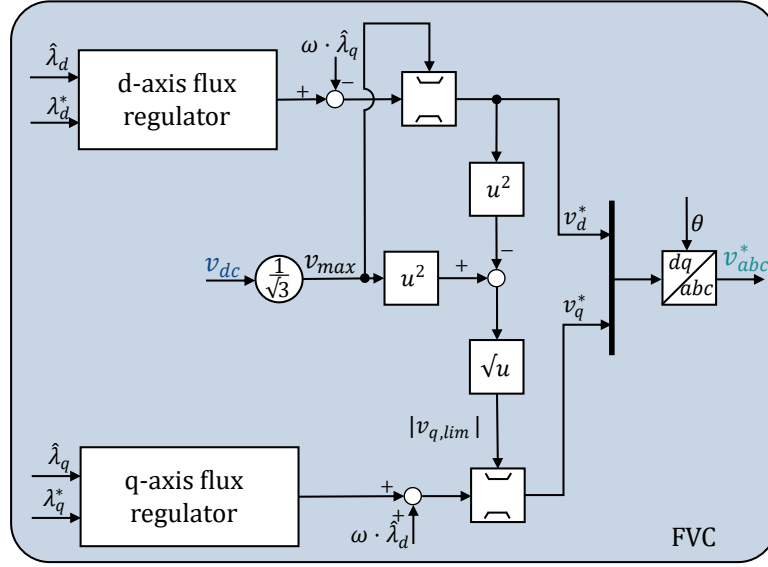


Fig. 4.22 Schematic of the FVC architecture: closed-loop regulation of stator dq flux linkages (λ_d, λ_q).

4.1.6 Rotor current regulator

Rotor current regulation is implemented via a PI controller. The voltage saturation limits are dynamically configured based on the power converter topology: $[0, v_{dc}]$ for unipolar excitation or $[-v_{dc}, v_{dc}]$ for H-bridge configurations. To account for the nonlinear magnetic behavior described in (2.5), the controller gains are adaptively tuned based on the real-time estimation of the rotor equivalent differential inductance $\tilde{l}_{eq,f}$ (see Fig. 2.35). Additionally, feed-forward compensation of the resistive voltage drop ($\tilde{R}_f \cdot i_f^*$) is implemented to enhance disturbance rejection, as shown in Fig. 4.23).

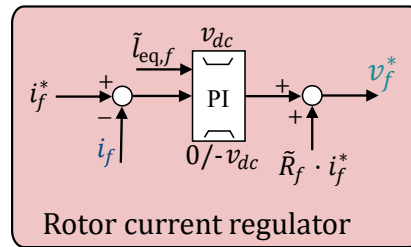


Fig. 4.23 Rotor current regulator.

4.1.7 PI regulators tuning

Although any regulator can be used for closed-loop control of each TCA, this work focuses on the most common PI regulators. The detailed description of how the PI regulators for flux and current control can be tuned is reported hereafter. It is noted that this discussion of PI regulator tuning is valid in any reference frame and for any axis. The subscripts "x" and "y" are used to identify a generic axis and the corresponding quadrature one, respectively.

Flux controller design

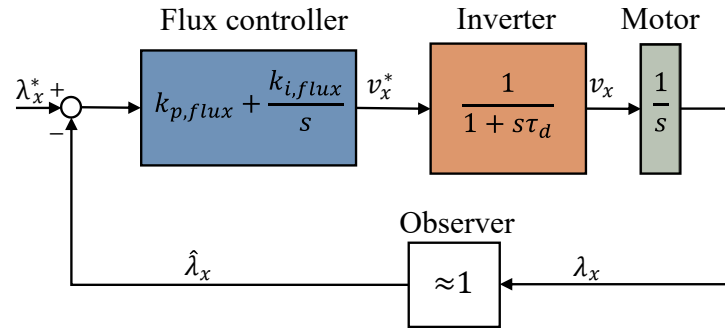


Fig. 4.24 Block diagram of the flux control loop in the Laplace domain.

The simplified flux control structure in the Laplace domain is depicted in Fig. 4.24. The inverter dynamics are modeled as a transport delay with unitary gain, approximated by a first-order transfer function:

$$H_{\text{inv}}(s) \simeq \frac{1}{1 + s \cdot \tau_d} = \frac{1}{1 + s \cdot (1.5 \cdot T_s)}, \quad (4.26)$$

where τ_d is the time delay introduced by the inverter to apply the commands (intrinsic digital processing delay plus the delay introduced by the symmetric PWM) and T_s represents the switching period (assumed equal to the sampling interval). Assuming ideal voltage reconstruction, the observer dynamics are treated as a unity gain. Consequently, the open-loop transfer function H_{OL} is derived as:

$$H_{\text{OL,flux}}(s) = \left(k_{p,\text{flux}} + \frac{k_{i,\text{flux}}}{s} \right) \cdot \frac{1}{s} \cdot \frac{1}{1 + s \cdot \tau_d} = k_{p,\text{flux}} \cdot \frac{s - \omega_{z,\text{flux}}}{s^2} \cdot \frac{1}{1 + s \cdot \tau_d}, \quad (4.27)$$

where $\omega_{z,\text{flux}} = -k_{i,\text{flux}}/k_{p,\text{flux}}$ denotes the controller zero. The system is characterized by a double pole at the origin, high-frequency pole introduced by the inverter delay and it is independent of machine parameters and operating point variations.

The controller tuning is performed as follows:

- **Bandwidth selection:** The crossover frequency $f_{c,\text{flux}}$ is constrained by the inverter delay. A theoretical upper limit is established as:

$$\omega_{c,\text{flux},\text{max}} = \tau_d^{-1} \Rightarrow f_{c,\text{flux},\text{max}} \simeq \frac{f_s}{10}. \quad (4.28)$$

- **Zero placement:** The controller zero $\omega_{z,\text{flux}}$ is positioned to satisfy the desired phase margin Φ_{PM} :

$$\omega_{z,\text{flux}} = \frac{\omega_{c,\text{flux}}}{\tan(\Phi_{\text{PM}} + \arctan(\omega_{c,\text{flux}} \tau_d))} \quad (4.29)$$

subject to the stability constraint $\Phi_{\text{PM}} + \arctan(\omega_{c,\text{flux}} \tau_d) < 90^\circ$.

- **Gain calculation:** Following the analytical method reported in [80], the integral and proportional gains are computed as:

$$k_{i,\text{flux}} = \omega_{c,\text{flux}}^2 \cdot \sqrt{\frac{1 + (\tau_d \omega_{c,\text{flux}})^2}{1 + \left(\frac{\omega_{c,\text{flux}}}{\omega_{z,\text{flux}}}\right)^2}}; \quad k_{p,\text{flux}} = \frac{k_{i,\text{flux}}}{\omega_{z,\text{flux}}}. \quad (4.30)$$

Current controller design

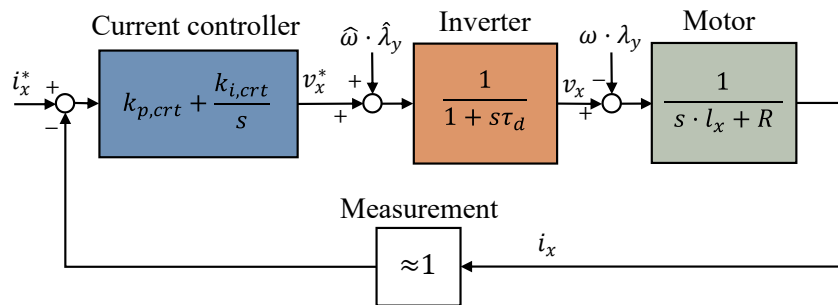


Fig. 4.25 Block diagram of the stator current control loop in the Laplace domain.

The current control scheme is illustrated in Fig. 4.25. Unlike the flux loop, the current plant accounts for the machine's resistive-inductive ($R - L$) dynamics. Assuming negligible measurement delay, the open loop transfer function is expressed as:

$$H_{OL,crt}(s) = \left(k_{p,crt} + \frac{k_{i,crt}}{s} \right) \cdot \frac{1}{1 + s \cdot \tau_d} \cdot \frac{1}{R + s \cdot l_x} = \frac{k_{p,crt}}{l_x} \cdot \frac{s + \omega_{z,crt}}{s \cdot (s + R/l_x)} \cdot \frac{1}{1 + s \cdot \tau_d}. \quad (4.31)$$

Applying the zero-pole cancellation technique, the controller zero is aligned with the machine's electrical pole ($\omega_{z,crt} = R/l_x$), yielding the relationship:

$$k_{i,crt} = k_{p,crt} \cdot \frac{R}{l_x}. \quad (4.32)$$

This simplification results in a first-order equivalent open loop transfer function:

$$H_{OL,crt}(s) = \frac{k_{p,crt}}{l_x} \cdot \frac{1}{s} \cdot \frac{1}{1 + s \tau_d}. \quad (4.33)$$

The design parameters are derived as follows:

- **Bandwidth and gains:** Selecting a typical phase margin $\Phi_{PM} = 60^\circ$, the crossover frequency and proportional gain are determined as:

$$\omega_{c,crt,max} = \frac{\tan(90^\circ - \Phi_{PM})}{\tau_d}; \quad k_{p,crt} = \omega_{c,crt} \cdot l_x. \quad (4.34)$$

The integral gain follows from (4.32).

Notably, the control gains in (4.34) explicitly depend on the differential inductance l_x . In EESMs, l_x is a highly nonlinear function of the state vector (i_d, i_q, i_f) due to magnetic saturation. Consequently, using constant gains is suboptimal. Instead, the controller employs adaptive gain scheduling based on the real-time estimation $\hat{l}_x$ provided by the magnetic model in the flux observer (see Fig. 4.3).

Rotor current controller design

The tuning methodology for the rotor current regulator is analogous to that of the stator current. The analytical framework remains valid upon substituting the stator parameters (R, l_x) with the rotor resistance R_f and the equivalent rotor differential

inductance $l_{eq,f}$. The latter is similarly estimated via the $I\theta$ observer to account for saturation effects. As indicated in Fig. 4.26, the disturbance term in this loop originates from stator-rotor cross-coupling rather than back-EMF; however, the feedback controller tuning remains identical.

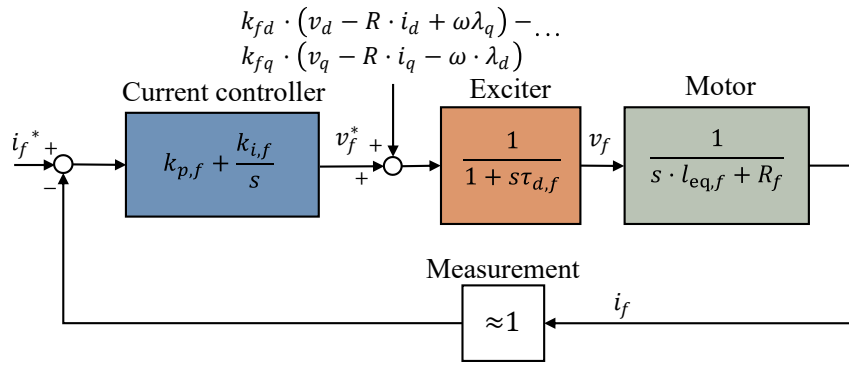


Fig. 4.26 Block diagram of the rotor current control loop in the Laplace domain.

Approximating the bandwidth ω_b of the PI regulator to the crossover frequency $\omega_b \approx \omega_c$, the tuning of the PI regulators for each TCA is summarized in Table 4.1.

Table 4.1 PI regulators tuning.

DFVC		FPC	
$ds\text{-axis } k_p$	$k_i \cdot \frac{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}{\omega_b}$	$ds\text{-axis } k_p$	$k_i \cdot \frac{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}{\omega_b}$
$ds\text{-axis } k_i$	$\omega_b^2 \cdot \sqrt{\frac{1 + (\omega_b \tau_d)^2}{1 + \left(\frac{\omega_b}{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}\right)^2}}$	$ds\text{-axis } k_i$	$\omega_b^2 \cdot \sqrt{\frac{1 + (\omega_b \tau_d)^2}{1 + \left(\frac{\omega_b}{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}\right)^2}}$
$qs\text{-axis } k_p$	$\omega_b \cdot \tilde{I}_{qs}$	$qs\text{-axis } k_p$	$k_i \cdot \frac{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}{\omega_b}$
$qs\text{-axis } k_i$	$\omega_b \cdot \tilde{R}$	$qs\text{-axis } k_i$	$\omega_b^2 \cdot \sqrt{\frac{1 + (\omega_b \tau_d)^2}{1 + \left(\frac{\omega_b}{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}\right)^2}}$
DTC		FVC	
$ds\text{-axis } k_p$	$k_i \cdot \frac{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}{\omega_b}$	$d\text{-axis } k_p$	$k_i \cdot \frac{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}{\omega_b}$
$ds\text{-axis } k_i$	$\omega_b^2 \cdot \sqrt{\frac{1 + (\omega_b \tau_d)^2}{1 + \left(\frac{\omega_b}{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}\right)^2}}$	$d\text{-axis } k_i$	$\omega_b^2 \cdot \sqrt{\frac{1 + (\omega_b \tau_d)^2}{1 + \left(\frac{\omega_b}{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}\right)^2}}$
$qs\text{-axis } k_p$	$\omega_b \cdot \tilde{I}_{qs}$	$q\text{-axis } k_p$	$k_i \cdot \frac{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}{\omega_b}$
$qs\text{-axis } k_i$	$\omega_b \cdot \tilde{R}$	$q\text{-axis } k_i$	$\omega_b^2 \cdot \sqrt{\frac{1 + (\omega_b \tau_d)^2}{1 + \left(\frac{\omega_b}{\tan(\Phi_{PM} + \arctan(\omega_b \tau_d))}\right)^2}}$
CVC		Rotor current	
$d\text{-axis } k_p$	$\omega_b \cdot \tilde{I}_{dd}$	k_p	$\omega_b \cdot \tilde{I}_{eq,f}$
$d\text{-axis } k_i$	$\omega_b \cdot \tilde{R}$	k_i	$\omega_b \cdot \tilde{R}_f$
$q\text{-axis } k_p$	$\omega_b \cdot \tilde{I}_{qq}$		
$q\text{-axis } k_i$	$\omega_b \cdot \tilde{R}$		

4.2 Dead-beat flux control

The torque control solutions described previously require parameter-dependent regulator tuning (CVC, DFVC, DTC), and their dynamic performance is inherently linked to the regulator bandwidth and, consequently, the digital control sampling frequency.

In the context of PMSM drives, DB control strategies are renowned for achieving torque settling within two sampling periods [120, 121], subject to converter voltage and current constraints. A distinct advantage of DB formulations over conventional PI synchronous frame regulators is their deterministic behavior under saturation limits and their ability to sustain performance at low switching-to-fundamental frequency ratios, provided that the predictive model is accurate. According to these features, Dead-Beat Direct Torque and Flux Control (DB-DTFC) architectures have been adapted for EESMs [137, 138], typically deriving the field flux reference to enforce unity power factor. However, unity power factor operation is not synonymous with efficiency maximization; achieving optimal energy conversion requires more sophisticated loss-minimization algorithms [139, 140, 134, 130, 131, 135, 141]. Furthermore, existing DB-DTFC schemes often necessitate complex current estimators [121], increasing computational burden. Moreover, the current state of the art lacks EESM torque-control solutions that eliminate inner-loop calibration; performance remains intrinsically coupled to inductance variations (i.e., magnetic saturation), regardless of the regulator topology.

To cope with all above mentioned limitations, this Section proposes a "plug-and-play" torque control framework based on Dead-Beat Direct Flux Control (DB-DFC) for both stator and rotor flux linkages. By directly regulating the flux state variables, whose discrete-time evolution is straightforward to model with high fidelity, the control bandwidth is rendered invariant to magnetic saturation and sampling frequency. Consequently, this approach eliminates the need for gain scheduling or extensive inner-loop calibration procedures, applicable to both steady-state and transient regimes. The proposed architecture synthesizes torque accuracy by combining minimization of total copper losses with an integral error compensation mechanisms. This hybrid structure ensures that parameter mismatches in the flux maps are automatically rejected, maintaining robust torque regulation.

Regarding the implementation topologies, the literature reports advanced FVC strategies for PMSMs in both rotor (dq) and stationary ($\alpha\beta$) coordinates [126, 116], as well as the recently introduced FPC in the stator flux (dqs) frame [80]. However, the application and comparative analysis of these direct flux methodologies within the EESM domain remain unexplored. Therefore, this work advances the state-of-the-art in EESM control through the following contributions:

- **Self-calibrated control architecture:** The first proposal of torque controllers for traction EESMs that eliminate demanding tuning and calibration procedures.
- **Dual direct flux regulation:** Implementation of simultaneous direct control of stator and rotor flux linkages, enabling ultra-high dynamic bandwidth while satisfying minimum copper loss criteria.
- **Comparative topological analysis:** The first rigorous comparison of DB-DFC implementations for EESMs, evaluating $\alpha\beta$, dq and dqs frames.
- **Experimental validation:** A comprehensive experimental assessment of the proposed strategies, demonstrating robustness against parameter mismatches and low sampling-to-fundamental frequency ratios.

The control structure proposed is the same as the one described before in this Chapter and reported in Fig. 4.1. A few control modules require modification or addition, as outlined below.

4.2.1 EESM discrete modeling

EESMs can be described in different reference frames, namely the stationary $\alpha\beta$, the rotor dq , and the stator dqs coordinates. As previously described in Chapter 2, the electromagnetic equations of the machine can be expressed in all of the aforementioned reference frames. When implemented in a discrete digital controller, each formulation exhibits specific characteristics, which are highlighted in the following.

Electromagnetic model in stationary $\alpha\beta$ coordinates

The stator voltage equation in the stationary $\alpha\beta$ frame is provided in (2.19). Considering a discrete-time domain, the discrete form of (2.19) using the forward Euler discretization is given by:

$$\vec{v}_{\alpha\beta}^k = R \cdot \begin{bmatrix} i_{\alpha}^k \\ i_{\beta}^k \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda_{\alpha}^{k+1} - \lambda_{\alpha}^k \\ \lambda_{\beta}^{k+1} - \lambda_{\beta}^k \end{bmatrix} \quad (4.35)$$

where T_s is the sampling period while k stands for the generic sampling time instant.

Electromagnetic model in rotor dq coordinates

The stator voltage equation in the rotor dq reference frame is expressed by (2.1). Assuming $\omega^{k+1} \simeq \omega^k$, (2.1) in discrete form becomes:

$$\vec{v}_{dq}^k = R \cdot \begin{bmatrix} i_d^k \\ i_q^k \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda_d^{k+1} - \lambda_d^k \\ \lambda_q^{k+1} - \lambda_q^k \end{bmatrix} + \omega^k \cdot \begin{bmatrix} -\lambda_q^k \\ \lambda_d^k \end{bmatrix} \quad (4.36)$$

Electromagnetic model in stator dqs coordinates

The voltage equation in the stator flux-oriented dqs frame can be expressed as (2.10), whose discrete form is:

$$\vec{v}_{dqs}^k = R \cdot \begin{bmatrix} i_{ds}^k \\ i_{qs}^k \end{bmatrix} + \frac{1}{T_s} \cdot \begin{bmatrix} \lambda^{k+1} - \lambda^k \\ \lambda^k \cdot (\delta^{k+1} - \delta^k) \end{bmatrix} + \begin{bmatrix} 0 \\ \lambda^k \cdot \omega^k \end{bmatrix} \quad (4.37)$$

Rotor model

As noted, the EESM also enables direct field regulation via the rotor current i_f (2.20). The discrete form of the rotor voltage equation is:

$$v_f^k = R_f \cdot i_f^k + \frac{\lambda_f^{k+1} - \lambda_f^k}{T_s} \quad (4.38)$$

Building on this modeling framework, the proposed controller structure presents some differences with the one already proposed before and shown in Fig. 4.2. In particular, in the next subsections, the discrete form of the stator flux observer, the rotor flux observer, the integral correction loops and the DB control schemes are presented.

