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Variable Flux Permanent Magnet Machines for Sustainable Traction Electrification

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Declaration

I hereby declare that, the contents and organization of this dissertation constitute my own original work and does not compromise in any way the rights of third parties, including those relating to the security of personal data.

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2026

* This dissertation is presented in partial fulfillment of the requirements for **Ph.D. degree** in the Graduate School of Politecnico di Torino (ScuDo).

I would like to dedicate this thesis to my loving parents

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Abstract

Variable Flux Machines (VFMs) are a promising solution for improving the sustainability and operating flexibility of electric drives, particularly in traction applications. By exploiting low-coercivity permanent magnets, VFMs allow their magnetization state (MS) to be intentionally adjusted during operation, enabling adaption of the PM flux linkage according to torque and speed requirement. This feature can extend the operating speed range, improve efficiency over the driving cycle and reduce the dependence on heavy rare earth permanent magnets. However, these advantages are accompanied by significant challenges in machine design, electromagnetic modeling and control, mainly due to the irreversible demagnetization and remagnetization behavior of low-coercivity magnets and the high-amplitude current pulses required for MS manipulation.

This dissertation addresses these challenges through a framework covering machine geometry selection, control-oriented modeling, optimal strategy for MS regulation while maintaining smooth torque control and operating range evaluation under inverter voltage constraints. First, different rotor topologies for VFMs employing only low-coercivity magnets are comparatively analyzed. The results show that the Surface Permanent Magnet (SPM) topology provides the best compromise in terms of remagnetization current demand, robustness against unintended demagnetization, and torque controllability during MS variation transients. Based on this outcome, an SPM Proof-of-Concept (POC) machine is designed as the main reference case, and a traction-oriented VFM using FeN magnets is also developed for traction application case analysis.

The development of VFM control strategies requires extensive simulation stages. Compared to standard PMSMs, the VFM models inherently need to dynamically account for PM demagnetization. This is traditionally achieved by co-simulation merging Finite Element Analysis (FEA) software into the simulation environment,

with significant computational burden, time and storage demand, incompatible with control development and calibration purposes. To this end, a control-oriented electromagnetic model of the VFM is proposed. A baseline formulation is first developed to represent irreversible demagnetization and remagnetization through the use of magnetization maps, flux maps, and an MS updating mechanism. The model is subsequently extended to include asymmetric magnetization under load by introducing the complex magnetization state, an armature-flux formulation, and a two-virtual-magnet representation. [The proposed modeling methodologies are validated against FEA-based simulations and implemented in a real-time Hardware-in-the-Loop \(HiL\) environment, showing high accuracy together with a significantly improved computational effort.](#)

On the control side, two dedicated strategies are proposed to mitigate torque pulsation during MS variation under load. The first one is developed within Direct Flux Vector Control framework, while the second one is based on a modified Field Oriented Control structure, referred to as the $i_d - T$ strategy. Both approaches avoid the explicit estimation of the full current trajectory through intermediate magnetization states and proved to achieve smooth torque transitions during demagnetization and remagnetization processes.

Finally, the dissertation addresses the inverter voltage limit during remagnetization transient, which strongly constrains the feasible operating range of VFM under load. A computationally efficient method is proposed to determine the remagnetization feasibility boundary in the torque-speed plane by combining pre-calculated flux maps and magnetization maps. The analysis shows that remagnetization introduces a hysteresis effect in the practically accessible operating region and that multi-step remagnetization can mitigate this limitation.

This dissertation provides a coherent framework for the design, modeling and control of VFMs, and contributes practical methods for enabling their adoption in high-performance and more sustainable electric drive applications.

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Chapter 1

Introduction

1.1 Research Motivations

The electrification of transportation plays a key role in reducing greenhouse gas emissions and achieving global sustainability targets. In this context, electric traction systems have become a central technology in automotive and industrial applications. Among the available solutions, Permanent Magnet (PM) Synchronous Machines (PMSMs) are widely adopted due to their high efficiency, power density and reliability. However, their performance strongly depends on the use of rare-earth permanent magnets, such as Neodymium-Iron-Boron (NdFeB), which raise critical concerns in terms of cost, supply chain security and environmental impact.

Rare-earth (RE) elements are geographically concentrated and subject to market volatility, making the large-scale implementation of Electric Vehicles (EVs) highly dependent on constrained resources. Furthermore, their extraction and processing involve significant environmental and social impact. As a result, reducing or eliminating the reliance on RE materials in electric machines has become a major challenge for sustainable electrification.

Several strategies are currently being investigated to reduce the dependence on RE materials in electric drives. These include reducing the RE content in conventional PMSMs, for example through Dy-free NdFeB-based solutions and recycling RE materials.

Dy-free NdFeB-based PMSMs are attractive because they avoid the use of heavy RE elements, but their thermal robustness may be limited, since the magnetic characteristics of Dy-free NdFeB magnets are highly sensitive to temperature variation. Recycling is also an important research direction, but the collection and separation of RE elements from existing electric machines remain costly and technically challenging. Another possible route is the use of Low Coercivity Force (LCF) PMs. However, in a conventional PMSM setting, LCF PMs would be exposed to unintended demagnetization.

Variable Flux Machines (VFMs), also known as memory motors, represent a promising solution in this direction. By enabling controllable Magnetization State (MS), *that is, intentional demagnetization and remagnetization*, VFMs allow the adjustment of the magnet flux linkage according to operating conditions, improving efficiency over a wide speed range while maintaining good performance on torque capability. In particular, the use of LCF magnetic materials enables the reduction or elimination of RE usage in electric motors, and introduces new opportunities for the design of environmentally sustainable traction drives. However, this flexibility comes at the expense of increased complexity in machine design, modeling and control.

Key challenges arise from the complex behavior of the LCF magnets, including irreversible demagnetization and remagnetization phenomena. These effects are difficult to capture with conventional modeling approaches and require advanced tools for accurate prediction and control. Moreover, the need for high-amplitude current pulses to manipulate the magnetization state introduces additional constraints at both the inverter and control levels.

In this context, the present PhD research aims to develop advanced methodologies for the design, modeling and control of VFMs, with a particular focus on enabling the use of RE-free magnetic materials in traction applications. The ultimate goal is to contribute to the development of high performance and sustainable electric drives, reducing the dependence on critical raw materials while maintaining the performance required by modern electrified transportation systems.

1.2 Permanent Magnet Synchronous Machines

PMSMs have become the most preferred traction solution for modern electromobility [1–13]. These machines are typically classified into Surface-Mounted PM (SPM) and Interior PM (IPM) configurations. Among these, the IPM topology is widely adopted in automotive applications, including Battery Electric Vehicles (BEVs), Hybrid Electric Vehicles (HEVs) and Plug-in Hybrid Electric Vehicles (PHEVs), due to its high power density, superior torque-to-mass ratio and high efficiency over a wide speed range.

The IPM motor exhibits magnetic saliency, with the q -axis inductance (L_q) greater than the d -axis inductance (L_d). Its electromagnetic torque combines PM torque and reluctance torque, enabling high torque capability even in the Field Weakening (FW) region. Examples of the commercially available IPM machine designs utilized for traction applications are shown in Fig. 1.1 [14].

In recent decades, NdFeB has become the dominant PM material in PMSMs. Its high remanence (B_r) permits a high torque density, while its high coercivity (H_c) ensures resilience to demagnetization during operation. However, NdFeB properties are highly dependent on temperature, as illustrated in Fig. 1.2. Both B_r and H_c degrade as temperature increases, and this can lead to unwanted irreversible demagnetization and a significant loss in torque performance if thermal limits are exceeded.

To improve the thermal stability of NdFeB PMs, heavy rare-earth elements such as Dysprosium (Dy) or Terbium (Tb) are often added during the NdFeB manufacturing process [15–21]. However, the extraction of these heavy RE materials involves

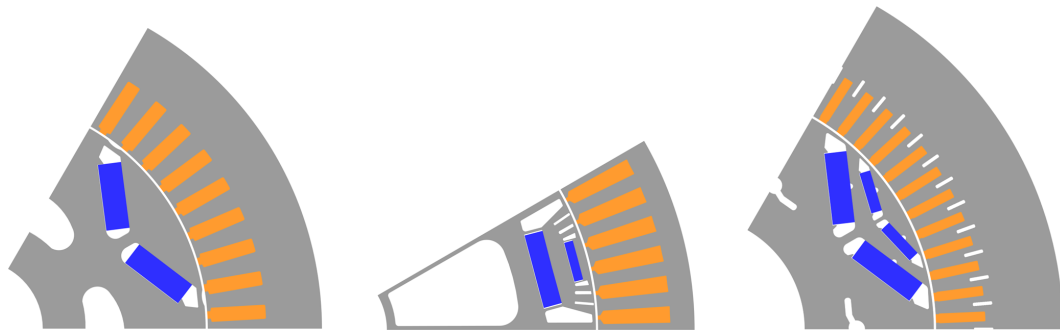


Fig. 1.1 Cross-section of PMSMs utilized for traction application: (a) Tesla Model 3 3D6, (b) BMW i3 and (c) Lucid Air.

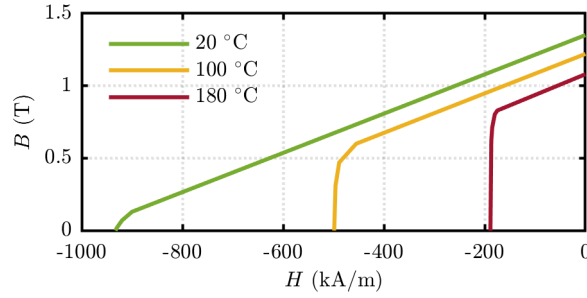


Fig. 1.2 Hysteresis curve of N45 NdFeB PM at different temperatures.

a significant environmental cost and supply chain risks [22–24]. Recent research has focused on RE recycling methodologies [25–34] and on the development of alternative RE-free magnetic materials, such as iron nitride (FeN) [35]. FeN exhibits high remanence and improved temperature stability, as shown in Fig. 1.3. However, its low coercivity makes it prone to unintentional demagnetization, limiting its applicability in conventional PMSM designs and motivating alternative machine concepts.