4.2.2 Discrete stator flux observer

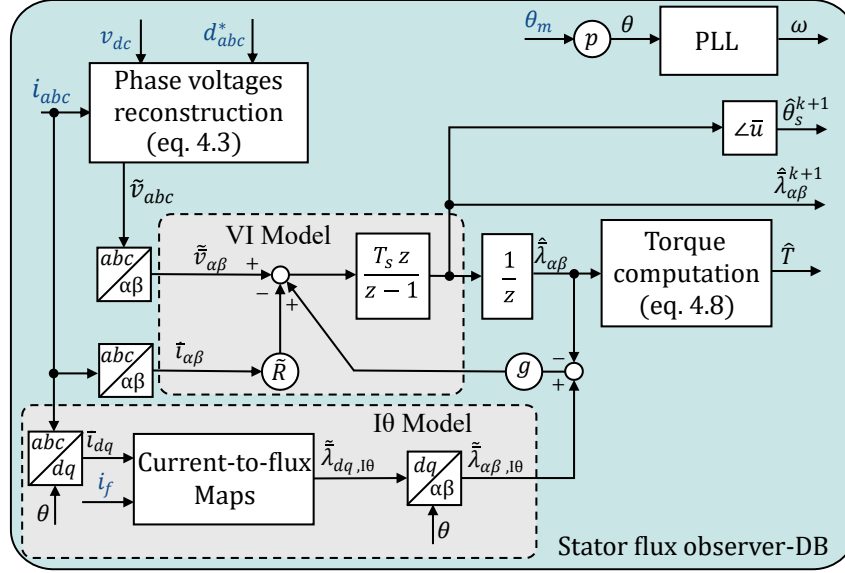


Fig. 4.27 Schematic view of the adopted discrete stator flux observer.

The stator flux vector is evaluated using a hybrid observer that integrates two model-based estimators as discussed in Chapter 4.1.1. Its discrete-time implementation is depicted in Fig. 4.27.

The discrete stator flux observer provides two key outputs for the implementation of the DB regulators:

- The predicted stator flux vector in the stationary $\alpha\beta$ frame $\hat{\lambda}_{\alpha\beta}^{k+1}$ and the corresponding position $\hat{\theta}_s^{k+1}$.
- The actual electromagnetic torque, given by:

$$\hat{T} = \frac{3}{2} \cdot p \cdot (\hat{\lambda}_{\alpha} \cdot i_{\beta} - \hat{\lambda}_{\beta} \cdot i_{\alpha}) \quad (4.39)$$

4.2.3 Discrete rotor flux observer

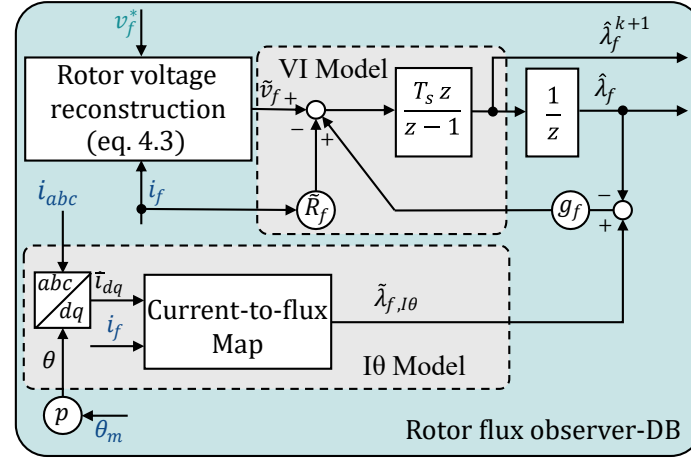


Fig. 4.28 Schematic view of the adopted discrete rotor flux observer.

The implementation of DB control for the rotor circuit requires a one-step prediction of the rotor flux linkage. This is realized via a dual-model observer topology, structurally analogous to the stator estimator, which computes the predicted state $\hat{\lambda}_f^{k+1}$ (see Fig. 4.28).

The Current Model ($I\theta$) incorporates the machine’s nonlinear magnetic topology via FEA or experimentally derived LUTs, as depicted in Fig. 2.23. While these maps are static, potential discrepancies arising from model mismatch are mitigated by a dedicated correction mechanism described in subsequent sections.

Operationally, this observer differs significantly from its stator counterpart. The $I\theta$ estimator dominates steady-state convergence and continuously initializes the backward Euler integrator of the Voltage Model (VI). Conversely, the VI estimator contribution is significant solely during dynamic transients (where $d\lambda_f/dt \neq 0$), as the excitation voltage v_f balances the resistive drop in steady-state ($v_f \approx R_f i_f$). Consequently, the VI path is critical for ensuring robust DB prediction during torque transients. To prioritize the VI dynamics during flux excursions while ensuring steady-state stability, the crossover gain g_f is set to a minimal value. Furthermore, the observer's integral architecture inherently acts as a low-pass filter, rendering the flux estimate immune to high-frequency measurement noise.

4.2.4 Stator and rotor flux references computation

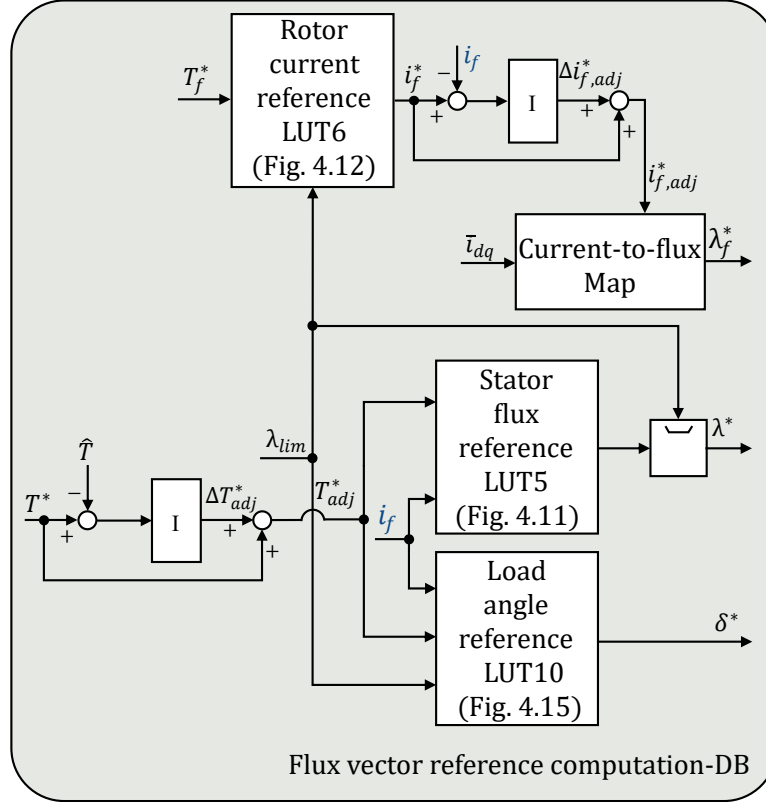


Fig. 4.29 Schematic view of the stator and rotor flux references computation.

Stator and rotor flux references are obtained from the torque references previously computed. Specifically, the stator torque reference T^* is used to determine the stator flux vector references of amplitude λ^* and load angle δ^* . In parallel, the rotor current reference i_f^* is derived from the rotor-side torque reference T_f^* and then converted into the corresponding rotor flux reference λ_f^* through the same current-to-flux map implemented in the rotor flux observer. For simplicity, the proposed torque controllers adopt an efficiency-oriented strategy based on minimizing the overall Joule losses of the machine, comprising both stator and rotor contributions [139].

Inaccuracies in the magnetic characterization maps inevitably propagate to the reference generation LUTs (LUT5, LUT10), introducing steady-state torque deviations. While this sensitivity is inherent to current-model-based observers [186, 187, 109], it is effectively mitigated in the medium-to-high speed regime where the estimator dynamics are dominated by the parameter-insensitive Voltage Model. Consequently,

the observed electromagnetic torque $\hat{T}$, derived via (4.8), serves as a reliable feedback variable for error compensation in these operating regions. To guarantee high-fidelity steady-state regulation, an outer-loop integral compensator is implemented [126]. This regulator dynamically adjusts the torque command T^* based on the discrepancy between the reference and the observed torque, generating the compensated input T_{adj}^* for LUT5 and LUT10 (see Fig. 4.29). The integrator bandwidth is selected to ensure convergence primarily during steady-state operation, leveraging the accuracy of the VI-dominated estimate at medium and high speed. An analogous topology is applied to the rotor current control to enforce consistency between the reference current i_f^* and the rotor flux reference λ_f^* , utilizing the measured rotor current i_f as the feedback variable (Fig. 4.29). The tuning of these integral regulators is straightforward: the gain is set as the reciprocal of the desired disturbance rejection time constant (typically in the range of 100–500 ms). Detailed design criteria for these compensation loops are provided in the experimental validation chapter.

4.2.5 DB control

Digital controllers inherently introduce a one-step execution delay between the computation of voltage references and their application via switch-mode power converter. Specifically, for a generic sampling instant k , the machine state variables (currents, fluxes, and torque) at $k + 1$ are determined by the voltage commands computed at $k - 1$. Consequently, the control action calculated at instant k effectively determines the system state at $k + 2$. The DB strategy addresses this latency by utilizing discrete-time state-space models to predict the evolution of controlled variables, computing the voltage vector required to drive the state at $k + 2$ to its reference value. Accordingly, the proposed torque controllers derive the voltage commands for instant $k + 1$ by enforcing the condition that the stator and rotor fluxes at $k + 2$ equal their respective targets. The reference discrete-time control laws are derived from the discrete voltage equations (4.35), (4.36), (4.37), and (4.38):

$$\bar{v}_{\alpha\beta}^* = R \cdot \begin{bmatrix} i_{\alpha}^{k+1} \\ i_{\beta}^{k+1} \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda_{\alpha}^* - \lambda_{\alpha}^{k+1} \\ \lambda_{\beta}^* - \lambda_{\beta}^{k+1} \end{bmatrix} \quad (4.40)$$

$$\bar{v}_{dq}^* = R \cdot \begin{bmatrix} i_d^{k+1} \\ i_q^{k+1} \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda_d^* - \lambda_d^{k+1} \\ \lambda_q^* - \lambda_q^{k+1} \end{bmatrix} + \omega^{k+1} \begin{bmatrix} -\lambda_q^{k+1} \\ \lambda_d^{k+1} \end{bmatrix} \quad (4.41)$$

$$\bar{v}_{dqs}^* = R \cdot \begin{bmatrix} i_{ds}^{k+1} \\ i_{qs}^{k+1} \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda^* - \lambda^{k+1} \\ \lambda^{k+1}(\delta^* - \delta^{k+1}) \end{bmatrix} + \begin{bmatrix} 0 \\ \lambda^{k+1} \omega^{k+1} \end{bmatrix} \quad (4.42)$$

$$v_f^* = R_f \cdot i_f^{k+1} + \frac{\lambda_f^* - \lambda_f^{k+1}}{T_s} \quad (4.43)$$

Strict adherence to these equations needs the prediction of the currents and of the electrical speed at step $k + 1$. However, the design of EESMs for traction applications justifies some approximations. Stator resistive drops are typically negligible except at very low speeds; moreover, at steady-state, the deviation between predicted and measured currents is minimal. Similarly, the rotor current dynamics are governed by a time constant significantly larger than the sampling period T_s , rendering detailed prediction non-critical. Furthermore, due to mechanical inertia, the electrical speed ω can be assumed constant over the prediction horizon ($\omega^{k+1} \approx \omega^k$). Consequently, the control laws are simplified as follows:

$$\bar{v}_{\alpha\beta}^* = R \cdot \begin{bmatrix} i_{\alpha}^k \\ i_{\beta}^k \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda_{\alpha}^* - \lambda_{\alpha}^{k+1} \\ \lambda_{\beta}^* - \lambda_{\beta}^{k+1} \end{bmatrix} \quad (4.44)$$

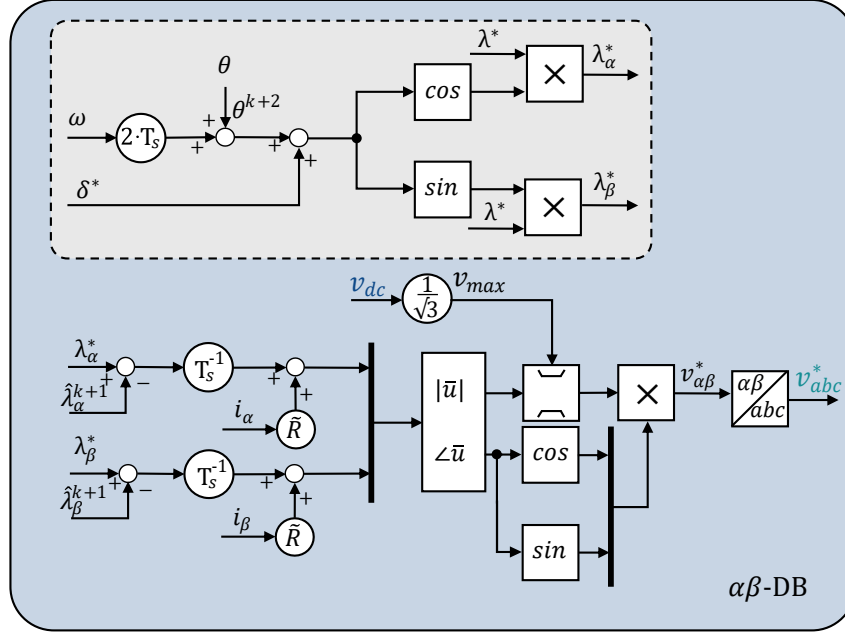
$$\bar{v}_{dq}^* = R \cdot \begin{bmatrix} i_d^k \\ i_q^k \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda_d^* - \lambda_d^{k+1} \\ \lambda_q^* - \lambda_q^{k+1} \end{bmatrix} + \omega^k \begin{bmatrix} -\lambda_q^{k+1} \\ \lambda_d^{k+1} \end{bmatrix} \quad (4.45)$$

$$\bar{v}_{dqs}^* = R \cdot \begin{bmatrix} i_{ds}^k \\ i_{qs}^k \end{bmatrix} + \frac{1}{T_s} \begin{bmatrix} \lambda^* - \lambda^{k+1} \\ \lambda^{k+1}(\delta^* - \delta^{k+1}) \end{bmatrix} + \begin{bmatrix} 0 \\ \lambda^{k+1} \omega^k \end{bmatrix} \quad (4.46)$$

$$v_f^* = R_f \cdot i_f^k + \frac{\lambda_f^* - \lambda_f^{k+1}}{T_s} \quad (4.47)$$

While these approximations may degrade performance during ultra-fast transients at very low speeds (where resistive drops and inverter dead-time effects are pronounced), such operating conditions are rare in traction profiles. Any resultant flux tracking error is typically transient and minor. At steady-state operation, residual discrepancies are effectively nullified by the outer integral compensation loops.

The proposed DB controllers implement direct flux vector control across three coordinate systems: stationary $\alpha\beta$, rotor dq , and stator dqs .

$\alpha\beta$ -DB Stator Flux ControlFig. 4.30 Schematic of the $\alpha\beta$ -frame DB flux control architecture.

The $\alpha\beta$ -frame implementation (see Fig. 4.30) necessitates a two-step prediction of the rotor electrical position to align the flux references:

$$\theta^{k+2} = \theta^{k+1} + \omega^{k+1} T_s \simeq \theta + 2 \cdot \omega \cdot T_s \quad (4.48)$$

The stator flux vector components are then derived from the amplitude λ^* and load angle δ^* commands:

$$\begin{cases} \lambda_{\alpha}^* = \lambda^* \cos(\theta^{k+2} + \delta^*) \\ \lambda_{\beta}^* = \lambda^* \sin(\theta^{k+2} + \delta^*) \end{cases} \quad (4.49)$$

Reference voltages are computed via (4.44). Given that the $\alpha\beta$ equations are decoupled, voltage saturation is applied to the vector magnitude $|v_{\alpha\beta}^*| \leq v_{max}$ while preserving the vector angle, thereby avoiding bias towards either axis. The final phase voltages v_{abc}^* are generated via the inverse Clarke transformation.

dq-DB Stator Flux Control

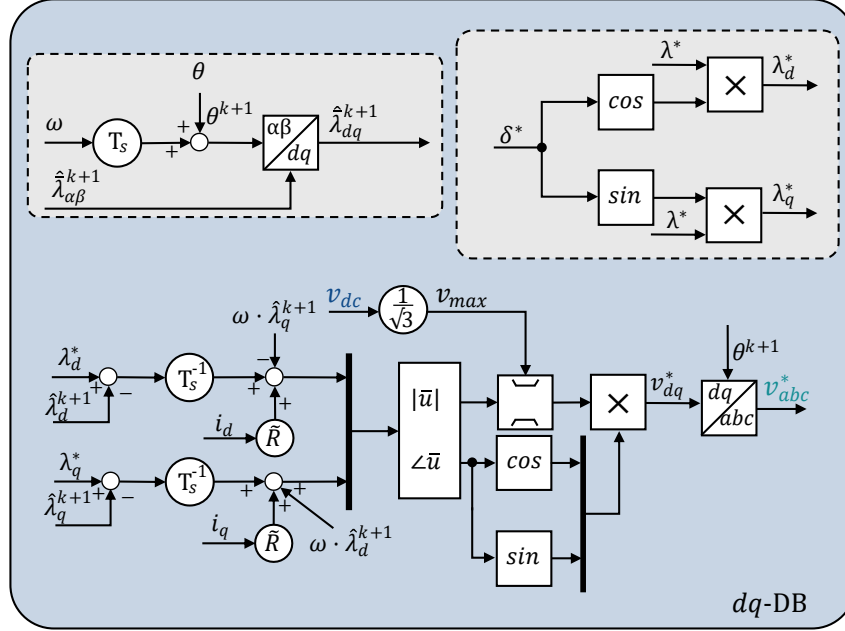
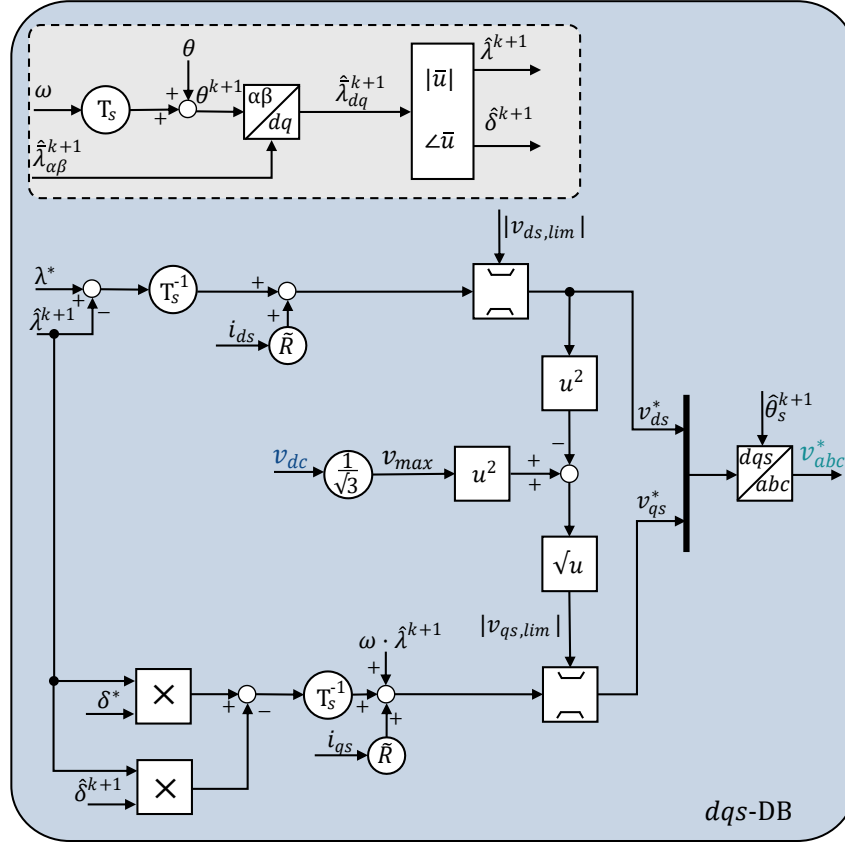


Fig. 4.31 Schematic of the *dq*-frame DB flux control architecture.