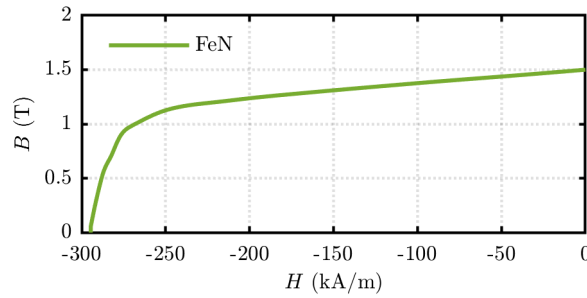


Fig. 1.3 Hysteresis curve of FeN PM.

1.3 Variable Flux Machines

Variable Flux Machines (VFMs), also referred to as memory machines [36], have attracted increasing research interest in recent years. VFMs belong to the PMSMs family but operate with LCF PM, such as AlNiCo [37–55], FeN [56–58], CeFeB [59], FeCrCo [60] and specific types of SmCo [61–65]. Among these materials, FeN is particularly attractive because it provides a RE-free PM solution, and its high remanence can satisfy the torque requirement of automotive traction applications.

The hysteresis curve of FeN PM is reported in Fig. 1.3 based on the datasheet available in [35]. Anyway, despite recent strong investments for their development, FeN magnets are currently still unavailable on the market.

The magnetic flux linkage of LCF PM is dynamically adjusted during operation by imposing dedicated current pulses: a negative d -axis current to reduce the PM flux linkage and a positive d -axis current to restore it. These processes are also called Magnetization State (MS) manipulation and enable active regulation of the effective PM flux. The MS is defined as:

$$MS = \frac{\lambda_m}{\lambda_{m,max}} \quad (1.1)$$

where λ_m is the PM flux linkage and $\lambda_{m,max}$ corresponds to the fully magnetized condition.

The MS manipulation process alters the hysteresis behavior of a LCF magnet, as illustrated in Fig. 1.4. Demagnetization shifts the operating point below the knee of the B-H curve, resulting in a reduced remanent flux after removing the dq current vector located in the second quadrant. Conversely, remagnetization restores the magnetization through a dq current vector located in the first quadrant.

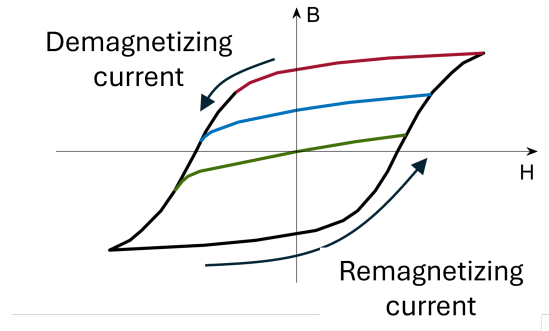


Fig. 1.4 Hysteresis curve of LCF PM and magnetic operating point at different MS levels.

This capability allows VFMs to overcome the fixed PM flux limitation of classical NdFeB-based PMSMs [53, 55, 65–69]. In conventional PMSMs, the constant PM flux leads to increased back Electromotive Force (EMF) at high speeds, requiring flux weakening (FW) current and causing additional losses. In contrast, VFMs can reduce λ_m at high speed operation, thereby lowering the back EMF and minimizing the need for FW current. At low speed and high torque, a higher MS improves torque

capability due to high PM flux linkage. As a result, VFMs can achieve a wider speed range and higher efficiency across the driving cycle, as shown in Fig. 1.5.

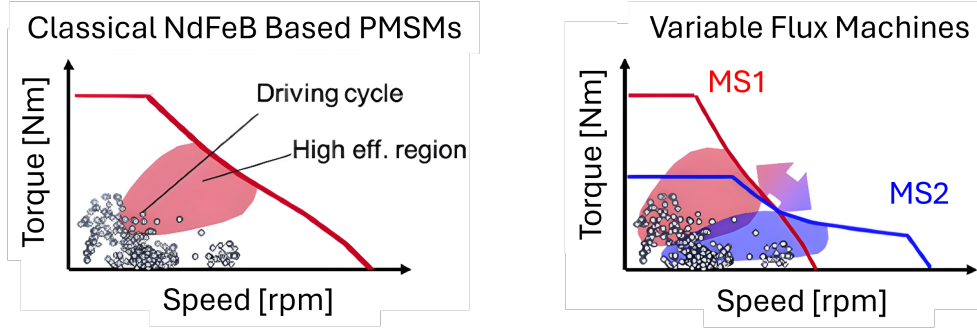


Fig. 1.5 Comparison between the torque-speed capability of a conventional NdFeB PMSM and a VFM.

The MS manipulation in VFMs is typically achieved through stator current injection, which introduces additional challenges in machine design [61, 70–72]. The demagnetizing current must be carefully limited, since unintentional demagnetization can lead to degraded torque and speed performance. On the other hand, the remagnetization process generally requires high-amplitude current injection, which may affect inverter sizing and system cost. Therefore, an appropriate balance between demagnetization and remagnetization current requirements must be carefully considered during the machine design stage.

VFMs can be classified into hybrid PM configuration [40–45, 48–55, 58, 59, 62–65] and single PM configuration [37–39, 46, 47, 60, 61]. Hybrid PM configurations are more commonly reported in the literature, combining high-coercivity NdFeB magnets with LCF magnets. In this arrangement, the NdFeB magnets provide a stable baseline flux, while the LCF magnets enable controllable flux variation through MS manipulation. However, this approach still relies on NdFeB magnets containing heavy RE elements, which raises concerns regarding cost, resource availability and environmental sustainability. In addition, the volume fraction of LCF magnets in hybrid designs is typically limited, restricting the achievable flux variation range. By contrast, the single PM configuration employs only LCF magnets, enabling full utilization of MS manipulation across the entire PM volume. This allows better flux regulation capability and eliminates the dependence on RE materials, making it a more sustainable solution. However, the absence of high coercivity magnets requires careful design to ensure sufficient baseline magnetization and stability under

operating conditions. A specific hybrid solution based on two LCF magnets has also been reported in [56, 57], where the two LCF magnets materials are FeN and AlNiCo.

Control strategies for MS regulation are critical. Simple approaches rely on injecting d -axis magnetization current pulse under no-load conditions [73–77], which inherently results in zero torque production. This method is well suited for applications such as washing machines, where operating modes are temporally separated: a high MS is maintained during the washing mode to provide high torque at low speed, while a low MS is used during the spin mode for high-speed, low-torque operation. In this type of application, the motor can be temporarily stopped during mode transitions, allowing MS manipulation without impacting performance. However, torque interruption is unacceptable in automotive traction applications, where continuous and smooth torque delivery during MS transition is required to avoid speed oscillations and potential loss of drivability. Therefore, advanced control strategies are required to enable MS variation while maintaining torque continuity and minimizing transient disturbances. Several control methods have been proposed to mitigate torque pulsations during MS variations, including decoupling-based current control relying on complex flux observers [78], modified i_q reference generation for demagnetization and remagnetization [79], hybrid control schemes combining sliding-mode speed control and MPC current control [80], and dedicated compensation schemes for dual three-phase VFMs [81]. Nevertheless, these approaches either introduce significant implementation complexity or still leave residual torque oscillations during MS transitions.

Due to the complexity of LCF PM behavior, Finite Element Analysis (FEA) is widely used in the design, modeling and control development of VFM drives, because FEA can accurately simulate the electromagnetic behavior of the machine, including nonlinear effects such as magnetic saturation and PM demagnetization and remagnetization [82]. However, this accuracy comes at the cost of high computational effort and long simulation times.

During the design stage, FEA is unavoidable for the optimization of the motor geometry. However, for control development, fast simulation models are required for testing and tuning the control systems, since a large number of simulations must be performed. Conventional electromagnetic models, such as those implemented in MATLAB/Simulink, usually do not capture the saturation and irreversible demagne-

tization phenomena of PMs. For this reason, VFM models often rely on FEA-based simulations or co-simulation approaches [83–85], accepting the associated computational burden. This represents a significant limitation for the control development of VFM drives.

1.4 Other Rare-Earth-Free Synchronous Machines

Other than VFMs, Electrically Excited Synchronous Motors (EESMs) and PM-assisted Synchronous Reluctance Machines (PMaSyRelMs) are two other typical rare-earth-free synchronous machines topologies widely adopted in traction applications.

EESMs, also referred to as Wound Rotor Synchronous Machines, are widely used by automotive OEMs such as Renault and BMW. They eliminate the need for permanent magnets by generating the excitation flux through a rotor winding [86–88]. This makes them an attractive rare-earth-free solution and provides high flexibility in flux control, especially in the field-weakening region. However, the presence of rotor excitation windings increases machine complexity and introduces additional losses and thermal management issues.

PMaSyRelMs [89–91] typically combine synchronous reluctance machine (SyRelM) with ferrite PMs. SyRelMs are attractive because of their simple rotor structure, while ferrite PMs provide an additional PM torque contribution at relatively low cost. However, the low remanence of ferrite magnets generally limits torque density and torque capability compared with high-performance rare-earth PM machines.

Similar to the excitation current injection in EESMs, VFMs also require current pulses to regulate the MS , especially during the remagnetization process. This introduces additional losses and thermal management issues. However, unlike EESMs, where the excitation current is continuously required during operation, the MS variation transients in VFMs are typically of very short duration. Therefore, the additional losses associated with MS regulation in VFMs are expected to remain limited, and the average thermal impact may be lower. However, the peak thermal stress during remagnetization can still be significant.