In the *dq*-frame implementation (see Fig. 4.31), the flux references are defined as:

$$\begin{cases} \lambda_d^* = \lambda^* \cos(\delta^*) \\ \lambda_q^* = \lambda^* \sin(\delta^*) \end{cases} \quad (4.50)$$

Prediction of the stator flux vector $\hat{\lambda}_{dq}^{k+1}$ requires transforming the observer output $\hat{\lambda}_{\alpha\beta}^{k+1}$ into the rotor frame using the predicted angle $\theta^{k+1} = \theta + \omega \cdot T_s$. The reference voltages are calculated using (4.45). Although the *dq* axes are dynamically coupled via back-EMF terms, a phase-preserving saturation strategy (identical to the $\alpha\beta$ case) is adopted to maintain vector directionality during deep saturation transients, avoiding complex axis-prioritization schemes. The final output v_{abc}^* is obtained via the inverse Park transformation using θ^{k+1} as rotation angle.

***dqs*-DB Stator Flux Control**Fig. 4.32 Schematic of the *dqs*-frame flux DB control architecture.

The *dqs*-frame approach, shown in Fig. 4.32, utilizes the predicted flux amplitude $\hat{\lambda}^{k+1}$ and load angle $\hat{\delta}^{k+1}$, which are extracted from the *dq*-transformed observer estimate $\hat{\lambda}_{dq}^{k+1}$ as feedback. Reference voltages are computed via (4.46). A key advantage of the *dqs* frame is the inherent decoupling of flux and torque production, analogous to a separately excited DC motor. The back-EMF is concentrated on the *qs*-axis. Consequently, the *ds*-axis requires only a minimal voltage margin $v_{ds,lim}$ to compensate for resistive drops and errors, typically selected as:

$$2 \cdot R \cdot i_{max} \leq v_{ds,lim} \leq 5 \cdot R \cdot i_{max} \quad (4.51)$$

This allows the q_s -axis to utilize almost the entire inverter voltage capability. The final phase voltages are generated via the inverse Park transformation oriented to the predicted stator flux angle $\hat{\theta}_s^{k+1}$ as rotation angle.

Rotor DB Flux Control

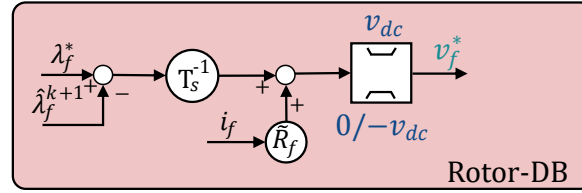


Fig. 4.33 Schematic of the rotor flux DB regulator.

The rotor flux is controlled by directly implementing (4.47) (see Fig. 4.33). The calculated reference v_f^* is clamped to the exciter's limits: $[0, v_{dc}]$ for unipolar topologies or $[-v_{dc}, v_{dc}]$ for bipolar configurations. The critical aspect of this implementation is the synchronization of stator and rotor control algorithms within a single digital unit; any timing drift would compromise the fidelity of the rotor flux reference λ_f^* , degrading performance.

4.3 Summary

This chapter presents a comprehensive unified torque control approach for EESMs in traction applications. The core of the control scheme developed is the UTC, a versatile architecture designed to implement multiple TCAs, including CVC, DTC, DFVC, FPC, and FVC, within a single topological arrangement.

The proposed UTC addresses the inherent complexity of EESMs, such as the strong stator-rotor magnetic cross-coupling and the disparate time constants between the two circuits. A pivotal contribution is the dual-torque reference strategy, which decouples the stator and rotor regulation paths (T^* and T_f^*) to ensure high-dynamic performance and transient stability even under constrained rotor current dynamics. The control scheme integrates variable current limits, minimization of the total Joule losses and a systematic control reference linearization methodology through LUTs based on a normalization technique, enabling accurate torque regulation across

the entire operating range for each of the TCAs considered. The description and experimental validation of the proposed UTC with variable current limits for EESMs have been published in the journal articles (J2) and (J3).

Furthermore, a "plug-and-play" DB-DFC strategy is proposed to be integrated in the UTC with minor modification to the control modules. By directly regulating the stator and rotor flux linkages in discrete-time, this approach eliminates the need for extensive gain scheduling and inner-loop calibration. The DB-DFC is rigorously derived and evaluated in stationary $\alpha\beta$, rotor dq , and stator dqs coordinates. The chapter concludes by demonstrating how the implementation of an integral error compensation mechanisms ensures robust steady-state accuracy and immunity to parameter mismatches, providing the DB-DFC as a high-performance solution for modern automotive powertrains.

Chapter 5

Experimental validation

The experimental validation of the proposed torque controllers has been carried out using as MUT a commercial sample of the EESM employed in the Renault ZOE EV R315, whose characteristics and primary data have been previously described in Chapter 2.3.

5.1 Test rig

The experimental validation was conducted on a dynamic test bench where the MUT operates in torque control mode, mechanically coupled to a high-speed DM serving as a speed-controlled prime mover (see Fig. 5.1). To assess the performance of the proposed torque control solution, the mechanical shaft torque is measured via a high-precision HBK T12HP torque transducer installed in-line between the machines, as shown in Fig. 5.1 (*b*).

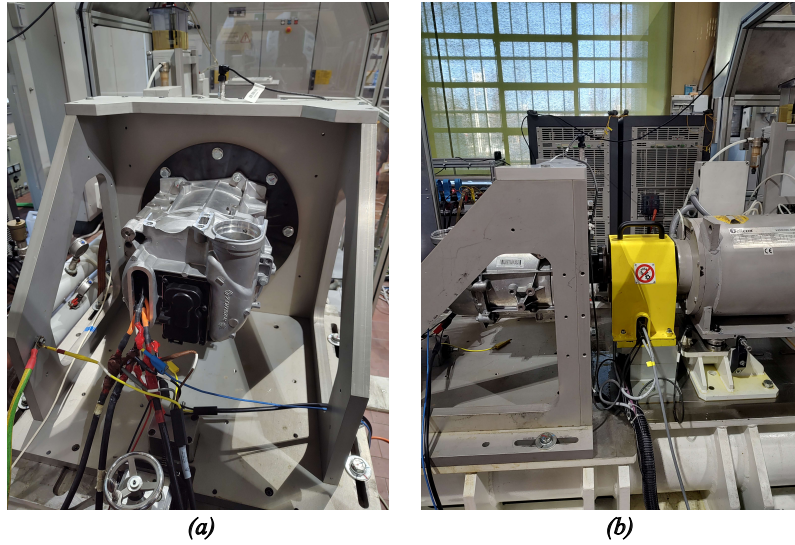


Fig. 5.1 View of the test rig: (a) EESM under test; (b) EESM coupled with the driving machine.

The MUT stator windings are fed by a Silicon Carbide (SiC) inverter (rated 800 V, 650 A), which is supplied by a bidirectional DC source regulated to the nominal DC-link voltage of 350 V. Rotor excitation is provided by a dedicated IGBT-based full-bridge power module fed at the same DC supply as the stator inverter. Both converters operate with a switching frequency synchronized to the control sampling frequency, utilizing single-edge PWM. The power converters stage are shown in Fig. 5.2.

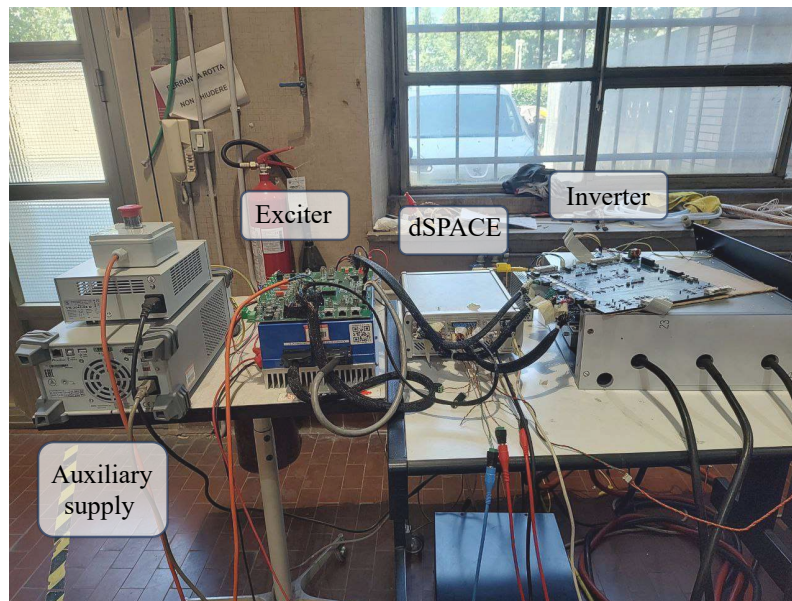


Fig. 5.2 View of the power converters (inverter and exciter).

Real-time control is implemented by a dSPACE MicroLabBox platform, which governs both the stator and rotor power stages. This unit handles algorithm execution, PWM signal generation, and synchronous signal acquisition. The control architecture was fully developed within the MATLAB/Simulink environment and deployed via automatic C-code generation. The real-time control of the speed controller software for the DM and the dSPACE Control Desk for the MicroLabBox is handled using two PCs, as shown in Fig. 5.3.



Fig. 5.3 View of the real-time control PCs setup.

The experimental setup is subject to operational constraints imposed by the mechanical coupling and the DM's performance. Consequently, the maximum torque and rotational speed are limited to 150 Nm and 10000 r/min, respectively. Nevertheless, this operating range permits rigorous validation of the proposed torque controllers in the machine's high-power region, enabling the testing of the peak power output of 100 kW.

5.2 Experimental results

The proposed UTC underwent a comprehensive experimental validation campaign designed to assess its performance across the machine's entire torque-speed operating range. The investigation focused on five primary performance metrics:

- **High-torque and deep FW capability:** Verifying the controller's ability to exploit the full torque and power potential of the machine, particularly within the voltage-limited deep FW and MTPV region.
- **Dynamic regulation:** Evaluating the tracking performance under rapid torque commands, including transient response at constant speed and fast reversals under variable speed conditions.
- **Steady-state torque accuracy:** Quantifying the precision of torque regulation across the speed range, ensuring strict adherence to voltage, current, and flux constraints.
- **Online limit variation:** Verifying the capability of the torque control algorithm to maintain accurate torque regulation even during the online variation of the current limits.
- **Efficiency:** Evaluating the efficiency of the proposed minimum Joule losses strategy.

Consequently, the validation protocol was structured into six representative test scenarios applied to all TCAs comprising the three DB-DFC implementations:

1. Maximum Torque Per Speed (MTPS) profile;

2. Torque transients;
3. Fast torque reversals;
4. Steady-state accuracy;
5. Variable current limits;
6. MTPV limitation.

All tests have been conducted using the TCAs discussed in this thesis: DFVC, DTC, FPC, CVC, FVC, and DB-DFC implemented in all three reference frames ($\alpha\beta, dq, dqs$). Except for the DB TCA, PI regulators have been employed, and for all controllers except CVC, the theoretical small-signal bandwidth of the stator control loops has been set to 200 Hz, while the rotor current control bandwidth has been set to 400 Hz. Due to the strong mutual coupling between the stator d -axis and the rotor excitation axis in the CVC strategy, tuning of the PI controllers becomes significantly more complex. To avoid instability and high-frequency noise in the torque response, the stator and rotor current control loop bandwidths for CVC have been limited to 50 Hz and 100 Hz, respectively. Although magnetic saturation has been rigorously accounted for by adjusting the PI regulator gains according to the design procedures reported in Chapter 4.1.7, these couplings force a trade-off in setting the small-signal bandwidth. On the one hand, keeping the small-signal bandwidth as low as possible mitigates oscillations in the dq and rotor currents, since the interaction between the stator and rotor PI regulators is reduced. On the other hand, a very low small-signal bandwidth of the current control loops does not allow for obtaining an acceptable dynamic performance of torque regulation. It is highlighted that the derating of the regulation bandwidth of the inner control loops has been applied only to CVC, since the other TCAs are ruled by state-space equations that are quite decoupled, thereby mitigating (DTC and DFVC) or avoiding (FPC and FVC) the interaction between the PI regulators of the inner control loops. For each PI regulator, the theoretical phase margin has been set to 60 degrees. The stator and rotor observer crossover frequencies (g, g_f) were set to 20 Hz (approximately 6% of the maximum fundamental frequency of 333 Hz at 10000 r/min). While voltage error compensation mechanisms [184, 187] typically allow for lower bandwidths, these conservative settings were maintained without specific calibration. Similarly, the flux-weakening voltage margin was fixed at $k_{fw} = 0.9$. The choice to avoid extensive optimization

procedures is made in order to streamline the tuning procedures. This way, it is also possible to demonstrate the "plug-and-play" nature of the proposed DB-DFC architecture, which does not require any PI regulators. Consequently, all control parameters have been selected according to standard heuristic guidelines. The outer-loop integral correction gains were tuned for moderate response times: the rotor current loop gain was set to 5 s^{-1} (residual error elimination within 0.2 s), and the torque loop gain to 2.5 s^{-1} (0.4 s settling time). These values, while sufficient for automotive requirements, were chosen to validate the robustness of the correction mechanisms without specific fine-tuning.

The LUTs governing the control maps of the PI-regulated TCAs have been computed from the experimentally derived current-to-flux maps of the EESM under test [81, 163]. Instead, to validate the integral correction loops of the DB-DFC approaches under practical detuning conditions, the control LUTs were populated using data derived exclusively from FEA [50, 188, 189]. This approach is typical in automotive development and introduces realistic parameter mismatches.

The input variables of such LUTs were discretized as follows: rotor current $i_f \in [0, i_{f,max} = 15]\text{ A}$ (16 points), flux limit $\lambda_{lim} \in [0, \lambda_{max} = 0.24]\text{ Vs}$ (25 points), stator current limit $i_{lim} \in [0, 385]\text{ A}$ with 11 A step (36 pt), rotor current limit $i_{f,lim} \in [0, i_{f,max} = 15]\text{ A}$ with 1 A step (16 points), and finally all the pu quantities $(i_{lim,pu}, T_{pu}, T_{f,pu}) \in [0, 1]$ with a 0.1 step (11 pt). Therefore, the dimensions of the LUTs were: LUT1 16×25 ; LUT2 $25 \times 36 \times 16$; LUT3 $25 \times 11 \times 16$; LUT4 25×16 ; LUT5 $11 \times 16 \times 11$; LUT6 $11 \times 25 \times 16$; LUT7 16×1 ; LUT8 16×1 ; LUT9 11×16 ; LUT10 $11 \times 11 \times 16$. The resulting memory footprint is approximately 2 MB. The control algorithms demonstrated a total execution time of approximately $30\text{ }\mu\text{s}$ on the dSPACE platform, confirming their feasibility for deployment on standard automotive-grade microcontrollers. It is highlighted that lower-resolution settings are allowed for most of the proposed LUTs. Indeed, the torque accuracy relies on LUT10 for all TCAs except DFVC and DTC. Therefore, only LUT10 should be prioritised if memory-saving optimization procedures become necessary.

To rigorously stress-test the robustness of the DB formulation, the sampling frequency was intentionally constrained to $f_s = 3\text{ kHz}$ for all tests where not specified. At the maximum speed of 10000 r/min (333 Hz fundamental frequency), this results in a sampling-to-fundamental frequency ratio of approximately 9. This aggressive constraint serves to validate the superior dynamic capability and stability

of the proposed predictive control under conditions where conventional PI regulators typically suffer performance degradation. Indeed, for the other PI-regulated TCAs, the switching frequency f_s has been set to 10 kHz.

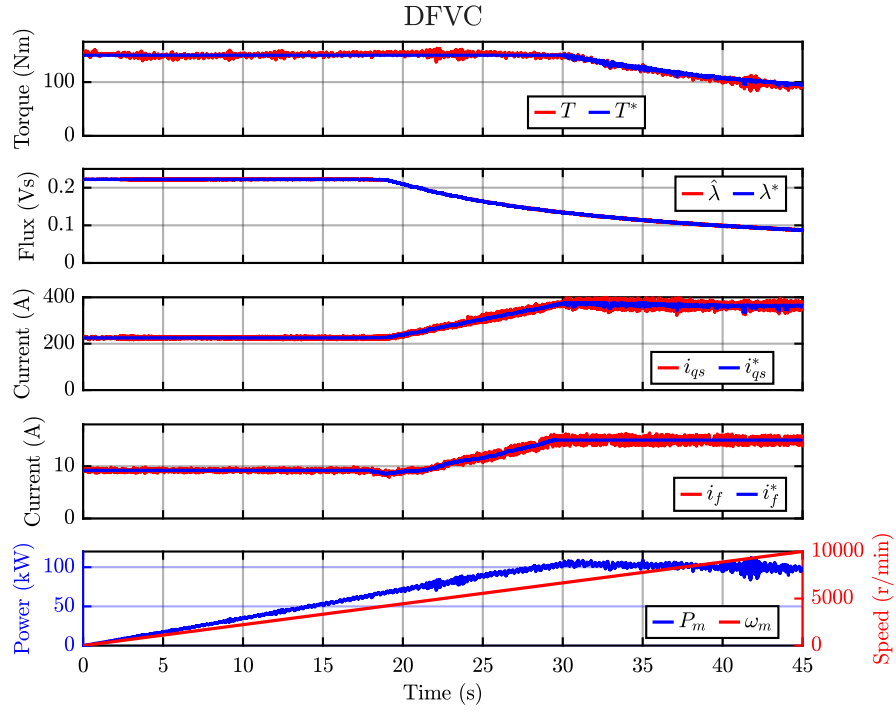
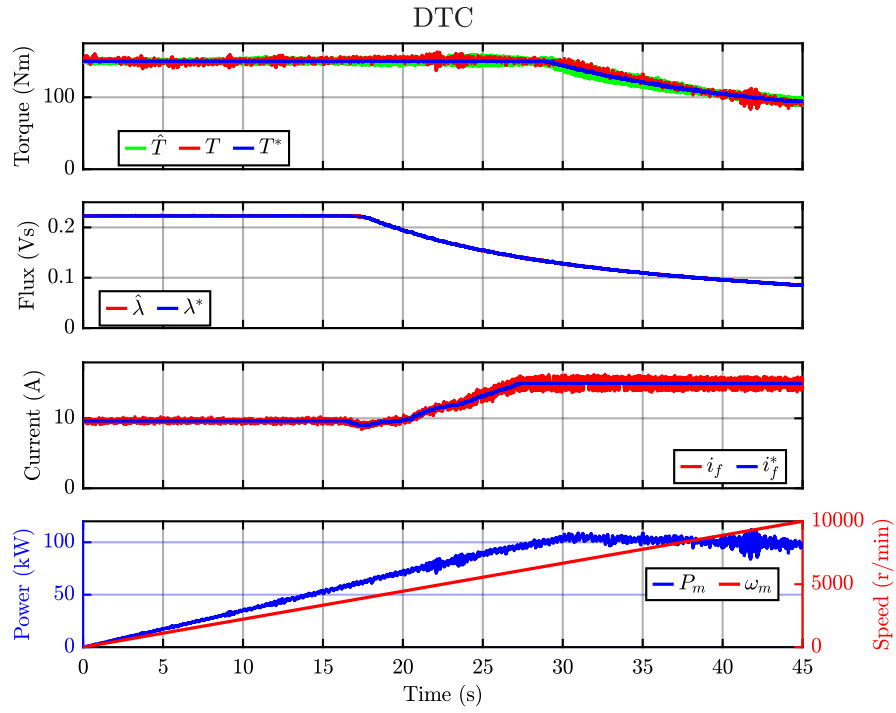
These experimental tests allow a detailed assessment of both steady-state and dynamic behavior of the controllers, while providing a direct comparison of the different TCAs implementations under various operational conditions, as discussed in the following subsections.