Compared with PMaSyRelMs, VFMs may offer higher torque density, since the LCF PMs considered for VFM applications (see Section 1.3) generally have higher

remanence than the ferrite magnets typically used in PMSyRelMs. However, this potential advantage comes at the cost of higher control complexity, since VFMs require dedicated demagnetization and remagnetization control strategies in order to regulate the MS during operation.

1.5 Contributions and Publications

The main contributions to the design, modeling and control of VFMs arising from this PhD research are summarized as follows:

- **Definition of magnetization maps for VFMs and their identification methods from FEA or experimental tests.**

Magnetization maps, including both demagnetization and remagnetization characteristics, are introduced to describe the evolution of the MS . The demagnetization map determines the MS when the motor is subjected to a demagnetization dq current vector located in the second quadrant, starting from a fully magnetized condition. The remagnetization map determines the MS when a remagnetization dq current vector is applied in the first quadrant, starting from a fully demagnetized condition.

- **Definition of flux maps for different MS levels and their identification methods from FEA or experimental tests.**

Flux maps corresponding to different MS levels are defined and extracted from FEA or experimental measurements. These maps capture nonlinear magnetic effects, including PM saturation and cross-saturation, and provide the basis for accurate control-oriented modeling.

- **Identification of the SPM topology as the most suitable rotor configuration for VFMs equipped only with LCF PMs.**

The SPM topology requires lower remagnetization current compared to IPM configurations, thus reducing inverter oversize. Furthermore, due to its magnetic isotropy, the SPM machine ensures torque continuity during MS transitions. In contrast, the thesis demonstrates that smooth torque control during demagnetization transients is impossible with IPM motors presenting a relevant magnetic anisotropy.

- **Development of a computationally efficient electromagnetic model for VFMs.**

A Look-up-table (LUT) based model is developed to accurately capture irreversible and local demagnetization phenomena occurring during MS variation under load conditions. The model can be executed in standard software for control simulation (such as Simulink or PLECS), eliminating the need for FEA-based co-simulation. The proposed model achieves high accuracy while significantly reducing computational burden, and it is validated against FEA simulations.

- **Implementation of the proposed VFM model in a real-time Hardware-in-the-Loop (HiL) environment.**

thanks to its low computational burden, the proposed model could be implemented on a PLEXIM RT Box 2 for real-time simulation, with the control algorithm embedded on an STM32 NUCLEO G474RE microcontroller. The HiL results confirm the model accuracy through comparison with FEA simulations, while demonstrating its suitability for real-time applications and permitting the emulation of complex and long-duration driving cycles.

- **Development of closed-loop control strategies for MS regulation under load.**

Two control approaches are proposed to regulate the *MS* while mitigating the torque disturbances induced by high-amplitude magnetization current pulses. The first is based on a modified Direct Flux Vector Control scheme, while the second is based on a modified Field Oriented Control approach.

- **Definition of feasible operating regions under inverter voltage constraints.**

A methodology is developed to analytically determine the feasible torque-speed operating area, including remagnetization transients where high-amplitude current pulse can saturate the DC-link voltage. The approach entirely relies on the manipulation of precomputed flux maps and magnetization maps, enabling fast evaluation.

The research outcomes produced during the PhD activities are listed below:

1. C. Chen, R. Garino, P. Pescetto, S. Ferrari and G. Pellegrino, "A Seamless Magnetization State Transition Strategy for Variable Flux Machines Based

- on Direct Flux Vector Control,” *2026 IEEE Energy Conversion Conference Congress and Exposition (ECCE)*, Vancouver, BC, Canada, 2026.
2. C. Chen, R. Garino, P. Pescetto, S. Ferrari and G. Pellegrino, “Comparative Analysis of Rotor Topologies in Variable Flux Machines with Low Coercivity Magnets,” *27th International Conference on Electrical Machines (ICEM)*, Funchal, Portugal, 2026.
 3. C. Chen, R. Garino, N. Pinzan, P. Pescetto and G. Pellegrino, “Real-time HiL Model of Variable Flux Synchronous Motor Drives,” *16th International Conference of the IMACS TC ELECTRIMACS (ELECTRIMACS)*, Palermo, Italy, 2026.
 4. C. Chen, P. Pescetto, S. Ferrari, G. F. Olson, M. Diana, T. Thiringer and G. Pellegrino, “Fast Determination of Feasible Torque-Speed Range for Variable Flux Machines Including Remagnetization Voltage Limit,” *2025 IEEE Energy Conversion Conference Congress and Exposition (ECCE)*, Philadelphia, PA, USA, 2025, pp. 1-8, doi: 10.1109/ECCE58356.2025.11260140.
 5. R. Garino, P. Pescetto, C. Chen, S. Ferrari, M. Diana and G. Pellegrino, “Accurate Simulation Model of Variable Flux Machines under Asymmetrical Demagnetization without FEA Co-simulation,” *2025 IEEE Energy Conversion Conference Congress and Exposition (ECCE)*, Philadelphia, PA, USA, 2025, pp. 1-8, doi: 10.1109/ECCE58356.2025.11260258.
 6. M. S. Mirazimi, C. Chen, P. Pescetto, S. Ferrari, G. Pellegrino, M. Diana and T. Thiringer, “Accurate Modeling of Variable- Flux PMSMs without Electromagnetic Co-Simulation,” *2024 International Symposium on Power Electronics, Electrical Drives, Automation and Motion (SPEEDAM)*, Napoli, Italy, 2024, pp. 625-630, doi: 10.1109/SPEEDAM61530.2024.10609051.

In addition, one international patent has been filed:

1. M. Diana, P. Pescetto, G. Pellegrino and C. Chen, “Controlling the magnetization state of a variable flux machine,” Patent EP4 601 185A2, 2025.

The following publications have been submitted:

1. C. Chen, R. Garino, P. Pescetto, G. F. Olson, M. Diana and G. Pellegrino, “Advanced HiL Simulation of Variable Flux Synchronous Machine for Traction Applications,” *IEEE Access*.

Additionally, the following publication was produced during the MSc thesis research activity:

1. U. Montanaro, C. Chen, A. Rizzo and A. Sorniotti, “Output-based enhanced closed-loop model reference adaptive control and its application to direct yaw moment control,” *international journal of robust and nonlinear control*, vol. 34, no 14, pp. 9471-9500, Sept. 2024, doi: 10.1002/rnc.747.

Chapter 2

Variable Flux Machine Geometry Selection

The selection of the VFM geometry is a critical design step, particularly when exclusively employing LCF PMs. Unlike conventional PMSMs, VFMs rely on intentional *MS* manipulation through inverter-driven demagnetization and remagnetization current pulses. However, the stator current amplitude required for remagnetization is typically high, directly constraining the inverter rating and overall system feasibility. Moreover, *MS* regulation may lead to torque interruption, which is unacceptable in many practical applications, especially in traction systems where smooth torque delivery is required during *MS* transitions.

The objective of this study is to compare the rotor topologies based on the following criteria:

- the demagnetization and remagnetization current requirements;
- the torque-speed characteristics across different *MS* levels;
- the capability to perform remagnetization under load conditions.

Based on this comparative analysis, a suitable rotor topology is selected and used for the design and manufacturing of a Proof-of-Concept (POC) prototype for experimental validation. To ensure a realistic and industrially relevant assessment, a commercially available stator is adopted, and only commercially available LCF magnets are considered.

2.1 Motor Parametrization and Rotor Topologies

2.1.1 Magnet Selection

The candidate LCF magnets in this study are FeN and AlNiCo. The main magnetic properties of AlNiCo 5, AlNiCo 6, AlNiCo 9 and FeN are compared in Fig. 2.1 and Tab. 2.1, with FeN data obtained from [35]. All these materials exhibit good thermal stability, but they differ significantly in terms of remanence and coercivity, which strongly influence their suitability for VFM applications.

Table 2.1 PM materials properties at room temperature

Material	B_r (T)	H_c (kA/m)	B_{knee} (T)	Density (kg/m ³)
FeN [35]	1.5	298	1.15	5100
AlNiCo 5	1.25	51	1.00	7310
AlNiCo 6	1.08	60	0.74	7340
AlNiCo 9	1.12	109	0.89	7250

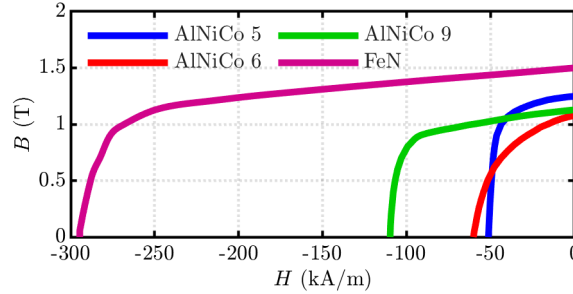


Fig. 2.1 B-H characteristics of AlNiCo and FeN

Among the considered options, FeN magnets are particularly attractive due to their high remanence and their rare-earth-free nature, which makes them the most sustainable solution. However, FeN technology is not yet sufficiently mature for industrial use and is currently not available on the market. For this reason, **AlNiCo 5 magnets are selected for the POC prototype**, as they are commercially available despite their reliance on critical raw materials.

2.1.2 Stator Parameters

According to [92], a 6-pole configuration provides a suitable compromise between remagnetization current requirements and overall efficiency. Therefore, this configuration is selected. A commercially available 36-slot stator is adopted and configured for a 6-pole, three-phase winding layout. The main geometrical and electrical parameters of the stator are listed in Tab. 2.2. The nominal current of the selected stator is 25 A.

Table 2.2 Stator Parameters

Outer diameter	200 mm	Turns per phase	120
Inner diameter	125 mm	Airgap	0.7 mm
Stack length	200 mm	Rated current	25 A _{pk}
Slots	36	Max current	100 A _{pk}
Pole pairs	3	Rated voltage	400 V _{rms}

It should be noted that the stator is not individually optimized for each rotor topology. Although a dedicated stator optimization for each rotor topology could improve the performance of the corresponding machine, such as torque density, power factor and current loading, such an approach would make the comparison less transparent, since the observed differences would arise from the combined optimization of both stator and rotor, rather than from the rotor topology itself. In particular, some IPM-based solutions could benefit from a stator and winding design better matched to their saliency characteristics. Therefore, the use of the same stator topology without individual optimization was an intentional choice, made to perform a controlled comparison among different rotor configurations under the same stator platform. By keeping the stator fixed, the influence of additional design variables, such as slot geometry, winding arrangement and thermal loading capability, is minimized. In this way, the comparison can focus more directly on the rotor-dependent effects that are most relevant for VFMs equipped with LCF PMs, namely remagnetization current requirement, robustness against unintended demagnetization, torque-speed capability at different *MS* levels, and torque controllability during *MS* transitions.

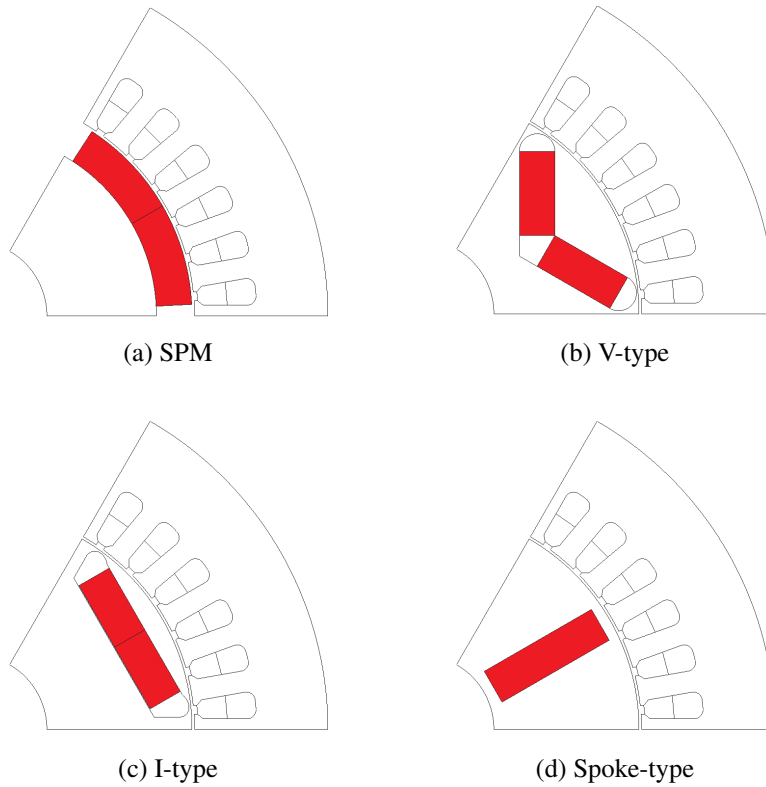


Fig. 2.2 Investigated VFMs with different rotor configurations: (a) SPM, (b) V-type, (c) I-type, and (d) Spoke-type.

2.1.3 Selected Rotor Topologies

Four rotor topologies are investigated in this study: an SPM configuration and three IPM configurations, namely V-type, I-type and spoke-type PM arrangements. The cross sections of these configurations are illustrated in Fig. 2.2. A uniform airgap thickness is maintained across all designs, consistent with mechanical constraints at maximum operating speed. This ensures that any observed performance differences can be mainly attributed to rotor geometry, rather than variations in airgap dimensions.

To ensure a consistent and fair comparison, the PMs are dimensioned according to two criteria. First, a uniform magnet thickness of 10 mm is used for all rotor configurations, providing comparable remagnetization capability and demagnetization behavior. Second, the magnet length is maximized within the geometric constraints of each rotor layout, with the objective of achieving the highest possible torque capability.

By maximizing the magnet length, each configuration operates under its most favorable magnetic loading condition, avoiding artificial penalization due to underutilization of the available rotor volume. This design choice ensures that the comparison highlights the structural characteristics of each topology, rather than differences due to suboptimal magnet sizing.

2.2 Performance Comparison of Different Rotor Configurations

The performance of the different rotor configurations is evaluated based on the following aspects:

- PM flux controllability, evaluated through the required demagnetization and remagnetization current magnitudes;
- robustness against unintended demagnetization under high load and high MS conditions;
- achievable torque-speed characteristics at different MS levels, including Maximum Torque per Ampere (MTPA) and Field Weakening (FW) operation;
- torque controllability during MS transitions.

Performing demagnetization and remagnetization processes under load can induce abrupt torque disturbances due to the instantaneous current excitation. Therefore, the last criterion specifically evaluates the capability of each topology to support effective torque mitigation strategies, which are essential for the practical application of VFM drives.

In addition, this section introduces the methodologies used to derive magnetization maps and flux maps at different MS levels, which form the basis for the subsequent analysis.

2.2.1 Synthetic Demagnetization and Remagnetization Characteristics

As a first step, a simplified analysis is carried out to evaluate the peak d -axis current required to reach a predefined MS during both demagnetization and remagnetization processes under no-load conditions. The results are presented in Fig. 2.3, where the q -axis current is set to zero.

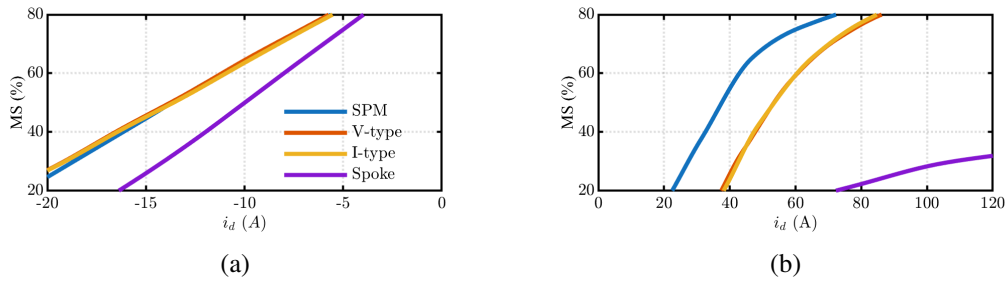


Fig. 2.3 FEA calculated magnetization characteristics of different rotor configurations with $i_q = 0$: (a) demagnetization and (b) remagnetization.

Starting from a fully magnetized condition, the demagnetization characteristic describes the negative i_d current required to reduce the MS to a desired level. Conversely, the remagnetization characteristic is obtained by applying positive i_d current, starting from a fully demagnetized state, and indicates the current required to restore the magnetization to a given MS . These curves provide an initial insight into the magnetization controllability of each rotor topology, as well as the associated current demand.

It can be observed that the SPM, V-type and I-type configurations exhibit similar behavior in terms of demagnetization, requiring comparable i_d current levels to reduce the MS . In contrast, a lower current is sufficient to significantly decrease the MS for the spoke-type configuration.

In terms of the remagnetization process, the SPM topology requires the lowest i_d current to restore the MS , making it the most favorable solution from the inverter sizing perspective. On the other hand, the spoke-type configuration is clearly penalized, as it demands significantly higher current for the remagnetization process. This characteristic makes the spoke-type configuration less suitable for practical applications, where current limitations and inverter constraints play a critical role.

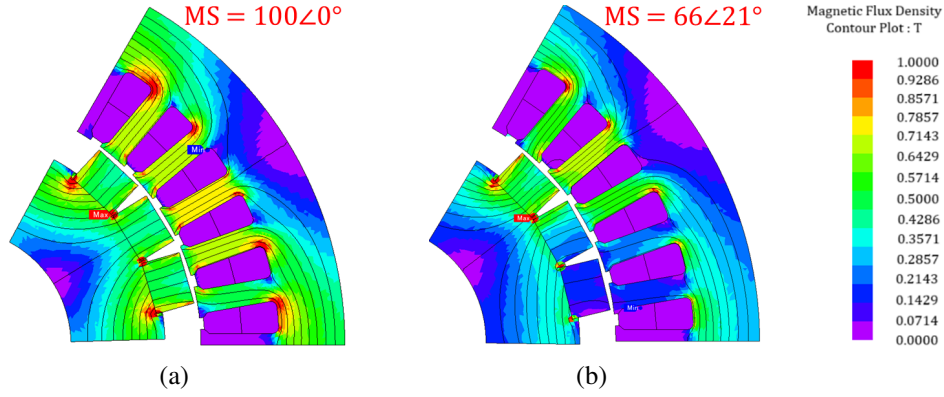


Fig. 2.4 Open circuit FEA results: (a) symmetric magnetization state and (b) deviated $\mathbf{MS}$ after asymmetric demagnetization.

2.2.2 Accurate Demagnetization and Remagnetization Maps and the Identification Method

Although the $\mathbf{MS}$ is mainly affected by the d -axis current component, the q -axis current also plays a significant role in the demagnetization and remagnetization processes. The presence of a non-zero i_q alters the magnetization direction, resulting in a distortion of the PM flux linkage with respect to the d -axis. This phenomenon leads to an angular deviation of the magnetization state, resulting in asymmetrical magnetization of the PM. An example of this asymmetry is illustrated in Fig. 2.4.