5.2.1 Maximum torque per speed (MTPS) profile

To characterize the MTPS performance, the MUT was accelerated from standstill to the mechanical limit of 10000 r/min utilizing the DM as a speed-controlled prime mover. A constant acceleration rate of approximately 230 r/min/s was applied to emulate quasi-steady-state operating conditions, while the torque reference T_{in}^* was set to the maximum test bench limit of 150 Nm.

The experimental trajectories of the regulated state variables for each TCA are presented in Figs. 5.4-5.11. Note that the mechanical power P_m is computed directly from the torque and speed data acquired via the high-precision transducer ($P_m = T \cdot \omega_m$). The results unequivocally demonstrate that the proposed UTC fully exploits the EESM's torque and power capabilities across the entire speed range, regardless of the specific TCA implementation. Furthermore, the regulation of stator and rotor quantities rigorously adheres to the minimum Joule loss trajectory, confirming the architecture's effectiveness in coordinating multiple system constraints while guaranteeing energy-efficient operation.

For clarity, it is highlighted that torque oscillations systematically affecting all TCAs at 20 s, 4500 r/min and at 40 s, 9000 r/min are due to the mechanical torsional resonances of the test rig, thus not depending on the proposed torque controller.

Fig. 5.4 Experimental results: MTPS profile with DFVC, $f_s=10$ kHz.Fig. 5.5 Experimental results: MTPS profile with DTC, $f_s=10$ kHz.

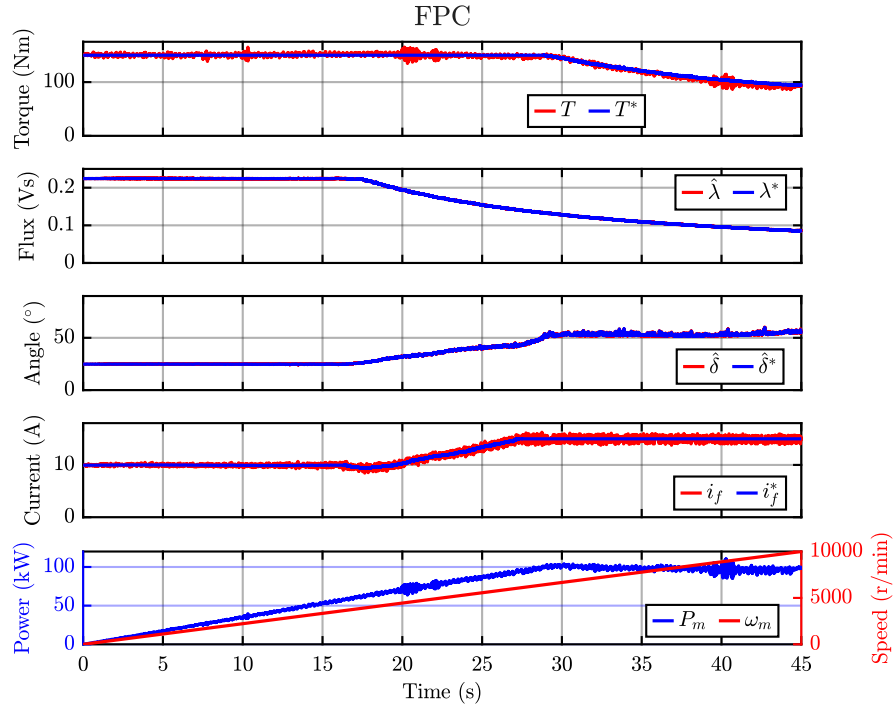


Fig. 5.6 Experimental results: MTPS profile with FPC, $f_s=10$ kHz.

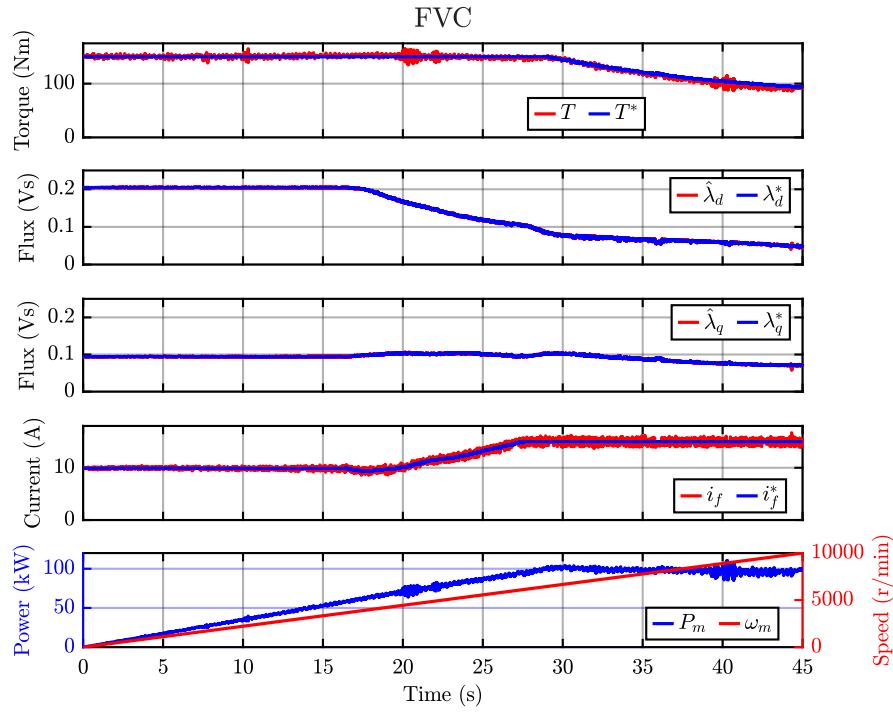
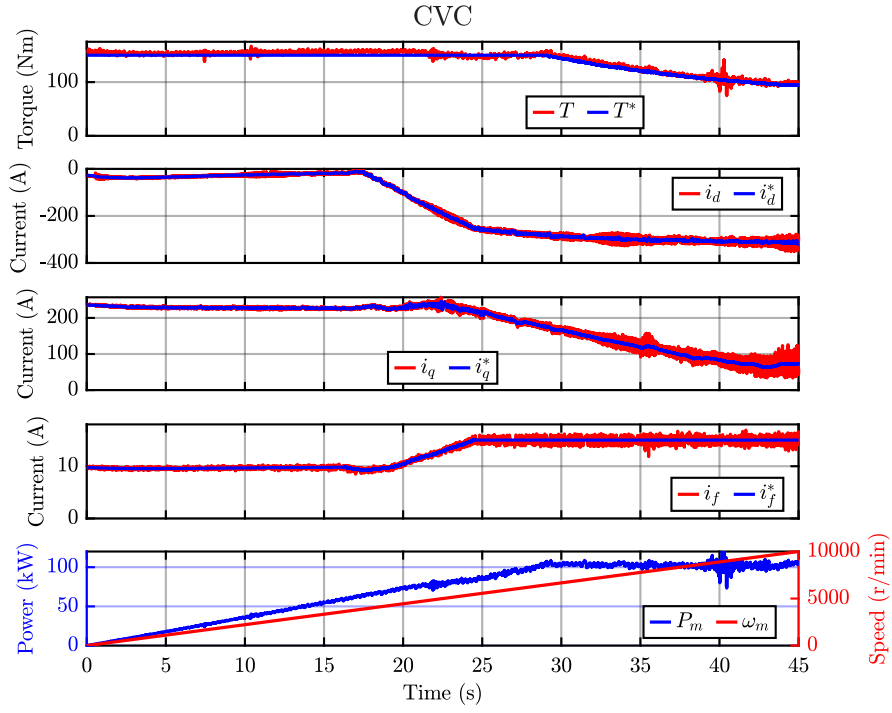
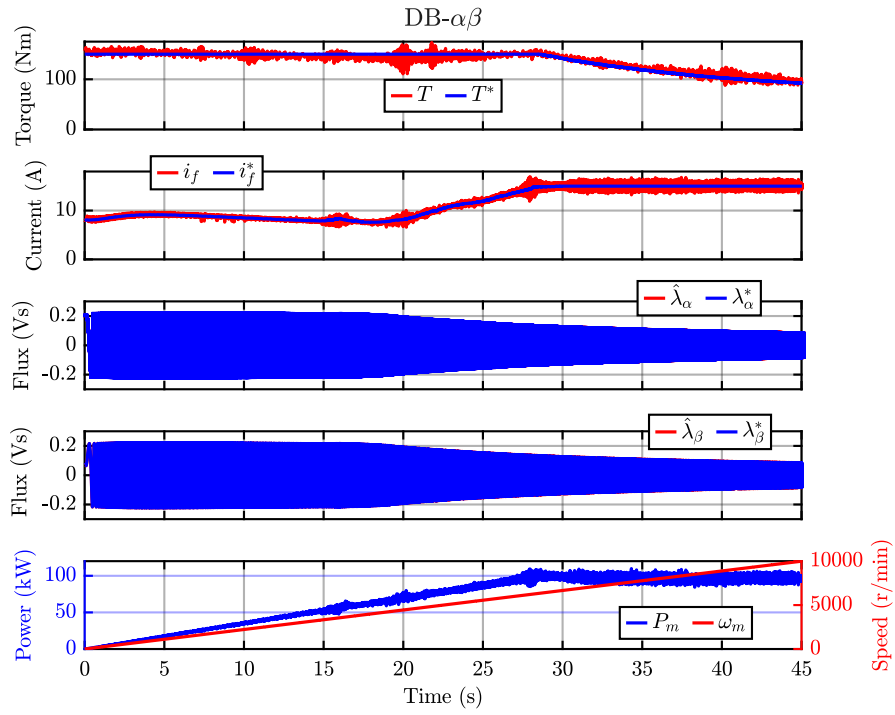


Fig. 5.7 Experimental results: MTPS profile with FVC, $f_s=10$ kHz.

Fig. 5.8 Experimental results: MTPS profile with CVC, $f_s=10$ kHz.Fig. 5.9 Experimental results: MTPS profile with DB- $\alpha\beta$, $f_s=3$ kHz.

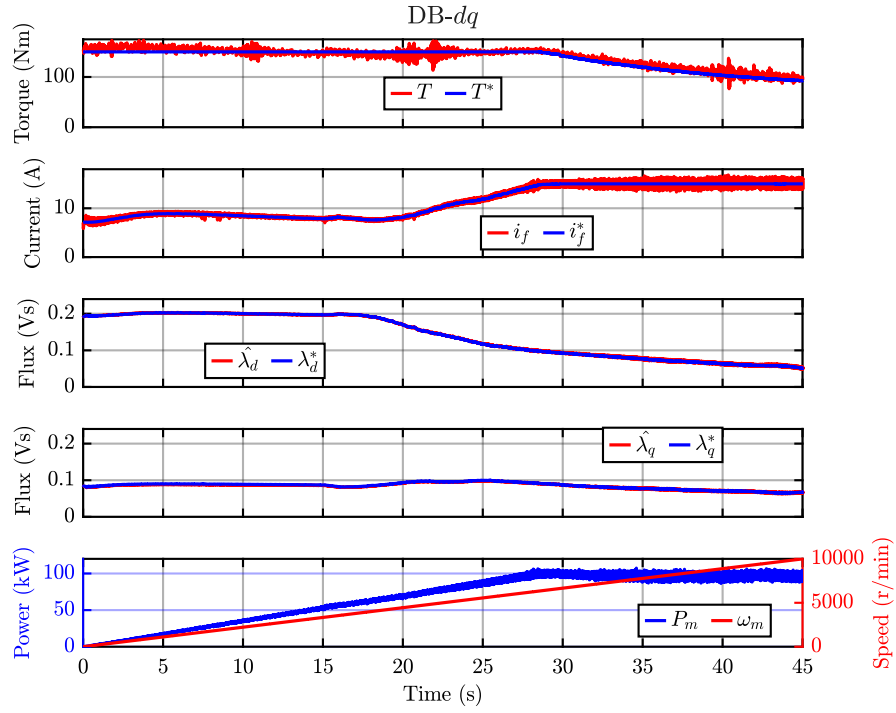


Fig. 5.10 Experimental results: MTPS profile with DB- dq , $f_s=3$ kHz.

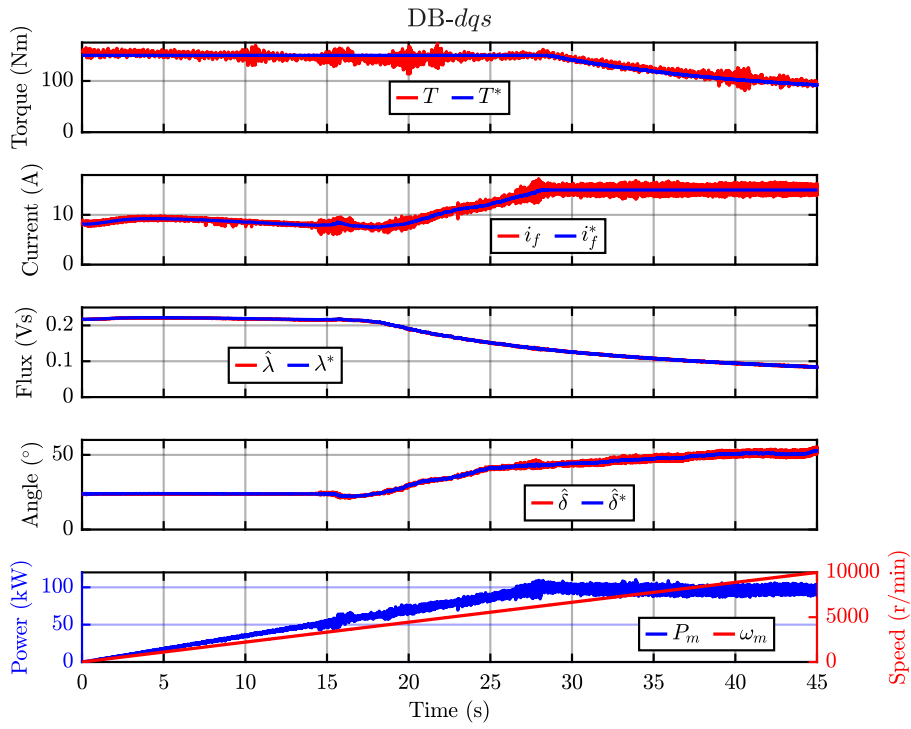


Fig. 5.11 Experimental results: MTPS profile with DB- dqs , $f_s=3$ kHz.

5.2.2 Torque transients

This section focuses on the characterization of the MUT intrinsic dynamic performance. In EESMs, the dynamic torque-generating capability is critically governed by electromagnetic cross-coupling between the stator and rotor circuits. To isolate these physical phenomena and rigorously evaluate the control topology without the limitations imposed by the finite bandwidth of conventional PI regulators, the subsequent experimental analysis is conducted exclusively utilizing the proposed DB-DFC architectures.

The dynamic performance of the proposed torque controllers was evaluated via fast transient response tests at a constant speed of 4000 r/min. Specifically, a torque step command of 100 Nm was applied with a slew rate of 5000 Nm/s.

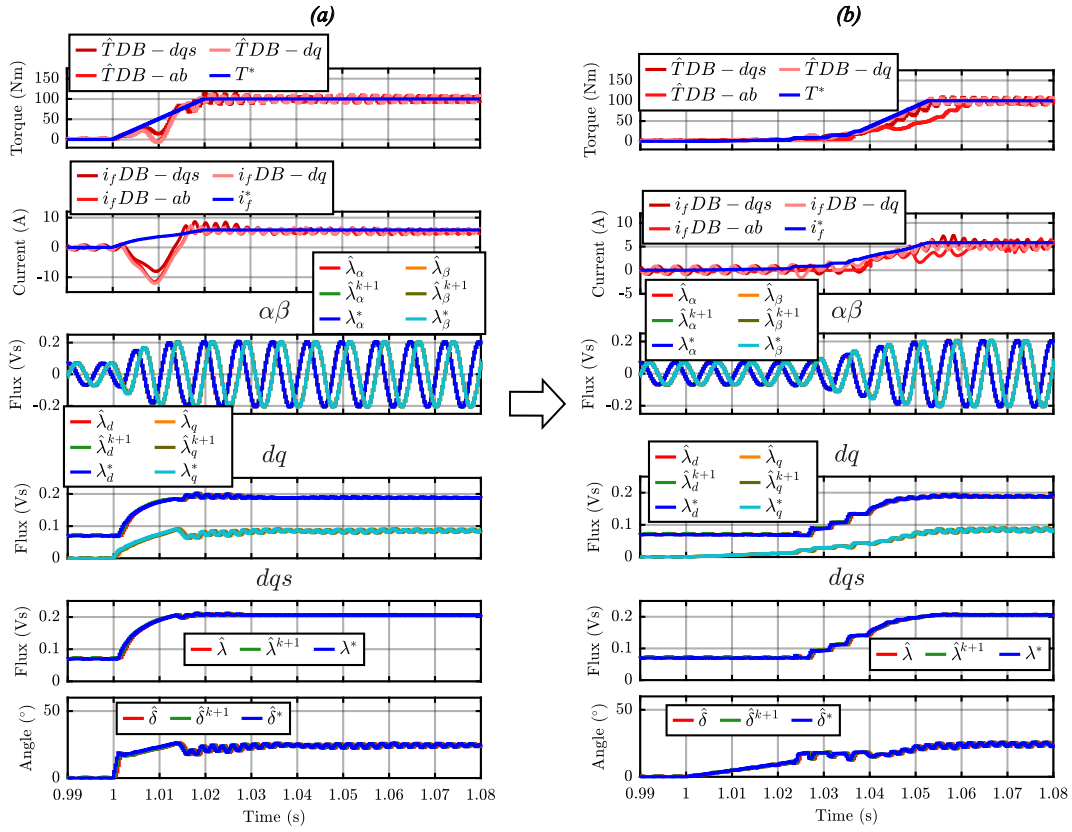


Fig. 5.12 Experimental results of torque transient response from 0 to 100 Nm, 5000 Nm/s, 4000 r/min: (a) without slew rate limitation for $i_f < 1$ A; (b) with slew rate limitation of 150 Nm/s for $i_f < 1$ A.

As illustrated in Fig. 5.12 (a), the controllers effectively track the torque reference; however, the rotor current exhibits a significant undershoot at the onset of the transient. This phenomenon stems from the armature reaction coupling between the stator and rotor windings, which becomes particularly critical at low excitation current levels. Analytically, the rotor current dynamics can be approximated as:

$$i_f = \frac{v_f - \frac{d}{dt}\lambda_f}{R_f} \simeq \frac{v_f - (\frac{di_f}{dt}l_{ff} + \frac{di_d}{dt}l_{fd})}{R_f} \quad (5.1)$$

where l_{ff} denotes the rotor self-differential inductance and l_{fd} represents the cross-coupling differential inductance between the rotor and the stator d -axis. Note that the cross-coupling term with the stator q -axis, l_{fq} (see Fig. 2.34), is negligible in this context [158].

As depicted in Fig. 2.33 and Fig. 2.35, although l_{ff} is approximately two orders of magnitude larger than l_{fd} , the high torque slew rate necessitates a rapid variation in stator flux, inducing a large positive derivative in the stator d -axis current ($\frac{di_d}{dt} \gg 0$, see Fig. 5.13). Under these conditions, the time derivative of the rotor flux $\frac{d\lambda_f}{dt}$ is dominated by the cross-coupling term $l_{fd}\frac{di_d}{dt}$.