To account for both amplitude and phase variations of the PM magnetization, the complex Magnetization State ($\mathbf{MS}$) is defined as:

$$\mathbf{MS} = \frac{\lambda_m}{\lambda_{m,max}} \cdot e^{j\phi_{MS}} \quad (2.1)$$

where λ_m represents the PM flux linkage, $\lambda_{m,max}$ corresponds to its value under fully magnetized conditions, and ϕ_{MS} denotes the phase of its fundamental component with respect to the rotor d -axis. In the case of symmetric magnetization ($i_q = 0$), $\phi_{MS} = 0$. However, when uneven local demagnetization occurs, the fundamental component deviates, leading to $\phi_{MS} \neq 0$. Accordingly, the $\mathbf{MS}$ can be expressed as a nonlinear function of the stator current vector:

$$\mathbf{MS} = f(i_d, i_q) \quad (2.2)$$

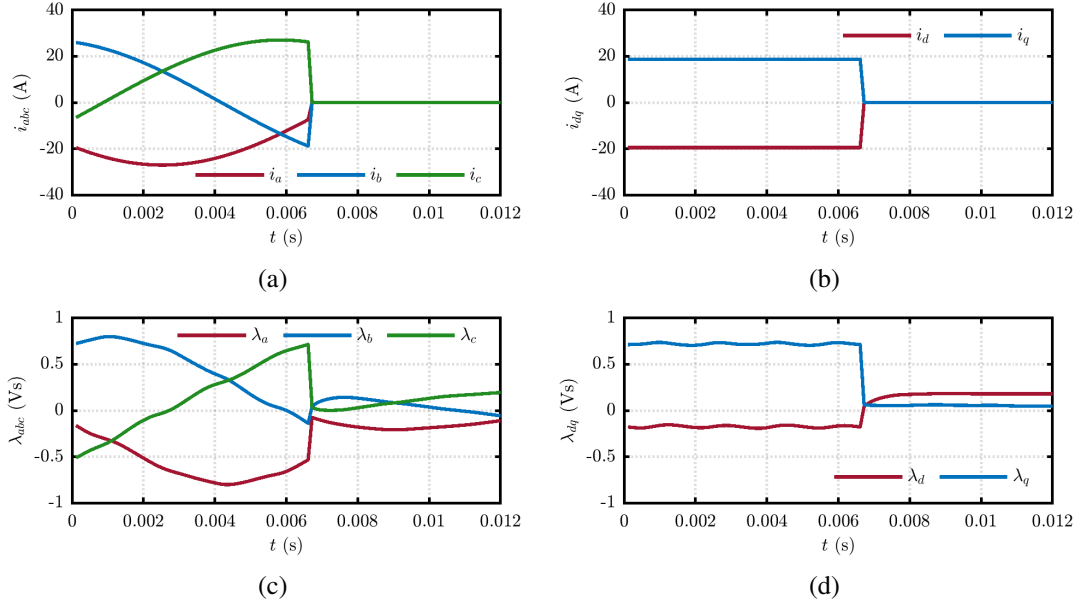


Fig. 2.5 Current and flux waveforms for a sample point **MS** measurement for V-type machine demagnetization map generation: (a) phase current, (b) dq current, (c) phase flux linkages and (d) dq flux linkage.

The extraction of the magnetization maps can be performed through FEA or experimental measurement. As an FEA example, the procedure for generating a single point of the demagnetization map is illustrated in Fig. 2.5, performed in JMAG using the V-type machine. The machine is operated at a constant speed and initialized with the PM fully symmetrically magnetized ($MS = 100\%$, $\phi_{MS} = 0$). A second-quadrant current pulse with $i_d = -19.5$ A and $i_q = 18.75$ A is applied to induce demagnetization, and is then returned to zero after the pulse. Under post-demagnetization condition, the demagnetized machine returns dq flux linkage of $\lambda_{d,PostMag} = 0.1810$ Vs and $\lambda_{q,PostMag} = 0.0474$ Vs. The **MS** is then calculated as:

$$MS = \frac{\lambda_{PostMag}}{\lambda_{m,max}} \cdot 100 \quad (2.3a)$$

$$\phi_{MS} = \tan^{-1} \left(\frac{\lambda_{q,PostMag}}{\lambda_{d,PostMag}} \right) \quad (2.3b)$$

where the $\lambda_{m,max}$ in this example is 0.7388 Vs, measured through λ_d without current injection. The flux amplitude after the **MS** transition $\lambda_{PostMag}$ is computed as:

$$\lambda_{PostMag} = \sqrt{\lambda_{d,PostMag}^2 + \lambda_{q,PostMag}^2} \quad (2.4)$$

The calculated complex magnetization state is $MS = 25\%$ and $\phi_{MS} = 14.7^\circ$.

By repeating this procedure for different current vectors, the **MS** demagnetization map is systematically explored, as illustrated in Fig. 2.9.a, c, e and g.

It is important to ensure that the duration of the magnetization current pulses is sufficient to achieve a steady magnetization condition. This can be verified by monitoring the dq flux linkages during the transients. The pulse duration must be long enough for the flux linkages to reach steady-state values before the magnetization current is removed, as shown in Fig. 2.5.d.

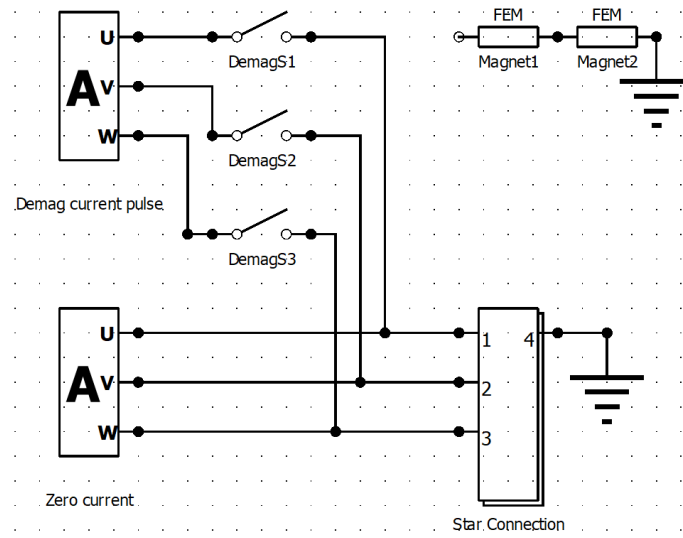


Fig. 2.6 Circuit setting for demagnetization map identification in JMAG.

An example of the circuit implementation used in JMAG for the demagnetization map identification is shown in Fig. 2.6. **JMAG is adopted as the FEA software throughout this dissertation.** Similar implementations can be realized in other electromagnetic simulation platforms, such as ANSYS Motor-CAD, since the underlying principle remains unchanged.

In this setup, the VFM circuitual model, represented by the Star Connection block, is connected to two three-phase current sources. Under normal no-load conditions, the machine is supplied by a zero-current source. When a demagnetization event is required, a second current source injects a three-phase current pulse corresponding to the desired dq current vector. The transition between the two operating conditions is managed through controlled switches (DemagS1, DemagS2 and DemagS3). These switches are closed while injecting the demagnetization current pulse and opened

once the pulse injection is completed, restoring the zero-current condition. Since the imposed current pulse is defined in the dq reference frame, it must be converted into three-phase quantities before being applied to the machine, by inverse Park and Clarke transformations, ensuring the injected currents correctly reproduce the desired dq excitation in the FEA model.

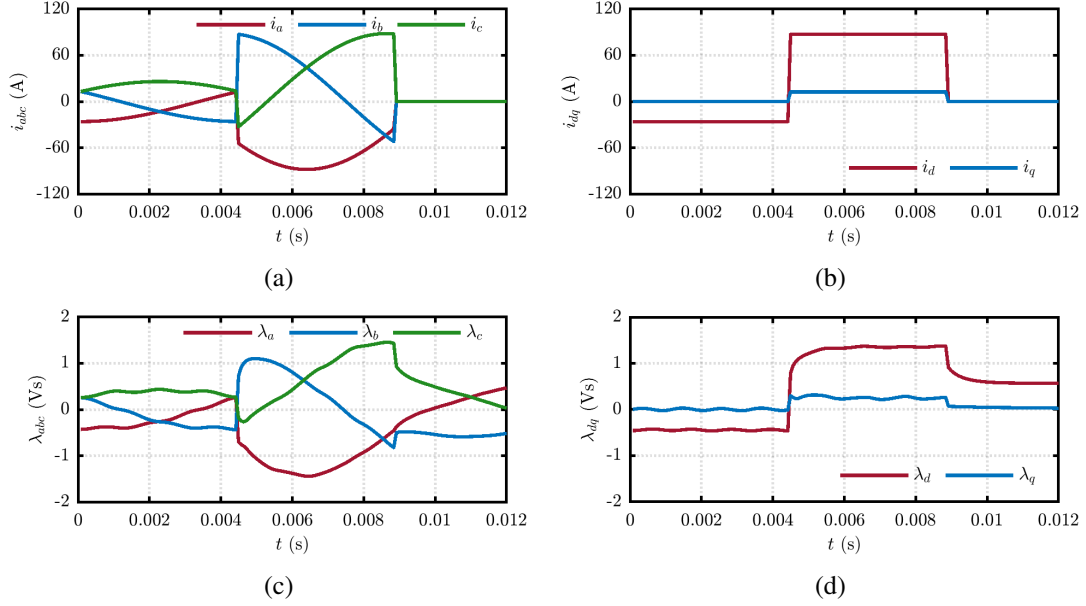


Fig. 2.7 Current and flux waveforms for a single point **MS** measurement for V-type machine remagnetization map generation: (a) phase current, (b) dq current, (c) phase flux linkages and (d) dq flux linkage.