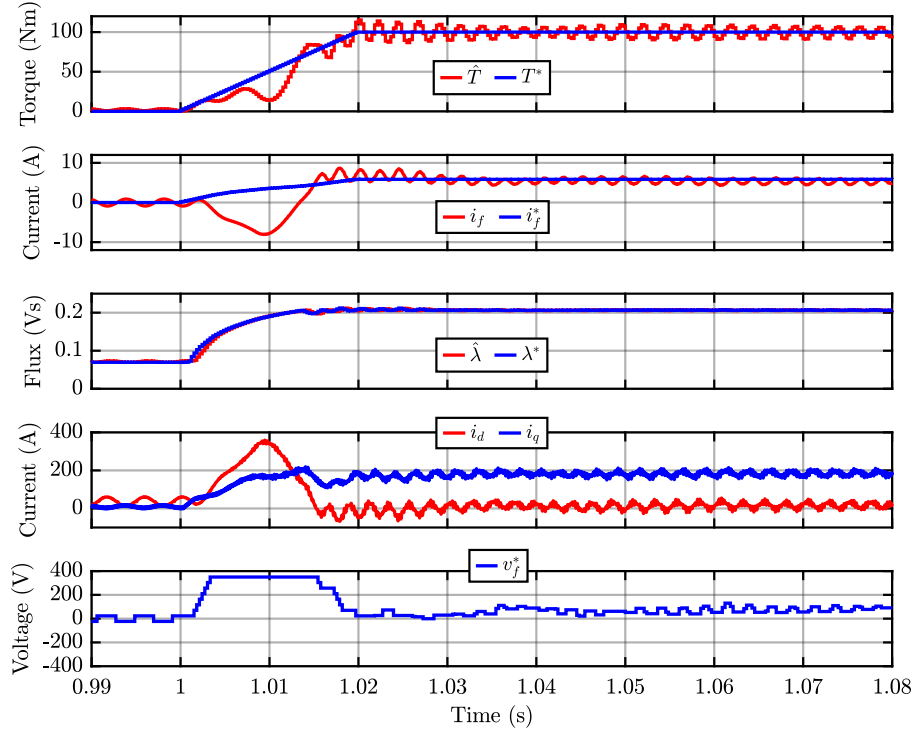


Fig. 5.13 Experimental results of torque transient response from 0 to 100 Nm, 5000 Nm/s, 4000 r/min using DB- dq s: stator dq currents and reference rotor voltage v_f^* .

According to (5.1), this rapid positive variation in i_d induces a substantial rotor back-EMF that opposes the control voltage v_f . Consequently, operating conditions may arise where $|\frac{d\lambda_f}{dt}| > |v_f|$, momentarily driving the rotor current into the negative region. In this scenario, the rotor flux regulator saturates the available voltage in an attempt to restore i_f , leading to a temporary loss of controllability, as shown in Fig. 5.13. To mitigate this effect, an adaptive slew rate limitation of 150 Nm/s is imposed whenever the rotor current falls below 1 A ($i_f \leq 1$ A). The results in Fig. 5.12 (b) confirm that this strategy effectively stabilizes the rotor current without compromising the overall dynamic performance, as the machine still achieves the torque set point in less than 0.06 s. It is noteworthy that the control LUTs are intentionally discretized for positive i_f values only. This design choice ensures that transient rotor current undershoots do not propagate into the stator control maps, preventing potentially destabilizing interactions between the control loops.

Further validation is provided by the step-response test reported in Fig. 5.14, where the torque reference steps from 30 Nm to 100 Nm, bypassing the low-current slew rate limiter. Despite voltage margin constraints preventing a true DB response,

the controller achieves the target torque within the minimum number of sampling periods without overshoot. The decoupled reference structure (T^* for stator, T_f^* for rotor) allows the stator loop to deliver rapid torque response while the rotor current converges according to its slower natural dynamics, optimizing efficiency.

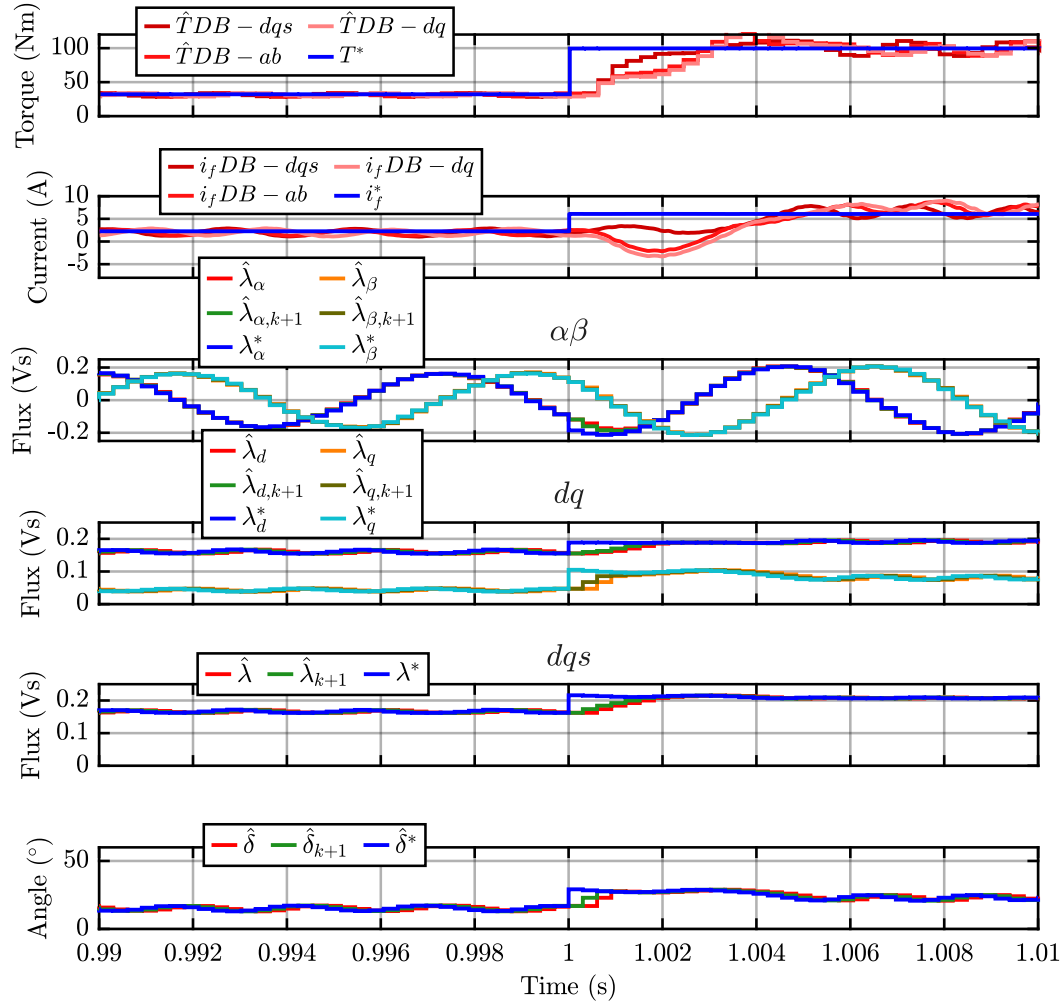
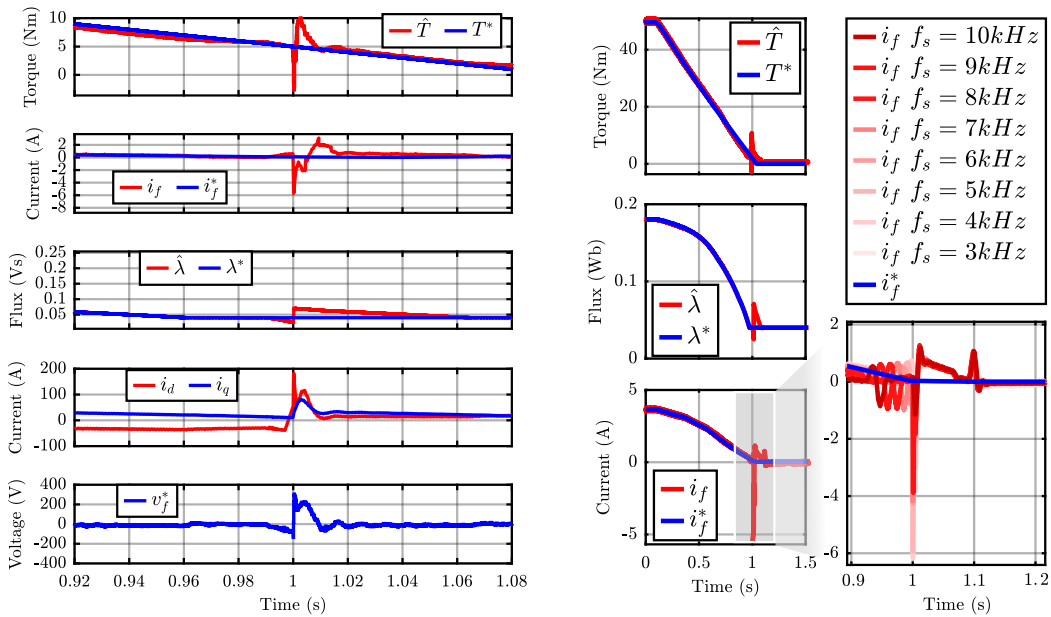


Fig. 5.14 Experimental results of torque step response from 30 to 100 Nm, 4000 r/min.

In Fig. 5.14, minor oscillations in the estimated torque $\hat{T}$ correspond to the transient behavior of the rotor current. Notably, the DB- dqs implementation exhibits a smaller rotor current undershoot compared to alternative DB approaches. This advantage is attributed to the inherent decoupling of the dqs frame, where the stator flux amplitude regulation requires minimal voltage margin (4.51), thereby reducing the cross-coupling interaction between the stator d -axis and the rotor circuit.

The stator-rotor interaction is most critical during torque-zeroing transients, where the stator flux must be reduced to its minimum value ($\lambda_{\min} = 0.04$ Vs), defined to ensure always the correct functioning of the flux observers. This process drives the stator d -axis current toward zero from negative values, resulting in a positive stator current transient $\frac{di_d}{dt}$. According to (5.1), both the control voltage v_f and the induced cross-coupling voltage $l_{fd} \frac{di_d}{dt}$ act in the same direction (negative), potentially driving the rotor current negative if the loop response is insufficient. This phenomenon is illustrated in Fig. 5.15 (a) for a torque reduction from 50 Nm to 0 Nm.



(a) Magnification of stator dq currents and reference rotor voltage v_f^* , $f_s = 3$ kHz.

(b) Effect of sampling frequency on rotor current undershoot.

Fig. 5.15 Experimental results of torque transient from 50 to 0 Nm, 50 Nm/s, 4000 r/min using DB- dqs .

The impact of sampling frequency on this undershoot is analyzed in Fig. 5.15 (b). Increasing the sampling frequency significantly attenuates the undershoot, rendering it negligible at $f_s = 10$ kHz due to the enhanced bandwidth and damping capability of the control loop.

Finally, Fig. 5.16 shows the transient response of the three DB approaches under identical conditions (150 Nm to 0 Nm step, 1000 Nm/s slew rate, $f_s = 10$ kHz).

The results reinforce the superior dynamic performance of the DB- dqs strategy, consistent with the findings from the step response analysis.

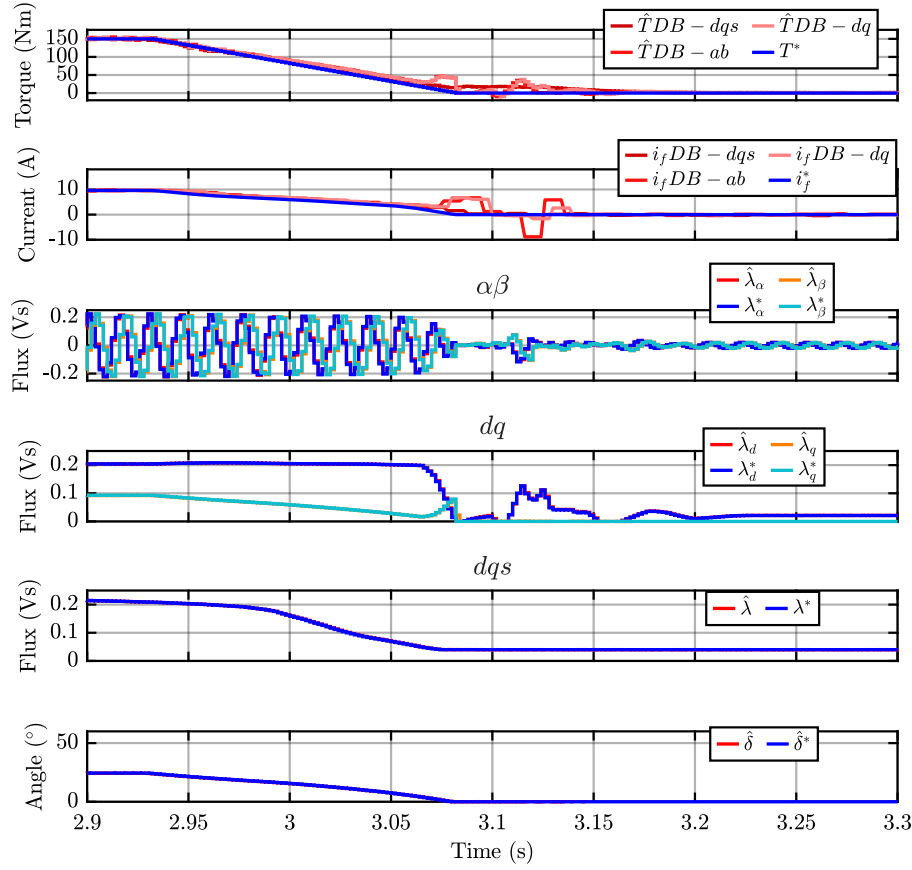


Fig. 5.16 Comparative experimental results of torque transient from 150 to 0 Nm, 1000 Nm/s, 4000 r/min, $f_s = 10$ kHz for different DB implementations (downsampling factor: 20).

5.2.3 Fast torque reversals

To rigorously evaluate the dynamic performance and robustness of the proposed torque controllers, fast torque reversal tests were conducted along the MTPS trajectory. The MUT was accelerated from standstill to the maximum speed of 10000 r/min utilizing the DM as a prime mover, with a constant acceleration of approximately 230 r/min/s. While the reference torque magnitude was maintained at the maximum admissible value ($|T_{in}^*| = 150$ Nm), the torque sign was periodically reversed to induce rapid and repetitive transitions between motoring and generating modes.

For the proposed DB-DFC schemes, the testing protocol specifically investigated the impact of sampling frequency on dynamic capability. At a constrained sampling frequency of $f_s = 3$ kHz, the reversal frequency and torque slew rate were set to 0.2 Hz and 50 Nm/s, respectively. As observed in Fig. 5.22, Fig. 5.24 and Fig. 5.26 the rotor current exhibits temporary, though non-critical, loss of controllability during rapid torque reductions. This phenomenon is consistent with the armature reaction effects previously analyzed during torque-zeroing transients. Conversely, increasing the sampling frequency to typical automotive values ($f_s = 10$ kHz) enabled significantly higher dynamic performance, permitting a reversal frequency of 0.5 Hz and a torque slew rate of 150 Nm/s. Under these conditions, the machine's peak power of 100 kW was reversed in approximately 1.2 s at maximum speed without rotor current instability, as shown in Fig. 5.23, Fig. 5.25 and Fig. 5.27.

The high-dynamic profile (0.5 Hz, 150 Nm/s) was successfully executed by all flux-based strategies (DB-DFC, DFVC, DTC, FPC, FVC). In contrast, the CVC strategy required a specific derating of the test parameters due to its inherently limited small-signal bandwidth, a consequence of the strong cross-coupling between the stator d -axis and the rotor circuit. To ensure stable operation, the CVC test protocol was restricted to a reversal frequency of 0.2 Hz, with a torque slew rate linearly decreasing from 150 Nm/s at standstill to 25 Nm/s at 10000 r/min (see Fig. 5.21).

The experimental results, illustrated in Figs. 5.17-5.27, evaluate both motoring and braking performance, confirming that the proposed UTC maintains rigorous regulation of all stator and rotor state variables across the entire speed range that includes the deep flux-weakening region. The absence of overcurrent events or instabilities during these severe dynamic stress tests validates the robustness of the proposed control scheme. Nonetheless, as illustrated in Fig. 5.21, the CVC strategy exhibits a noticeably reduced torque dynamic response, justified by its limited control bandwidth.

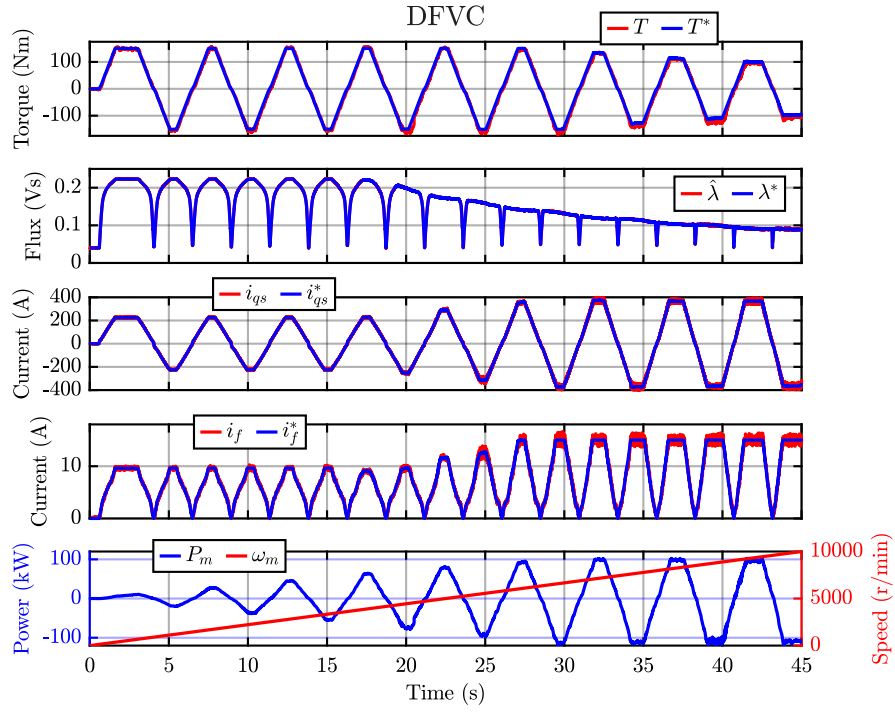


Fig. 5.17 Experimental results of fast torque reversals along the MTPS profile with DFVC, $f_s=10$ kHz.

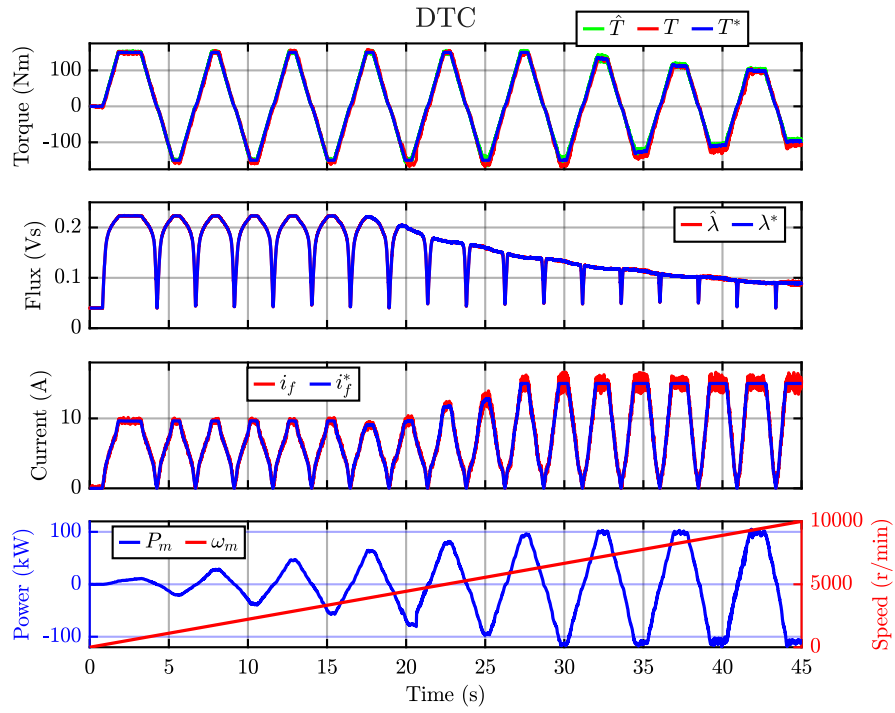


Fig. 5.18 Experimental results of fast torque reversals along the MTPS profile with DTC, $f_s=10$ kHz.

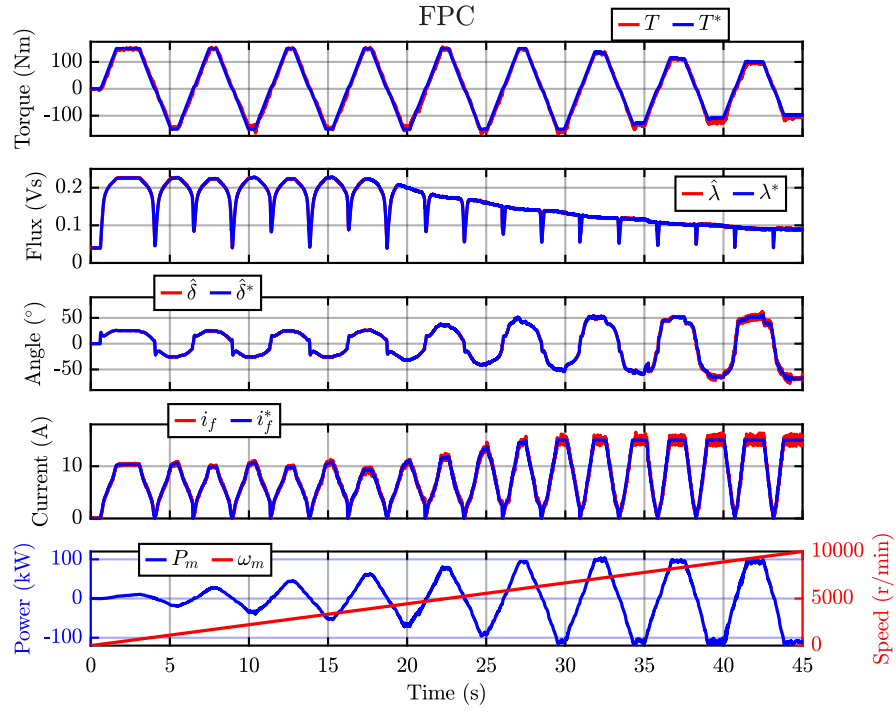


Fig. 5.19 Experimental results of fast torque reversals along the MTPS profile with FPC, $f_s=10$ kHz.

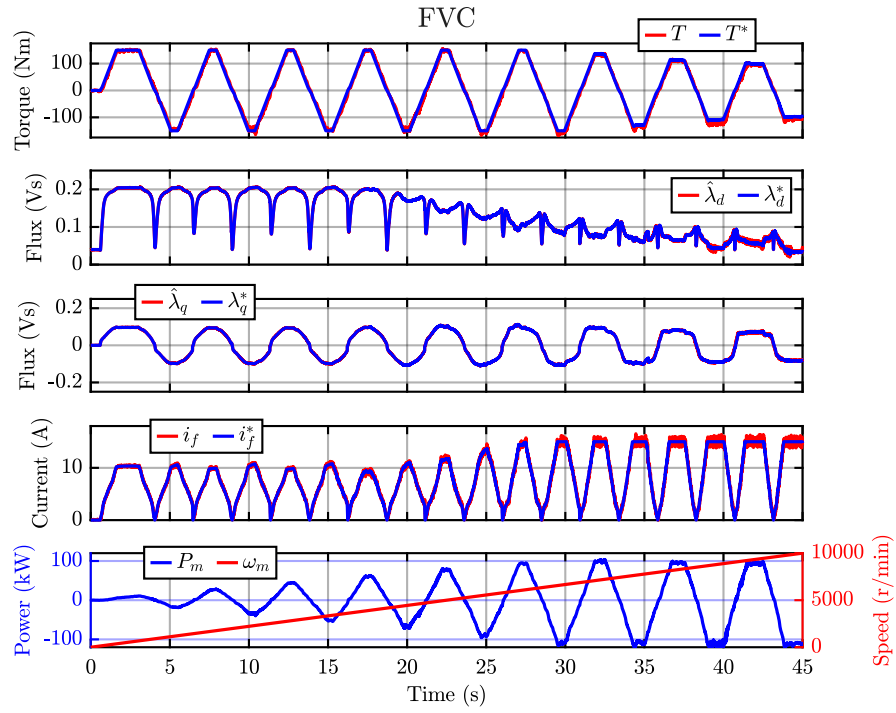


Fig. 5.20 Experimental results of fast torque reversals along the MTPS profile with FVC, $f_s=10$ kHz.

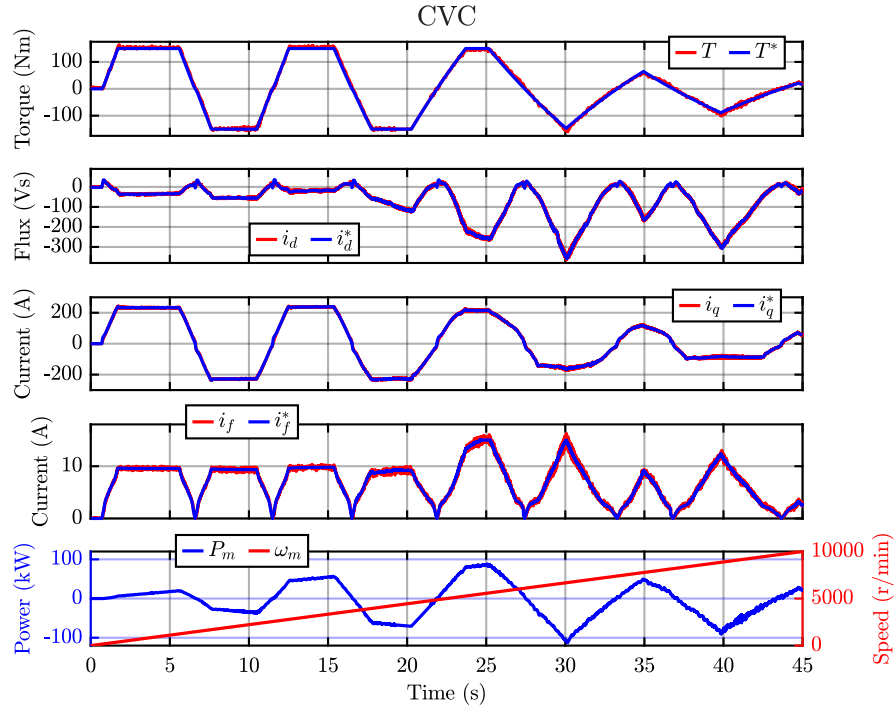


Fig. 5.21 Experimental results of fast torque reversals along the MTPS profile with CVC, $f_s=10$ kHz.

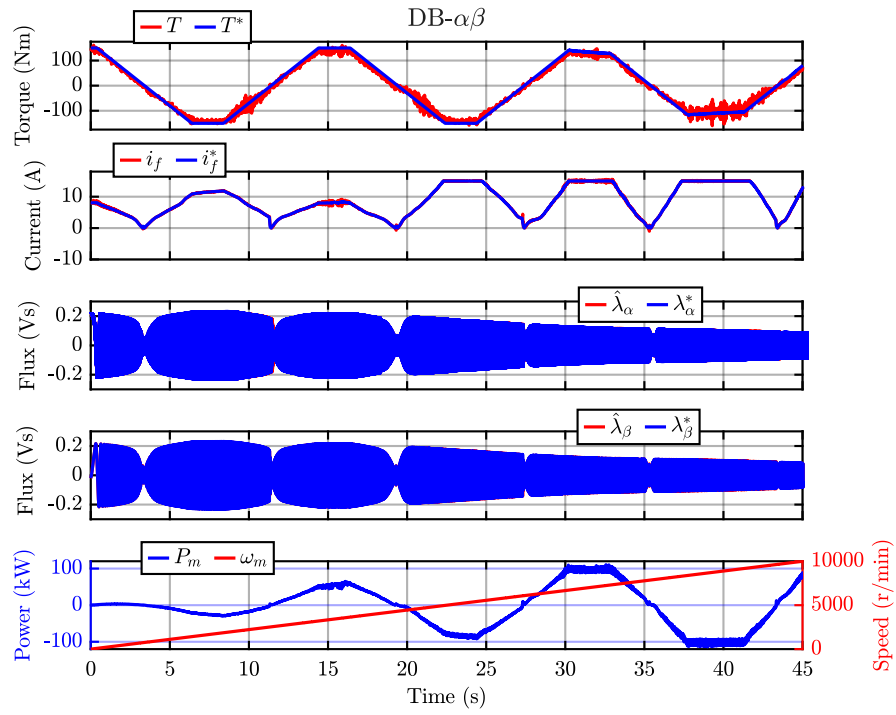


Fig. 5.22 Experimental results of fast torque reversals along the MTPS profile with DB- $\alpha\beta$, $f_s=3$ kHz.

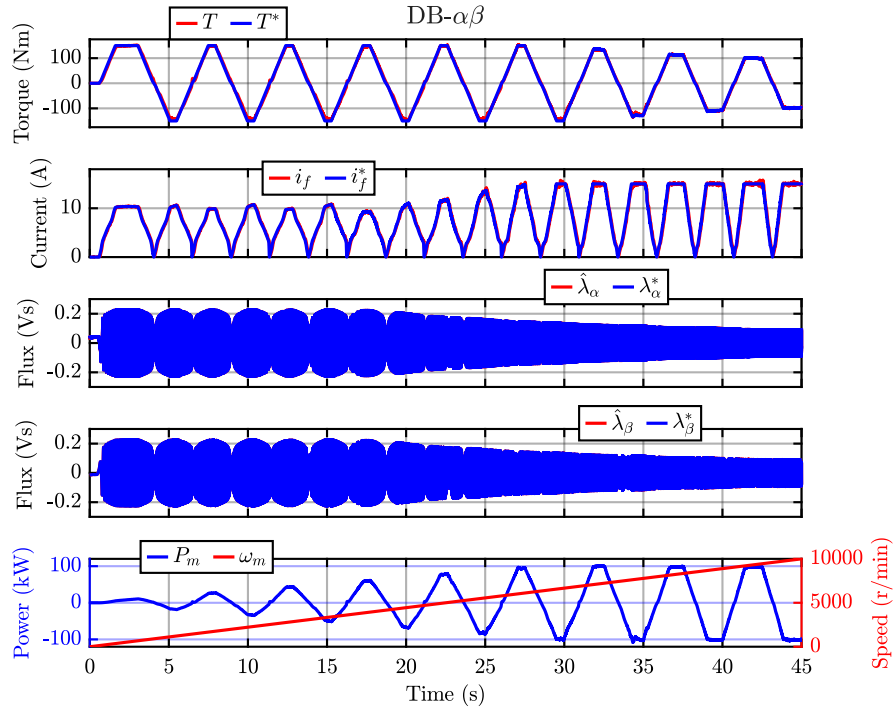


Fig. 5.23 Experimental results of fast torque reversals along the MTPS profile with DB- $\alpha\beta$, $f_s=10$ kHz.

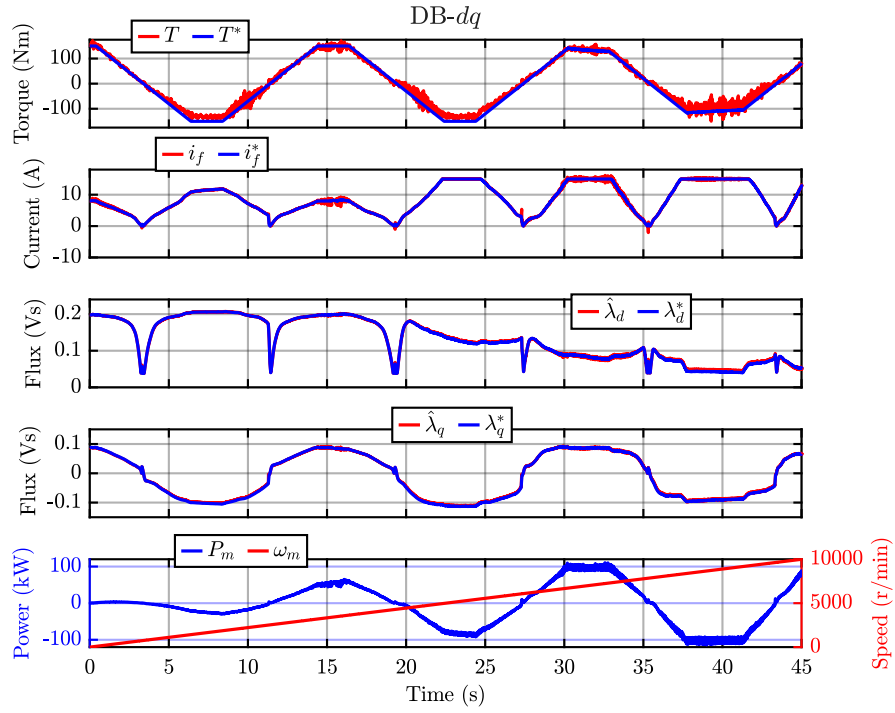


Fig. 5.24 Experimental results of fast torque reversals along the MTPS profile with DB- dq , $f_s=3$ kHz.

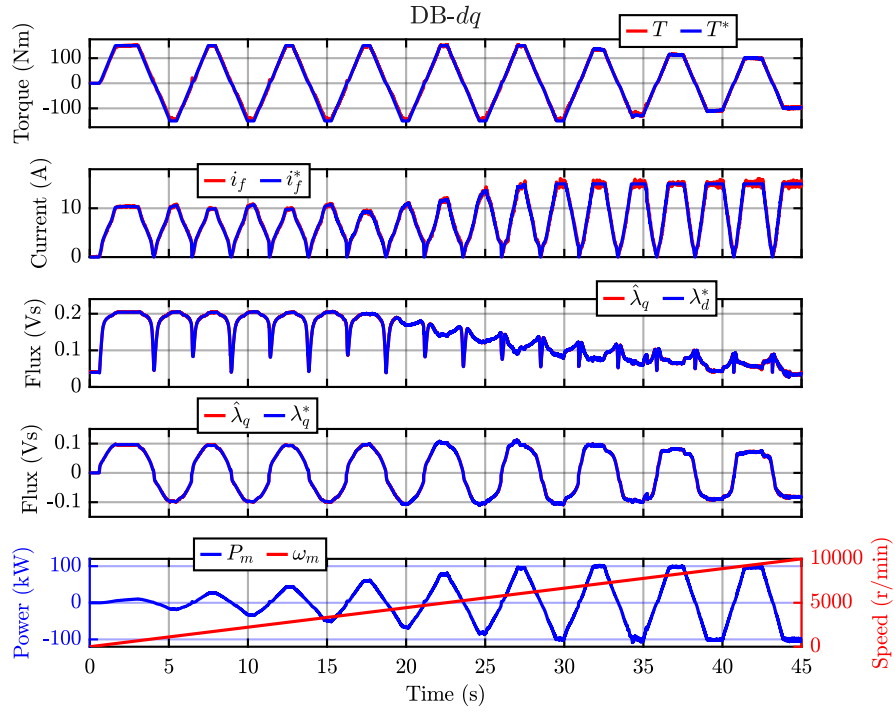


Fig. 5.25 Experimental results of fast torque reversals along the MTPS profile with DB- dq , $f_s=10$ kHz.

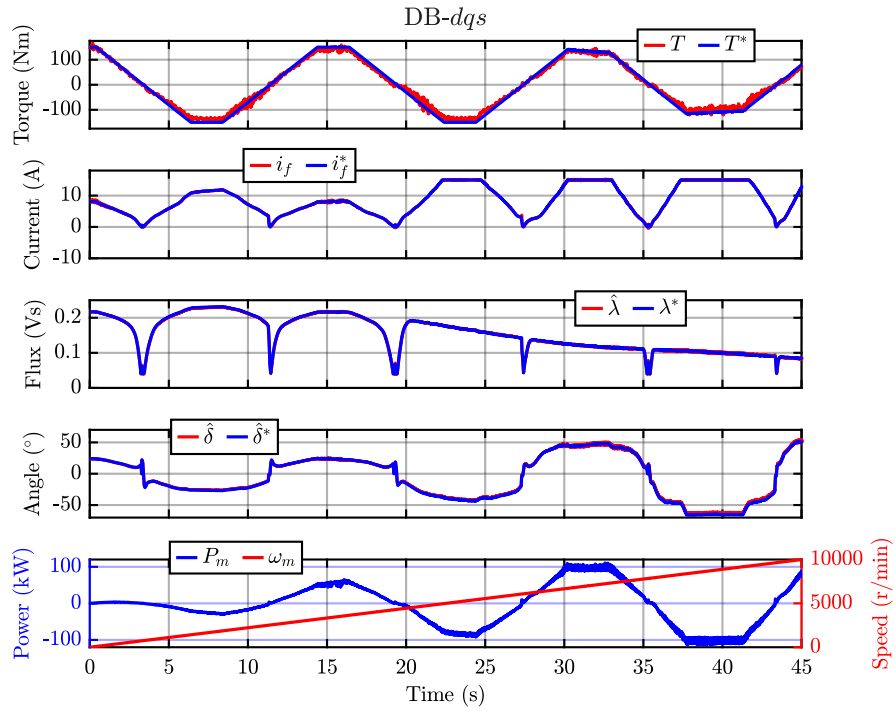


Fig. 5.26 Experimental results of fast torque reversals along the MTPS profile with DB- dqs , $f_s=3$ kHz.

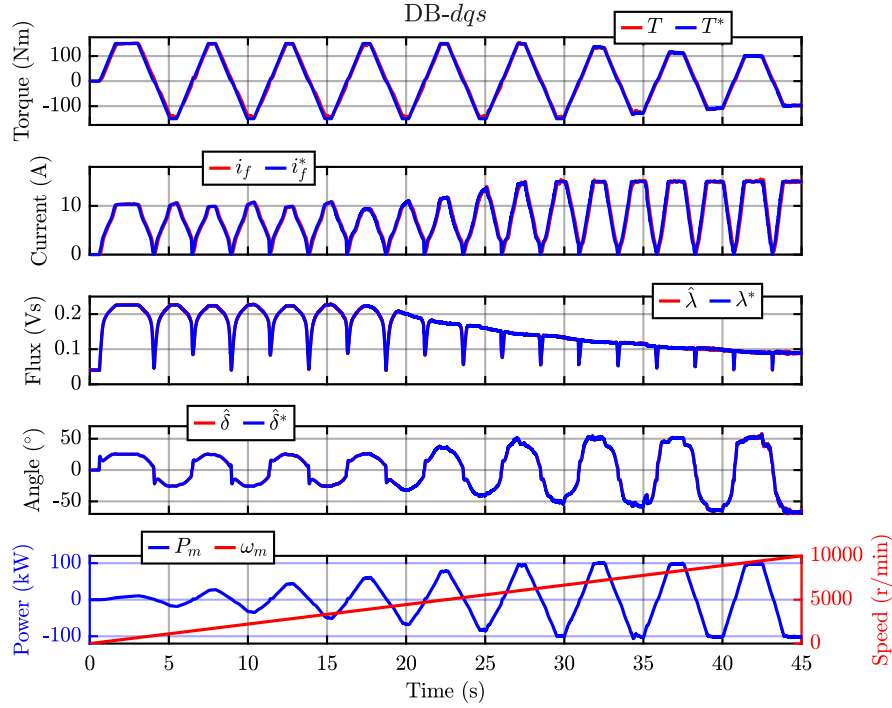


Fig. 5.27 Experimental results of fast torque reversals along the MTPS profile with DB-dqs, $f_s=10$ kHz.