The identification of the remagnetization map follows a similar procedure to that adopted for the demagnetization case, with the key difference that the machine must be initially demagnetized to a fully symmetrically demagnetized condition ($MS = 0$, $\phi_{MS} = 0$). An example of single point **MS** generation for the V-type machine is represented in Fig. 2.7. An initial demagnetization current is implemented to fully demagnetize the machine symmetrically, obtained from the demagnetization map, with $i_d = -26$ A and $i_q = 0$ in this case. A remagnetization current vector with $i_d = 87.3$ A and $i_q = 12.5$ A is then injected. Once the remagnetization current is removed, the resulting flux linkages are $\lambda_{d,PostMag} = 0.5676$ Vs and $\lambda_{q,PostMag} = 0.0279$ Vs. The $\lambda_{m,max}$ is 0.7388 Vs, and the resulting **MS** = $77\% \angle 2.8^\circ$ is obtained by using eq. (2.3b). By repeating this process for different current vectors, the remagnetization maps of the investigated motor topologies are obtained, as illustrated in Fig. 2.9.b, d, f and h. The circuital implementation used for remagnetization in JMag is illustrated

in Fig. 2.8. The machine is initially demagnetized by closing the switches DemagS1, DemagS2 and Demag S3. These switches are opened once the remag current pulse is imposed, closing the switches RemagS1, RemagS2 and RemagS3. Finally, all the switches are opened for the post-remagnetization phase.

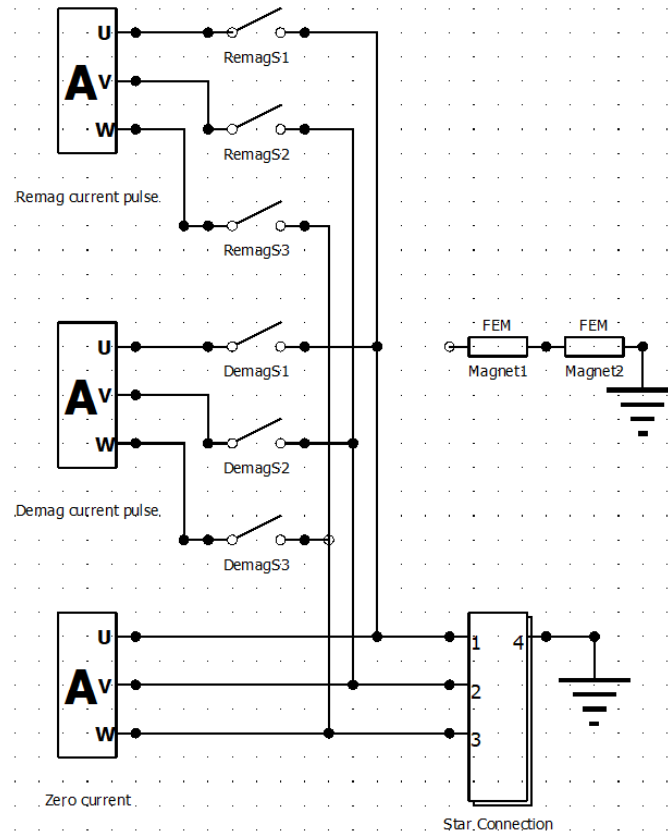


Fig. 2.8 Circuit setting for remagnetization map identification in JMAG.

It should be noted that the inverter voltage limitation can be neglected in FEA simulations, allowing the application of high-amplitude remagnetization current without constraints. However, in experimental implementations, it is recommended to operate at low rotor speeds, since the combination of high speed and large remagnetization current can cause the back-EMF to exceed the available inverter voltage.

The magnetization map results shown in Fig. 2.9 clearly indicate that the spoke-type rotor exhibits the poorest controllability of the magnetization state. It shows a very high sensitivity to demagnetization, as the MS drops below 80 % already with a relatively small negative i_d current, approximately 5 A. This high susceptibility

to demagnetization is coupled with an extremely large remagnetization current requirement. Restoring the PM to high MS levels (e.g., above 90 %) requires a current amplitude exceeding 700 A, which is approximately seven times higher than that required by the other topologies. **Such a high current demand severely penalizes the inverter rating and makes the spoke-type configuration impractical for real applications.** Its only advantage is the relatively limited ϕ_{MS} variation, indicating a more uniform magnetization distribution compared to the other designs under load.

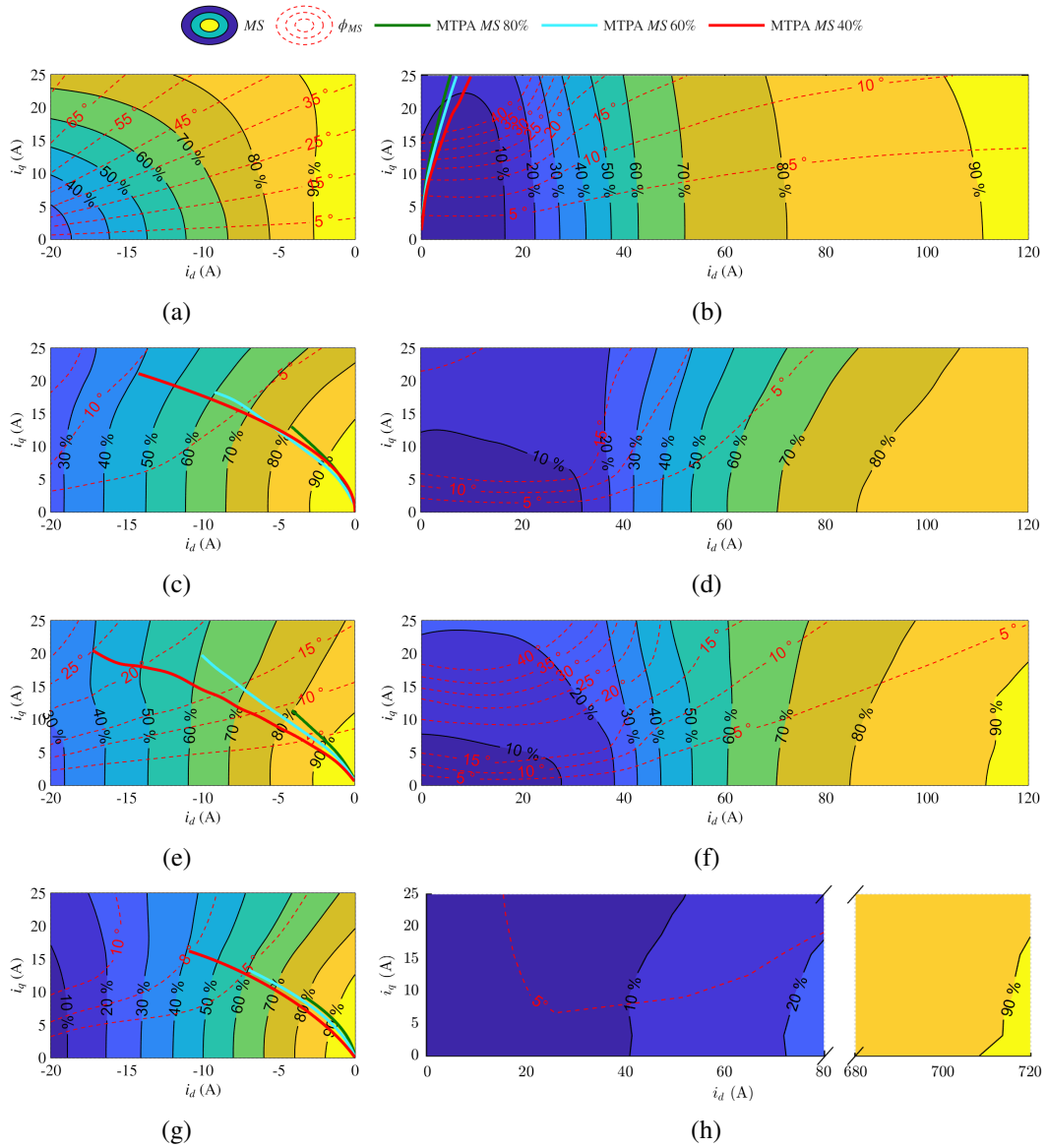


Fig. 2.9 MS amplitude and phase variation considering load current: (a) SPM demag, (b) SPM remag, (c) V-type demag, (d) V-type remag, (e) I-type demag, (f) I-type remag, (g) Spoke-type demag and (h) Spoke-type remag maps.

The V-type and I-type IPM configurations exhibit similar behavior in terms of demagnetization and remagnetization current requirements, representing a reasonable compromise from the perspective of current demand. However, differences arise when considering the phase distortion. The I-type topology shows a higher sensitivity to ϕ_{MS} variation, indicating a greater tendency toward asymmetrical magnetization

under load conditions. This may negatively impact torque production and control stability.

The SPM topology stands out as the most favorable solution in terms of remagnetization current requirement. It achieves the desired magnetization levels with the lowest current amplitude among all considered configurations, which is a key advantage for reducing inverter oversizing. Additionally, the SPM rotor requires slightly higher demagnetization current compared to the other topologies, which is beneficial in preventing unintended demagnetization during normal operation. However, this configuration also exhibits the highest sensitivity to phase distortion. This effect must be carefully considered in the control design.

The MTPA curves of the investigated VFM topologies are also represented in Fig. 2.9, obtained from the flux maps that will be discussed in the following section. The three IPM types exhibit MTPA trajectories in the second quadrant due to their anisotropy. At high MS , the demagnetization contour is reached at relatively low current, preventing full exploitation of the current rating. For example, the green curve in Fig. 2.9.c shows that the V-Type IPM motor can operate at $MS = 80\%$ and MTPA conditions in the current range 13.7 Apk or lower, which is well below the stator thermal current limit of 25 A. The spoke-type is even more penalized (Fig. 2.9.g), as the MTPA and the 80% demagnetization limit meet already at 9.6 Apk.

The SPM machine has its MTPA curves in the first quadrant of the dq plane due to ϕ_{MS} variation under load. This guarantees robustness against unintentional demagnetization under torque, including full load when the MS is in its high range, because torque operation always resides in the remagnetization quadrant and not in the demagnetization quadrant, unlike the IPM motors.