5.2.4 Steady-state torque accuracy

The steady-state torque regulation accuracy of the proposed control architecture was extensively validated through a comprehensive mapping characterization.

The experimental characterization comprised a grid sweep across the MUT operating range, incrementing the rotational speed from standstill to the 10000 r/min rig limit in discrete 500 r/min steps. At each speed step, ten equidistant torque setpoints were applied, spanning from zero to the maximum admissible torque in motoring mode $T_{\max}$ according to the mechanical speed.

The experimental results for the PI-based TCAs, presented in Figs. 5.28-5.32, demonstrate that the torque error ΔT , evaluated both in absolute terms and as a percentage of $T_{\lim}$, remains consistently within acceptable bounds, despite the absence of explicit iron and mechanical loss compensation in the λ^* and i_f^* references. However, a comparative analysis reveals a distinct performance divergence above the base speed. Specifically, the CVC strategy exhibits a marked degradation in torque accuracy compared to the flux-based methodologies (see Fig. 5.32). This deviation

derives from the unmodeled influence of frequency-dependent iron losses, which decouple the dq current references from the actual torque-producing components. Conversely, the flux-based strategies (DFVC, DTC, FPC, and FVC) demonstrate superior robustness. This superior performance is attributed to their reliance on the direct regulation of stator flux quantities (amplitude or vector), whose estimation is driven by the Voltage Model (VI) of the flux observer (see Fig. 4.3). Since the VI estimator naturally integrates the back-EMF, it implicitly captures the effects of iron losses, thereby enhancing torque accuracy without requiring explicit compensation mechanisms.

Regarding the proposed DB-DFC strategies, no significant performance discrepancies were observed among the varying coordinate implementations ($\alpha\beta, dq, dqs$), as steady-state accuracy is primarily dictated by the common stator flux observer. Consequently, the analysis focuses on the DB- dqs implementation as the representative case.

At a sampling frequency of 3 kHz, a non-negligible degradation in torque regulation accuracy is observed (see Fig. 5.33 (a)), which is commensurate with the characteristic degradation of controllability associated with a limited sampling-to-fundamental frequency ratio. As this ratio falls below 10 at maximum speed, the system becomes increasingly susceptible to discretization errors, elevated current ripple, and flux observer estimation inaccuracies. Conversely, the torque regulation precision of the DB-DFC is significantly enhanced when the sampling frequency is increased to a typical operating value of 10 kHz, as illustrated in Fig. 5.33 (b).

This high degree of accuracy is ensured by the implementation of integral correction loops for the torque ΔT_{adj}^* and rotor current $\Delta i_{f,adj}^*$ references (described in Chapter 4.2.4), as depicted in Fig. 5.33.

The subsequent analysis focuses exclusively on the controller behavior at a 10 kHz sampling frequency. The derated controllability at 3 kHz, despite errors remaining below 10% of the maximum torque, complicates the isolation of specific torque error sources. Analyzing the compensator outputs at 10 kHz reveals the distinct error sources across the speed range:

- **Low Speed:** The torque adjustment ΔT_{adj}^* is significant, required primarily to compensate for inverter voltage errors and dead-time effects not fully captured in the predictive equations (4.44)–(4.47).

- **Medium Speed:** The adjustment magnitude decreases, as voltage errors become negligible and iron losses remain low, indicating that the baseline control maps are well-tuned.
- **High Speed:** The torque adjustment increases again to compensate for iron losses, utilizing the feedback provided by the VI flux estimator.

Regarding the rotor current regulation, the adjustment $\Delta i_{f,adj}^*$ remains limited across the entire speed range. Specifically, the maximum adjustment is consistently less than 4%, which is primarily attributed to the detuning of the FEA-based current-to-flux maps utilized to derive the rotor flux reference from the corresponding rotor current (see Fig. 2.23).

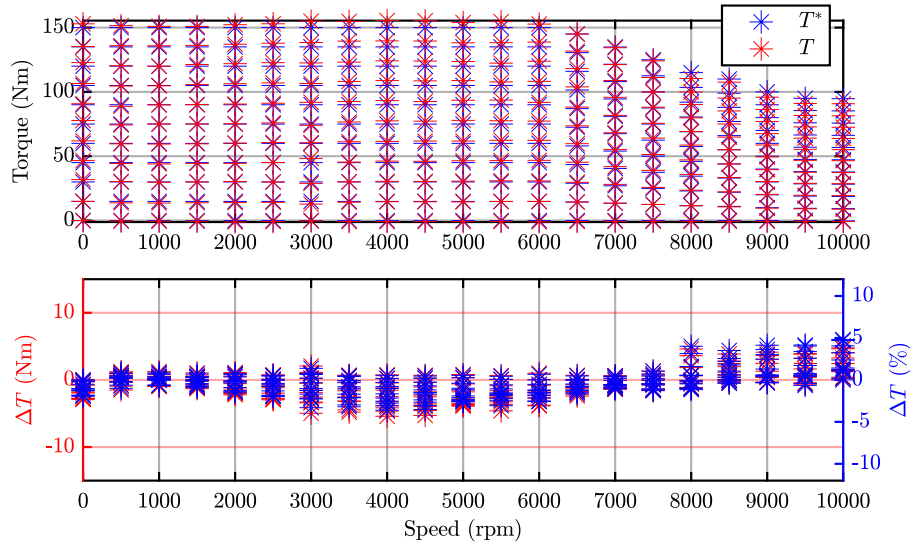


Fig. 5.28 Steady-state torque accuracy with DFVC: experimental results.

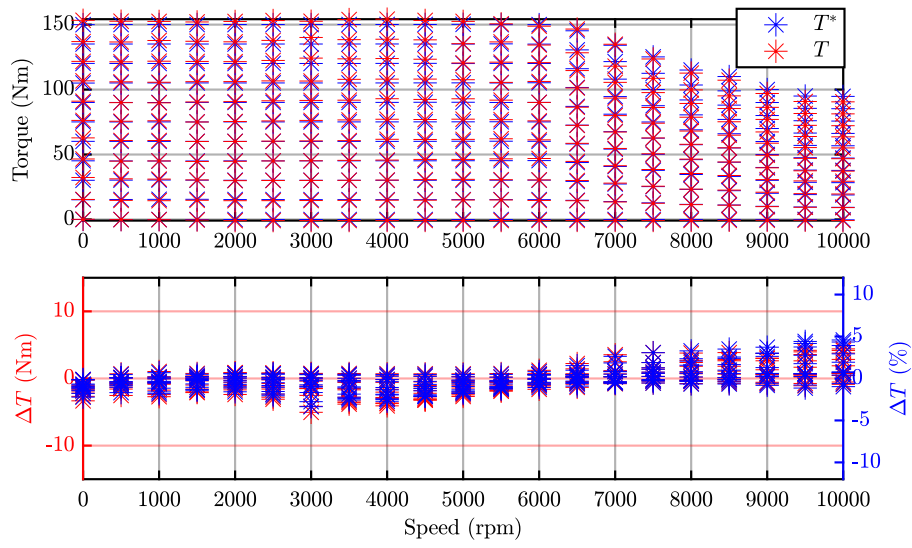


Fig. 5.29 Steady-state torque accuracy with DTC: experimental results.

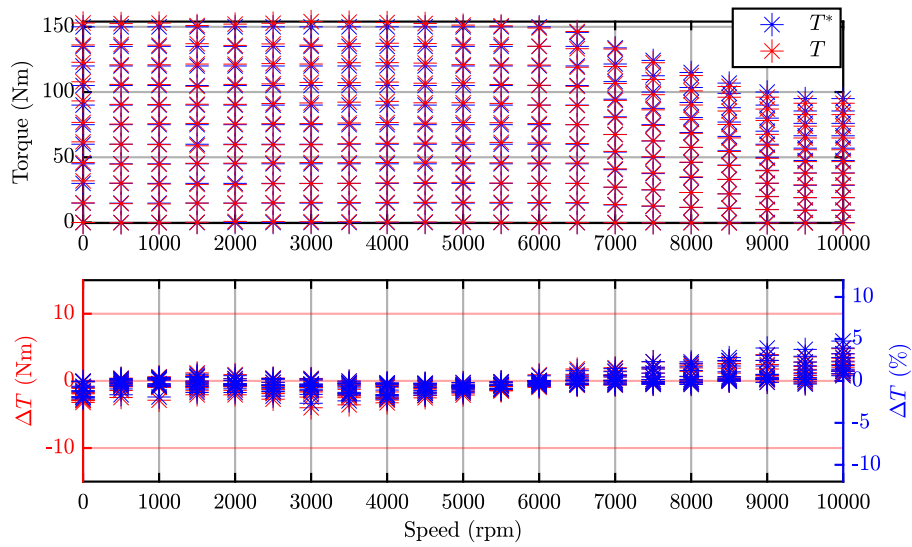


Fig. 5.30 Steady-state torque accuracy with FPC: experimental results.

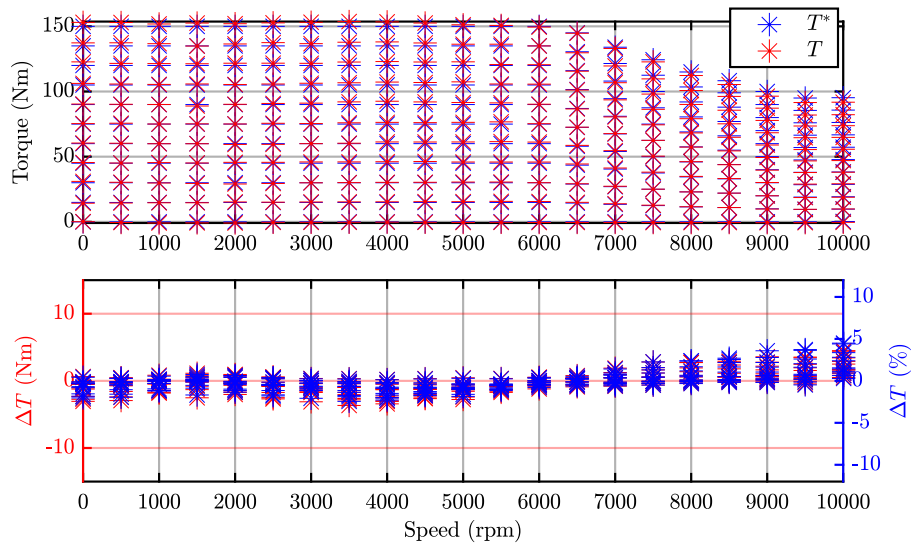


Fig. 5.31 Steady-state torque accuracy with FVC: experimental results.

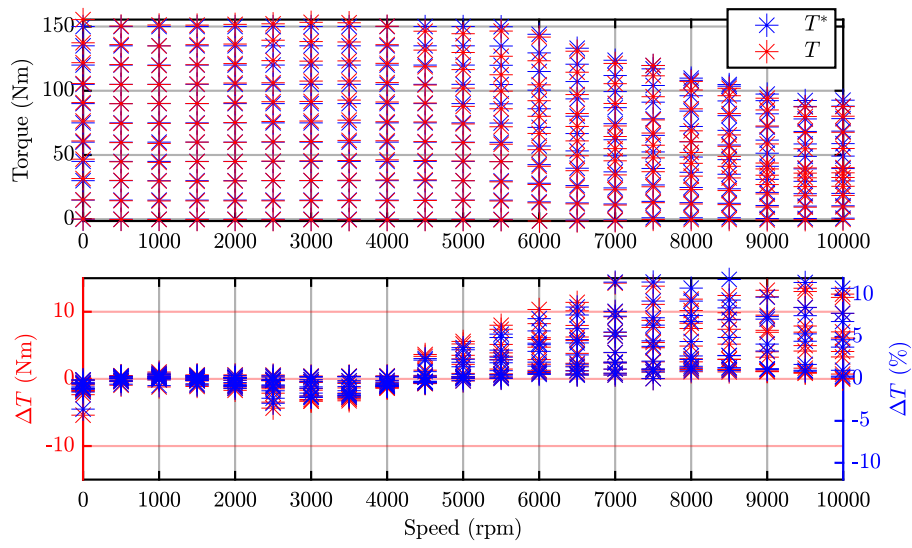


Fig. 5.32 Steady-state torque accuracy with CVC: experimental results.

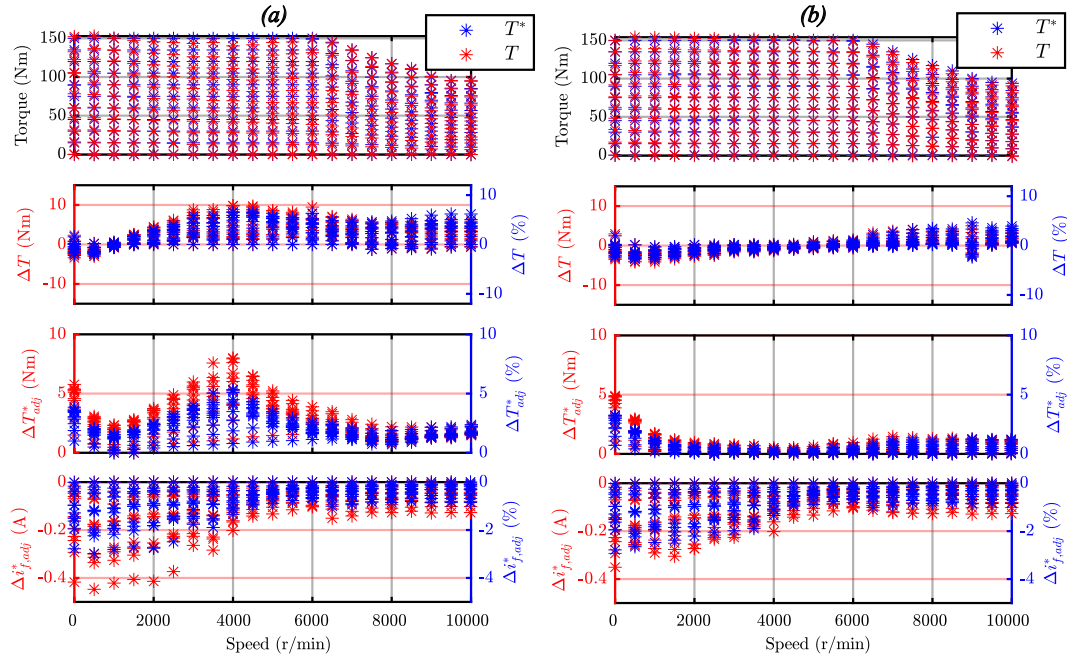


Fig. 5.33 Experimental results of steady-state torque accuracy using DB-dqs: (a) $f_s = 3$ kHz; (b) $f_s = 10$ kHz.

5.2.5 Variable current limits

A distinguishing feature of the proposed UTC is its inherent capability to accommodate advanced thermal management strategies reported in the literature. This is achieved by dynamically adapting the stator i_{lim} and rotor $i_{f,\text{lim}}$ current limits online. To validate this functionality, a specific dynamic derating test was performed. As this feature of the developed torque controller is independent from the TCA employed, the DFVC is taken as reference in this test campaign. The MUT was operated at a constant speed of 500 r/min imposed by the DM, with the torque reference T_{in}^* set to the maximum capability of 150 Nm. Initially, the current limits were fixed at their peak values as per Table 2.1 ($i_{\text{lim}} = i_{\text{max}} = 385$ A, $i_{f,\text{lim}} = i_{f,\text{max}} = 15$ A).

The test sequence is described as follows (see Fig. 5.34):

- Rotor derating ($t = 7.5$ s): The rotor current limit $i_{f,\text{lim}}$ is ramped down from 15 A to 5 A at a rate of 2 A/s. Despite the reduction in the rotor current, the controller maintains the output torque at 150 Nm. This is achieved by the torque control algorithm, which automatically compensates for the reduced i_f

by increasing the stator flux amplitude reference λ^* and adapting the torque-producing current i_{qs}^* according to (2.11). Although the theoretical torque limit T_{lim} decreases, it remains sufficiently high to satisfy the requested load torque.

- Stator derating ($t = 17.5$ s): While maintaining the reduced rotor current limit, the stator current limit i_{lim} is reduced from 385 A to 150 A at a rate of 15 A/s. In this regime, the combined constraints make the initial torque setpoint unachievable. Consequently, the controller actively clamps the reference torque T^* to the instantaneous limit T_{lim} . Even under these heavily constrained conditions, the algorithm ensures the exploitation of the maximum feasible torque while strictly adhering to the minimum copper loss trajectory.

In summary, this test demonstrates the controller's ability to seamlessly integrate with higher-level thermal supervision strategies, ensuring the active protection of the machine windings while maximizing performance within the available thermal range.

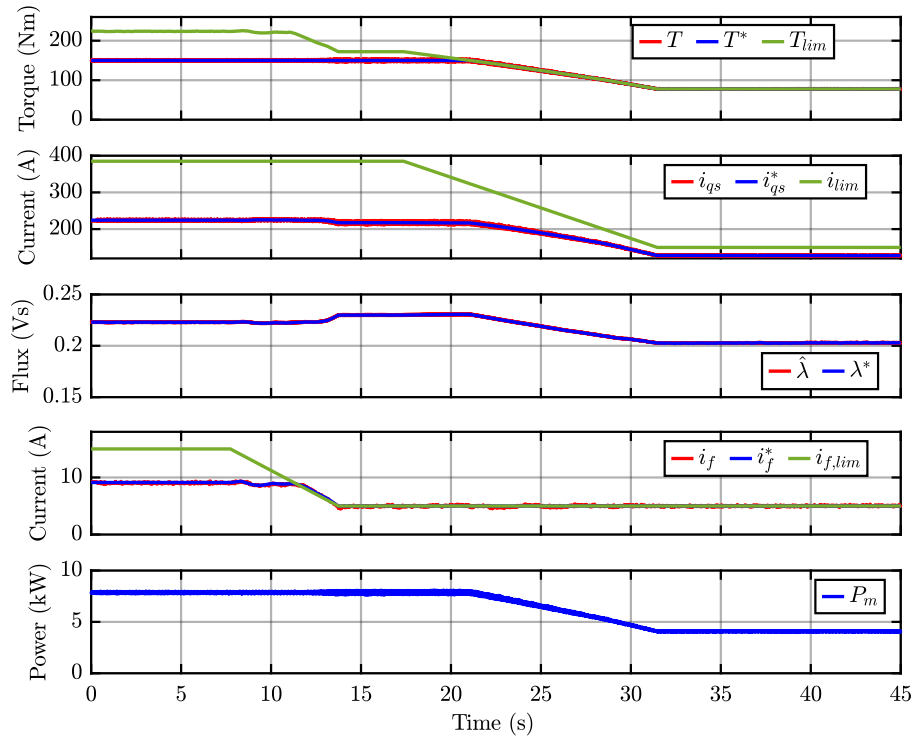


Fig. 5.34 Experimental validation of online current limit adaptation (thermal derating emulation).

5.2.6 MTPV limitation

To rigorously evaluate the effectiveness of the MTPV limitation strategy, a specific test scenario was designed to force the machine into deep flux-weakening operation where stability is critical. The rotor current limit $i_{f,\text{lim}}$ was forced to 2.5 A. This reduction lowers the machine's characteristic current to approximately 200 A, a value significantly below the active stator current limit ($i_{\text{lim}} = i_{\text{max}} = 385$ A), thereby exposing the MTPV asymptote within the achievable operating range. Under these constrained conditions, an MTPS trajectory was executed in motoring mode. The torque reference T_{in}^* was set to 100 Nm, consistent with the reduced excitation torque capability. The speed was ramped from standstill to 10000 r/min at a rate of 230 r/min/s utilizing the DM.

The stability of the MTPV loop relies on the tuning of the load angle limiting PI regulator. The calibration followed a heuristic design criterion aimed at ensuring spectral separation between the outer limit loop and the inner control dynamics. Specifically, the proportional $k_{p,\text{MTPV}}$ and integral $k_{i,\text{MTPV}}$ gains were refined experimentally to achieve an optimal trade-off between prompt intervention capability and noise rejection on the torque regulation.

The experimental results, illustrated in Fig. 5.35, demonstrate the effectiveness of the proposed strategy. As the speed exceeds 4500 r/min, the controller detects the approach to the stability boundary. Consequently, the load angle limiter (see Fig. 4.6) activates, generating a torque correction term ΔT_{MTPV} . This action dynamically reduces the effective torque reference T^* , ensuring that the observed load angle $\hat{\delta}$ remains strictly bounded by the theoretical limit δ_{lim} . The test confirms that the control architecture successfully prevents the EESM from exceeding the pull-out torque, maintaining stable regulation throughout the deep flux-weakening and MTPV region.

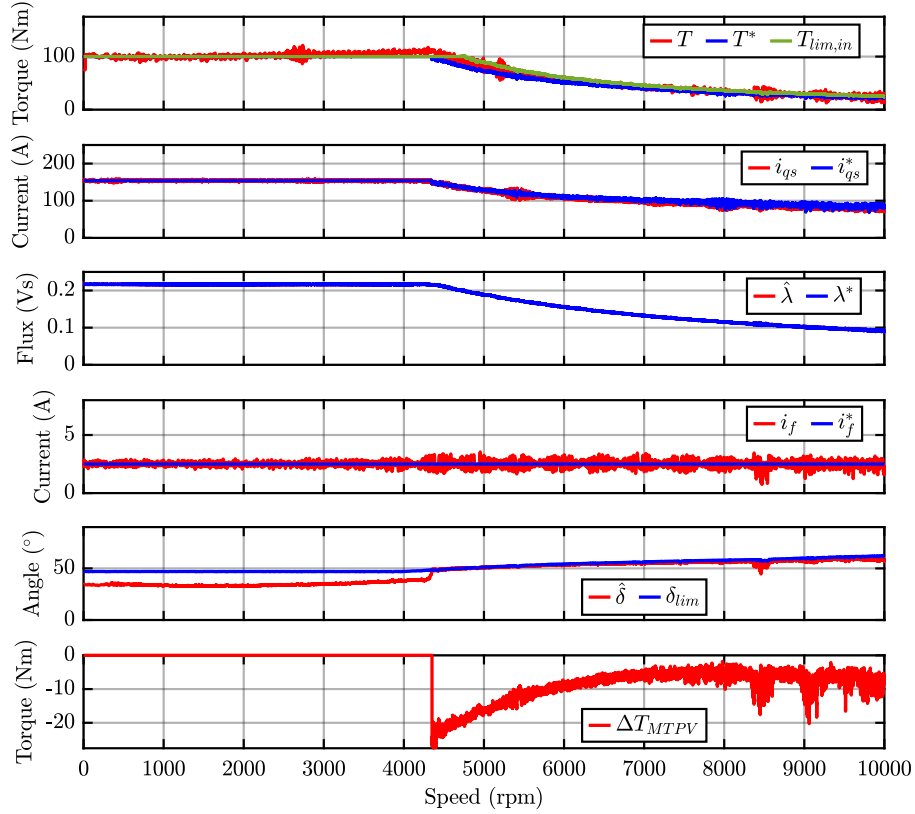


Fig. 5.35 Experimental validation of the MTPV limitation strategy (DFVC implementation).

5.2.7 Efficiency mapping

As previously discussed, the developed UTC framework is feasible for the implementation of arbitrary efficiency strategies. However, within the scope of this work and the presented experimental validation, the focus is restricted to the minimization of total Joule losses. To verify that the pre-processing phase for the definition of the control LUTs was correctly executed in accordance with the identified flux maps, the machine efficiency map has been experimentally characterized using the measurements acquired during the steady-state torque accuracy tests. For each tested point, the electric power is estimated from (4.9) and the mechanical power is calculated multiplying the measured torque by the measured speed $P_m = T \cdot \omega_m$. The total power losses are straightforwardly derived as:

$$P_{l,tot} = P_e - P_m. \quad (5.2)$$

Accordingly, the efficiency η is calculated:

$$\eta = \frac{P_m}{\underbrace{P_m + P_{l,tot}}_{P_e}} \quad (5.3)$$

Experimental results are compared with predictions derived from FEA-based results [152]. The FEA mapping algorithm evaluates machine performance across the torque-speed domain by iteratively calculating feasible operating points, determining associated loss components, and identifying optimal trajectories according to the defined efficiency objective.

As shown in Fig. 5.36 and Fig. 5.37, the total power loss $P_{l,tot}$ and efficiency η profiles exhibit an almost perfect match with the FEA, thereby validating the fidelity of the control LUT generation process. Furthermore, the experimental data reported in Fig. 5.37 confirm the high-efficiency operation achieved by the implemented minimum-Joule-loss approach.

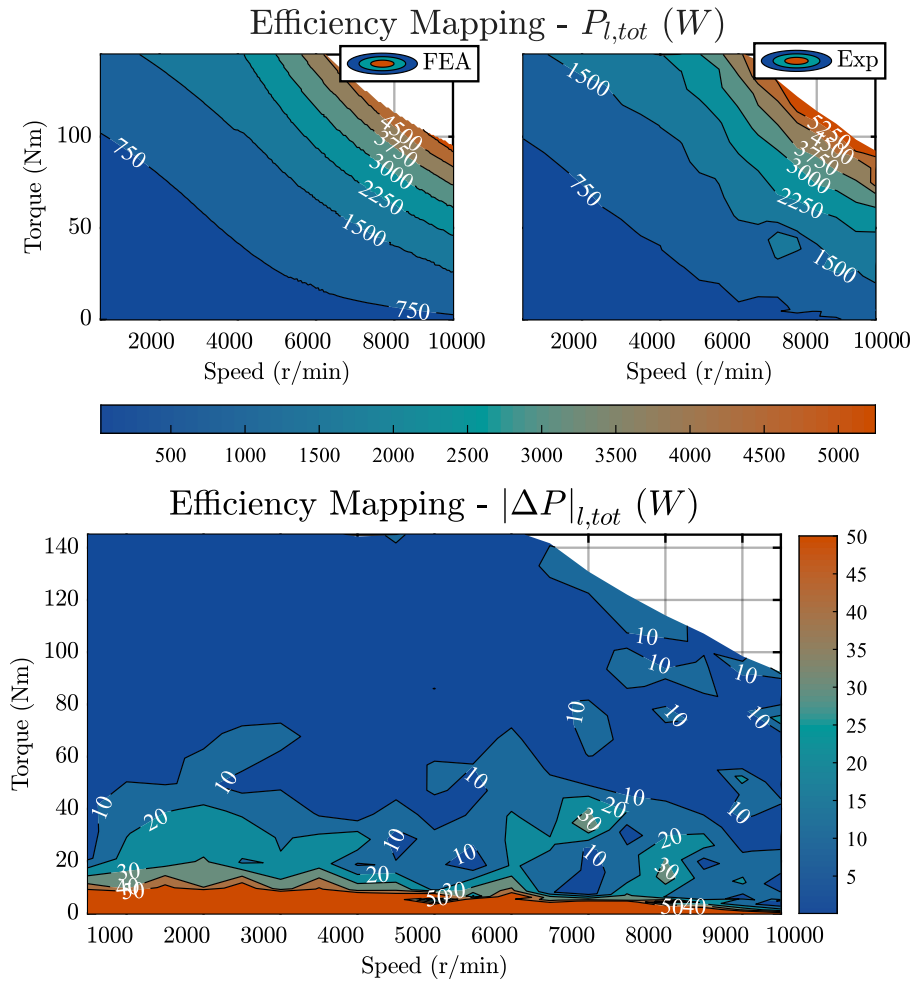


Fig. 5.36 Comparison between experimentally measured and FEA predicted total power losses $P_{l,tot}$.

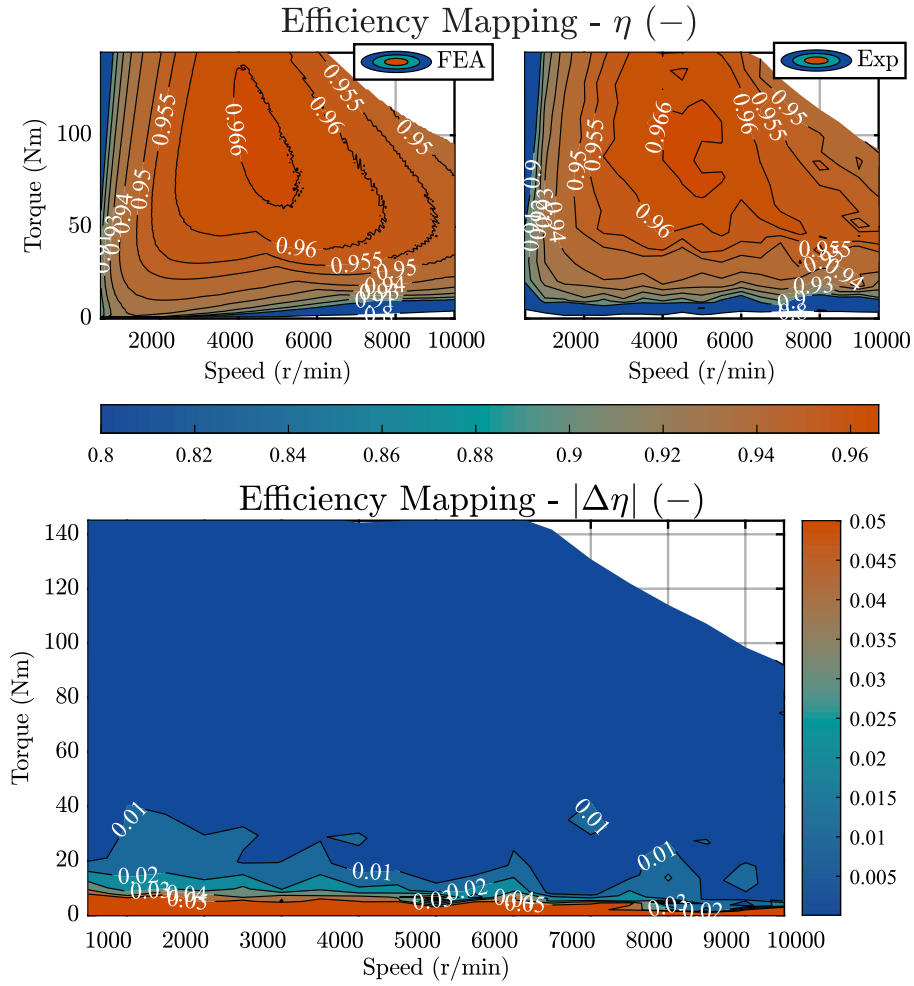


Fig. 5.37 Comparison between experimentally measured and FEA predicted efficiency η .

5.3 Summary

This chapter provided experimental evidence of the effectiveness and robustness of the proposed torque control architectures for automotive EESMs. The key findings from the experimental campaign are summarized below:

- **Setup and methodology:** The controllers were validated on a high-fidelity test bench up to 10000 r/min, 150 Nm and 100 kW. To strictly evaluate the "plug-and-play" capabilities of the DB-DFC, the control LUTs were derived solely from FEA data rather than experimental identification, and the sampling

frequency was intentionally limited at 3 kHz to show that torque control can be implemented with low sampling frequencies.

- **Static performance (MTPS):** All tested strategies (DFVC, DTC, FPC, FVC, CVC and DB-DFC) successfully exploited the MUT maximum torque and power capabilities across the full speed range. The proposed UTC ensured that stator and rotor quantities adhered to the minimum copper loss trajectories.
- **Dynamic performance:**
 - Flux-based approaches demonstrated superior dynamic performance compared to PI-based CVC, which suffered from bandwidth limitations due to stator-rotor cross-coupling.
 - A critical phenomenon of rotor current undershoot during fast torque transients was identified, caused by the armature reaction ($l_{fd} \frac{di_d}{dt}$). An adaptive slew-rate limiter was successfully implemented to mitigate this effects without compromising global performance.
 - Among the predictive strategies, the DB-*dqs* implementation exhibited the best dynamic behavior and stability, particularly during fast torque reversals in the deep flux-weakening region.
- **Steady-state accuracy:** The proposed flux-based controllers maintained high torque accuracy across the operating envelope. In contrast, the CVC strategy exhibited degradation at high speeds due to unmodeled iron losses. The DB-DFC proved robust to parameter detuning, due to FEA maps implementation, thanks to the integral correction loops, achieving high precision even at low sampling-to-fundamental frequency ratios, with optimal performance observed at standard automotive switching frequencies (10 kHz).
- **Thermal derating:** The flexibility of the UTC was confirmed through the successful validation of online thermal derating (variable i_{lim} and $i_{f,lim}$) and active MTPV limitation. The controller seamlessly adapted to dynamic constraints, ensuring safe operation without exceeding the machine's pull-out torque.

In conclusion, the experimental results validate the proposed UTC, especially with the DB-DFC approaches, as a high-performance, robust, and calibration-efficient solution suitable for next-generation electric traction drives. The most

relevant experimental results presented have been published in the journal articles (J2), (J3) and (J4).

Chapter 6

Conclusions and future works

This dissertation was focused on the modeling, simulation, and control of EESMs tailored for automotive traction applications. A key point of this research was the rigorous definition and integration of the magnetic cross-coupling phenomena between the stator and rotor circuits, identifying this interaction as a critical factor in both high-fidelity modeling and torque control performances.

The primary conclusions drawn from this work are summarized below, categorized by the specific research domain.

6.1 Conclusions

6.1.1 Experimental identification of stator and rotor flux linkages

The experimental identification of high-fidelity, multidimensional current-to-flux magnetic maps is one of the contributions of this work. While stator flux characterization leverages adapted PMSM-based protocols, the concurrent identification of rotor flux linkage is indispensable for accurately modeling the dynamic behavior of traction EESMs. Traditional methodologies, specifically locked-rotor tests using square-wave excitation, are largely inapplicable to high-performance traction designs. The combination of low stator inductance and significant armature reaction in such machines poses a risk to hardware integrity due to excessive current and torque transients.

To circumvent these limitations, a robust three-stage rotor flux identification sequence was proposed:

1. Rotor flux linkages are derived through time-domain integration of the rotor back-EMF during continuous sinusoidal excitation.
2. A PSO algorithm is implemented to eliminate fictitious hysteresis loops arising from thermal resistance drift and inverter non-linearities.
3. The integration constant $\lambda_{f,0}$ is identified by leveraging magnetic symmetry during open-rotor stator current reversals.

Validation against FEA benchmarks confirms the high fidelity of the 3D maps across the complete (i_d, i_q, i_f) operating range. Furthermore, experimental validation using a DB rotor flux control scheme demonstrates that these identified maps enable high-precision tracking, maintaining a residual rotor current error of less than 3%. These surfaces allow for the extraction of variable differential inductance profiles, providing the essential parameters for adaptive regulator tuning and saturation-aware torque control.

6.1.2 Advanced modeling and digital twin

To address the limitations of existing modeling approaches, this work proposed a novel Full-VBR model for EESMs. By leveraging multidimensional current-to-flux maps, the model accurately accounts for nonlinear magnetic saturation and, crucially, the dynamic cross-coupling effects between stator and rotor windings.

Key achievements in this domain include:

- **Analytical stator leakage inductance identification:** A novel procedure was developed to calculate the stator leakage flux and the effective current ratio between stator and rotor, overcoming the simplifications of standard identification methods.
- **Accuracy:** Compared to state-of-the-art solutions, the proposed VBR formulation is the first of its kind capable of accurately replicating armature reaction dynamics and stator leakage behavior under deep saturation.

- **Validation:** The model underwent rigorous validation through a "digital twin" approach. Comparison with experimental data from a commercial automotive EESM confirmed the model's high accuracy across all operating regimes, including the critical open-rotor circuit scenario.

These considerations and the results reported in this thesis regarding the proposed digital twin model have been published in the journal article (J1).

6.1.3 Unified torque control architecture

The dissertation introduced a UTC, a versatile control architecture designed to optimize performance and efficiency, whose key features are summarized hereafter.

- **Versatility:** The UTC features a modular structure compatible with multiple TCAs, including CVC, DTC, DFVC, FPC, FVC, and DB-DFC.
- **Efficiency:** The controller inherently performs accurate torque regulation across the entire torque-speed range while minimizing global copper losses (stator plus rotor) via an offline optimization included into LUTs.
- **Thermal integration:** The control scheme supports dynamic thermal management strategies by accepting variable stator and rotor current limits (i_{lim} , $i_{f,lim}$), allowing for seamless integration of higher-level thermal supervisors.
- **Accuracy and derating capability:** The torque regulation presents a competitive accuracy with all the considered TCAs. Furthermore, torque regulation accuracy was preserved even under significant dynamic reductions of the current limits, validating the controller's robustness for real-world thermal derating scenarios.

6.1.4 Comparative analysis of TCAs

Extensive experimental validation highlighted distinct performance characteristics among the implemented TCAs:

- **CVC limitations:** The CVC strategy exhibited significant performance degradation during fast transients due to the strong cross-coupling between the stator

d -axis and rotor excitation regulators. Furthermore, unmodeled iron losses at high speeds resulted in reduced steady-state torque accuracy compared to flux-based strategies.

- **MTPV management:** While DTC and DFVC require auxiliary control modules (often heuristically tuned) to limit the load angle during MTPV operation, strategies like FPC and FVC ensure this limitation inherently.
- **Calibration efficiency:** FPC, FVC, and DB-DFC strategies eliminate the need for inner-loop calibration while maintaining robust performance. This represents a significant advantage in reducing the time and cost associated with calibration procedures. A pivotal characteristic for industrial applications.

These analysis and the most relevant results reported in this thesis regarding the UTC and the dynamic thermal management strategies integration have been published in two journal articles (J2) and (J3).

6.1.5 Dead-beat direct flux control (DB-DFC)

The research culminated in the proposal of a "plug-and-play" DB-DFC methodology, implemented in $\alpha\beta$, dq , and dqs coordinates.

- **Robustness:** The DB-DFC approach demonstrated high torque regulation accuracy and minimized Joule losses without requiring inner-loop tuning. The controller maintained stability and performance even at extremely low sampling-to-fundamental frequency ratios (< 10).
- **Transients handling:** Experimental analysis of fast torque transients revealed that low sampling frequencies combined with high slew rates increase the risk of temporary rotor current uncontrollability due to the armature reaction and the exciter voltage constraints. However, results confirmed that under typical automotive sampling and switching frequencies, this effect is effectively mitigated.

The description of the DB-DFC methodology with the torque transient analysis has been published in the journal article (J4).

6.2 Future works

Future developments of this research will focus on enhancing both the modeling fidelity and the control capabilities of EESM drives.

6.2.1 Advanced modeling

Further refinements to the VBR model will include:

- **Iron loss modeling:** Implementation of frequency-dependent skin effect and iron loss phenomena to provide a precise estimation of overall machine efficiency in the simulation environment.
- **Spatial harmonics:** Integration of the functional dependency of fluxes, inductions, and torque on the rotor mechanical position to capture torque ripple and spatial harmonic effects.

6.2.2 Advanced control

Potential enhancements to the UTC include:

- **Real-time thermal optimization:** Incorporating variable stator and rotor winding temperatures into the loss minimization algorithm to optimize efficiency under conditions deviating from the rated thermal state.
- **Active thermal limiting:** Developing an integrated closed-loop thermal controller that actively limits current references based on real-time temperature feedback.
- **Total loss minimization:** Expanding the reference generation LUTs to account for iron and mechanical losses, thereby minimizing the total drive losses rather than copper losses alone.
- **Overmodulation:** Extending the direct flux control range into the deep overmodulation region (up to six-step operation) to maximize power delivery for a given DC-link voltage.

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