2.2.3 Flux Maps of Different Magnetization States and Torque-Speed Profiles

Flux maps represent the relationship between stator currents and flux linkages in the dq reference frame. These maps capture nonlinear magnetic phenomena such as saturation and cross-coupling effects, and they form the basis for torque computation, control design and performance evaluation across different operating conditions. These maps are typically obtained through FEA or experimental measurements.

In VFMs, flux maps depend on both the current vector and the magnetization state:

$$\begin{cases} \lambda_d = \lambda_d(\mathbf{i}_{dq}, MS) \\ \lambda_q = \lambda_q(\mathbf{i}_{dq}, MS) \end{cases} \quad (2.5)$$

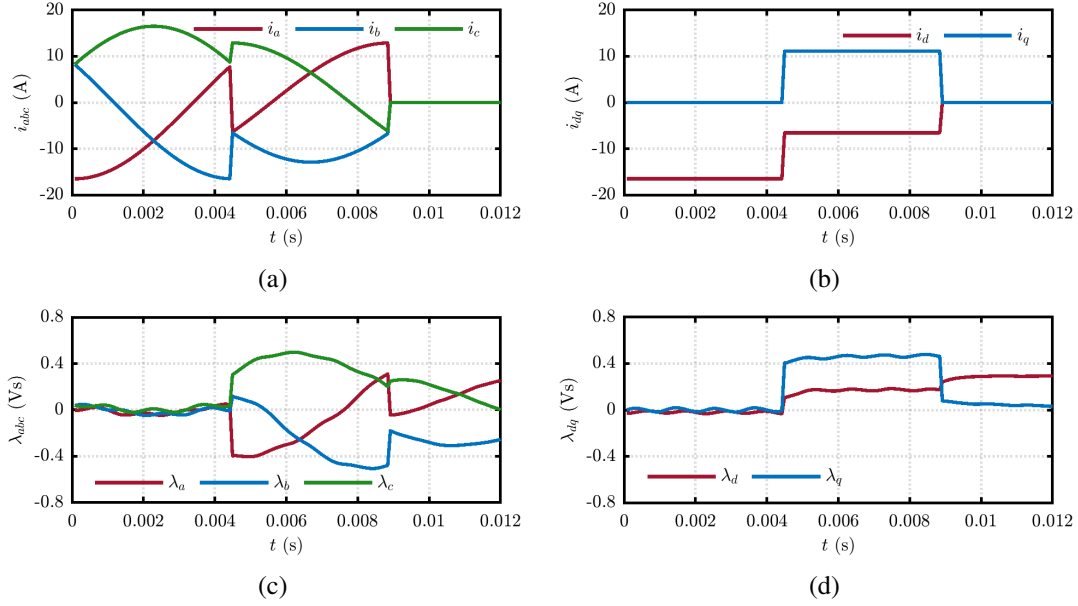


Fig. 2.10 Current and flux waveforms for a single point flux map generation for V-type machine at MS 40%: (a) phase current, (b) dq current, (c) phase flux linkages and (d) dq flux linkage.

The flux maps are obtained by initializing the machine at a specific MS at a constant speed, then applying load current vectors and recording the corresponding flux linkages. An example of single point flux map generation for the V-type motor at MS 40% is shown in Fig. 2.10. The machine is symmetrically demagnetized to $MS = 40\%$ initially by imposing a current vector obtained from the demagnetization map (Fig. 2.5.c), corresponding to $i_d = -16.5$ A and $i_q = 0$. A load current vector is then injected. In this example, the i_d and i_q load current are -6.55 A and 11.11 A. The corresponding flux linkages λ_d and λ_q of 0.1737 Vs and 0.4663 Vs are recorded. The current supply is interrupted at the end to verify if the MS remains at the target value of 40%.

The process is repeated for different current vectors and magnetization state. The circuital implementation used in JMAG is presented in Fig. 2.11, where the machine is initially demagnetized by closing the switches DemagS1, DemagS2 and DemagS3.

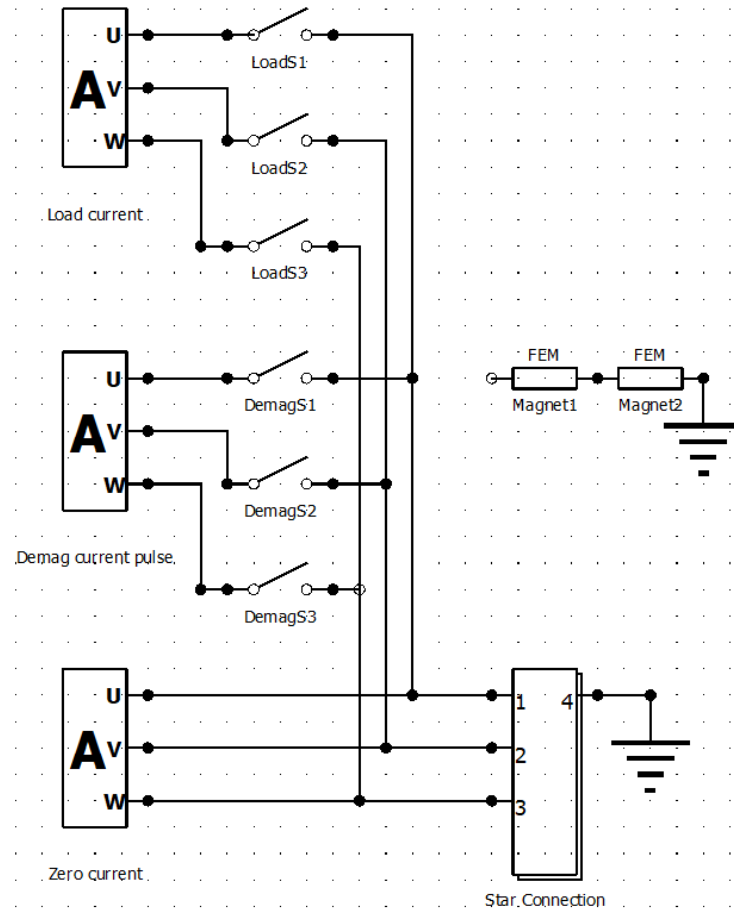


Fig. 2.11 Circuit setting for flux map generation in JMAG.

These switches are opened once the load current is imposed, closing the switches LoadS1, LoadS2 and LoadS3. The zero current is used to verify if the final *MS* remains at the target value.

The resulting flux maps for the investigated rotor topologies are represented in Fig. 2.12, with three *MS* levels of 40%, 60% and 80% selected. The current domains are different for each *MS*, according to the corresponding limit indicated in the magnetization maps (Fig. 2.9). The SPM, V-type and I-type IPM topologies have similar operating area, while the spoke-type has a reduced working limit in the second quadrant due to its weakness against the demagnetization current, and it exhibits extremely high current amplitude in the first quadrant, as the required remagnetization current is approximately 550 Apk.

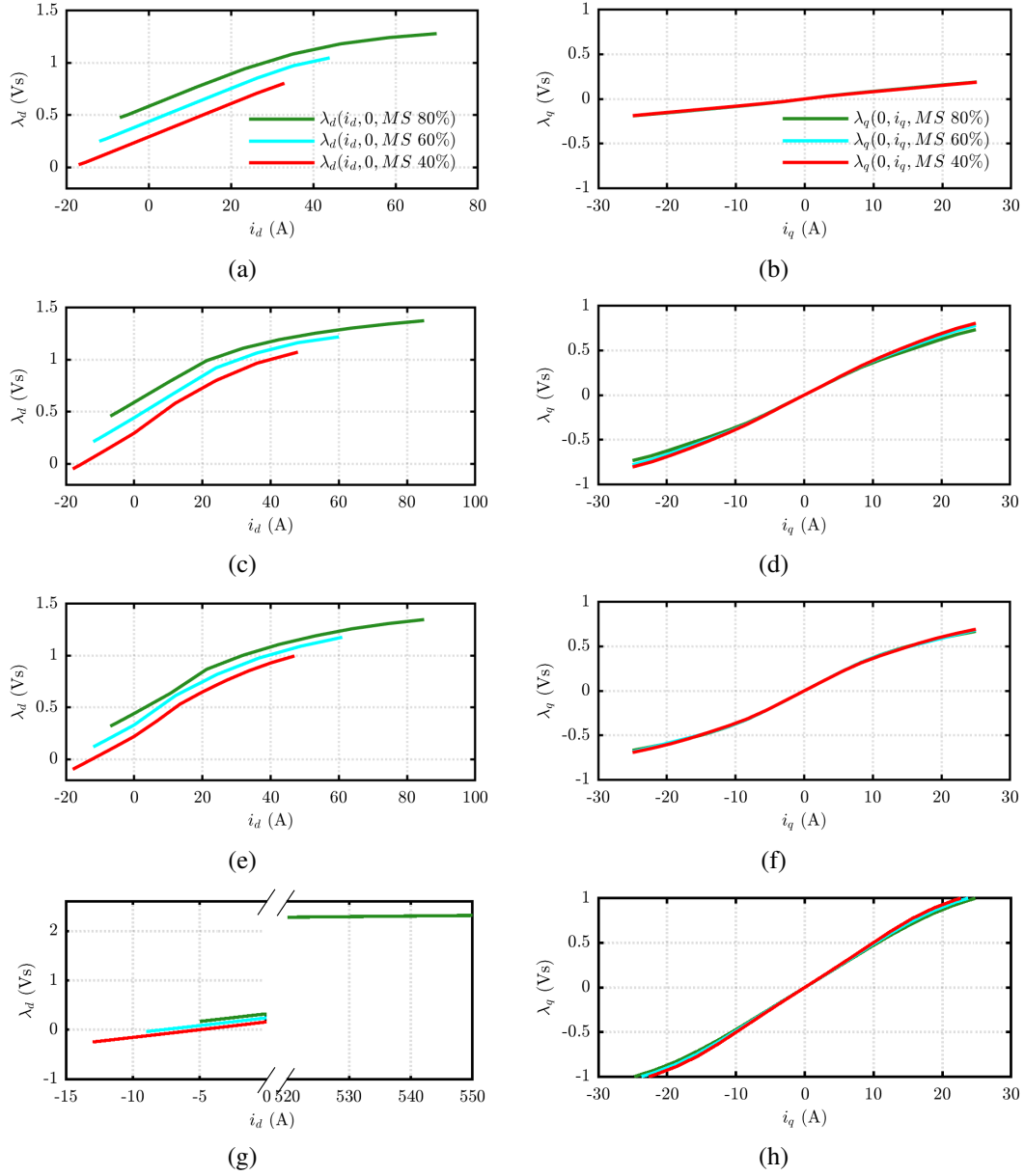


Fig. 2.12 Flux maps of different motor topologies: (a) SPM λ_d , (b) SPM λ_q , (c) V-type λ_d , (d) V-type λ_q , (e) I-type λ_d , (f) I-type λ_q , (g) Spoke-type λ_d and (h) Spoke-type λ_q .

The torque-speed characteristics can be derived from the flux maps, as represented in Fig. 2.13. High MS levels are preferred in the high-torque and low-speed region to extend the torque limit, whereas lower MS levels are utilized to extend the operating speed and minimize losses in the medium and high speed regions.

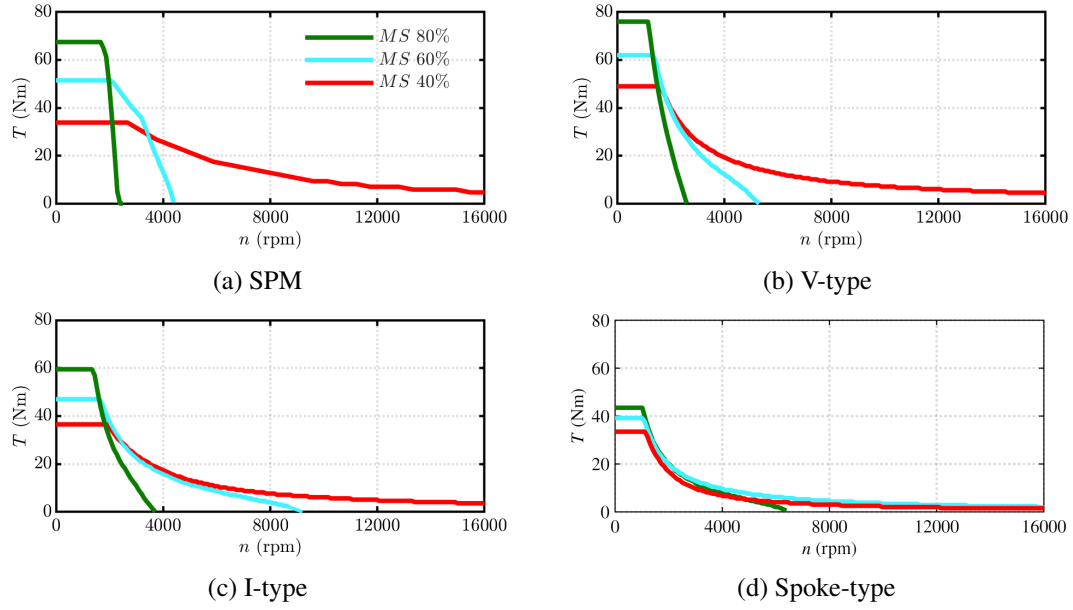


Fig. 2.13 Torque-speed profiles of the investigated VFMs topologies at MS 40%, 60% and 80%.

The V-type rotor provides the highest torque limit in constant torque region, with 76 Nm, 62 Nm and 48 Nm at MS of 80%, 60% and 40%, respectively. The SPM and I-type topologies exhibit slightly lower torque limits, respectively 68 Nm and 60 Nm at $MS = 80\%$. The spoke-type rotor has a maximum torque of 42 Nm at $MS = 80\%$, nearly half of that of the V-type. In addition, both peak torque and FW profiles at different MS levels are very close to each other, making MS regulation ineffective.

Based on this analysis, as well as on the results of Section 2.2.2, the spoke-type is retained unsuitable for future development because of its poor torque capability, limited effect of MS regulation, highest vulnerability to demagnetization under load and, most importantly, the highest remagnetization current requirement. A dedicated redesign of the stator and winding could potentially improve the torque capability and reduce the remagnetization current requirement. However, this would no longer be consistent with the comparative framework adopted, in which all rotor configurations are evaluated under the same stator settings. Therefore, within the adopted design conditions, **the spoke-type rotor configuration is not further discussed in this dissertation.**

2.2.4 Torque Controllability during MS Manipulation

The torque performance during *MS* transitions is a key design aspect of VFMs, particularly for traction applications where torque continuity is required during *MS* variations.

Under normal operating conditions, the *dq* current follows the MTPA law. However, during remagnetization or demagnetization transients, the current operating point must shift from the MTPA curve toward the first or second quadrant, respectively. The MTPA curves of the IPM topologies are located in the second quadrant, whereas for the SPM geometry the MTPA curve is located in the first quadrant, as illustrated in the Fig. 2.9. If the *MS* transition is actuated under load, the *dq* current vector must follow an iso-torque contour for having a smooth torque response.

The iso-torque contour is calculated from the electromagnetic torque equation:

$$T = \frac{3}{2}p [\lambda_d(i_d, i_q, MS) \cdot i_q - \lambda_q(i_d, i_q, MS) \cdot i_d] \quad (2.6)$$

where p is the number of pole pairs. The flux linkage λ_d and λ_q as functions of the current vector and *MS* are obtained from the flux maps, as discussed in Section 2.2.3.

The iso-torque curves for the SPM, V-type and I-type machines are illustrated in Fig. 2.14.a-c. The contour of the SPM case is the most regular one, especially at *MS*=80%, providing the feasibility of both demagnetization and remagnetization under constant torque. Conversely, the two IPM types have diagonal torque lines and even a pseudo-vertical zero-torque line in the first quadrant, where the PM and reluctance torque components compensate each other.

In the remagnetization scenario, the current operating point located along the MTPA curve is required to move toward the first quadrant. For the V-type and I-type IPM machines, remagnetizing the PM up to *MS* = 80% requires current vector with i_d component above 80 A (Fig. 2.9.d and Fig. 2.9.f). However, their zero-torque curve are located around $i_d = 20$ A, meaning that [the torque interruption is inevitable for the investigated IPM types, regardless of the adopted control strategy, because the remagnetization trajectory crosses a region in which the reluctance torque counteracts the PM torque contribution. IPM configurations with reversed saliency characteristics \(\$L_d < L_q\$ \), also referred to as flux-intensifying designs \[61\],](#)

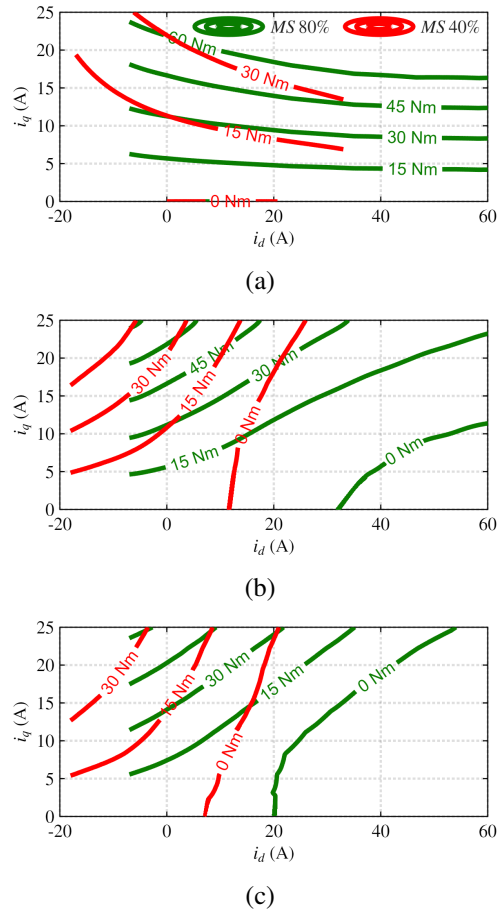


Fig. 2.14 FEA calculated iso-torque curves for different *MS* levels of (a) SPM, (b) V-type and (c) I-type rotor configurations.

could exhibit a different torque behavior during remagnetization and therefore may not suffer from the same limitation.

In the demagnetization case, both investigated IPM rotor configurations exhibit continuous iso-torque lines throughout the second quadrant, demonstrating that constant torque control is possible during demagnetization.

Returning to the SPM machine, its isotropic geometry results in negligible reluctance torque, avoiding the occurrence of zero torque conditions in the first quadrant of the current plane. In Fig. 2.14.a, it is evident that the 0 Nm trajectory coincides with the horizontal axis and does not interfere with demagnetization or remagnetization operations. Consequently, the iso-torque curves can be followed in both the first and second quadrants, thus **permitting torque control under load**

during MS variations if a proper control strategy is implemented. This provides a distinct advantage over the anisotropic IPM designs.

2.3 Proof-of-Concept VFM Motor

The SPM rotor geometry is selected for proof-of-concept (POC) experimental validation, due to its lower remagnetization current requirement, good torque-speed profile and most importantly, its capability to maintain the torque during MS variation with load.



Fig. 2.15 Proof-of-concept VFM.

However, the ideal SPM uses arc-shaped magnets, which increase manufacturing complexity. Therefore, a segmented PM shape is adopted to simplify fabrication, with the 10 mm thick AlNiCo 5 magnets, as shown in Fig. 2.15. A carbon fiber sleeve is used for structural integrity, and for this reason the magnetic air-gap is increased with respect to the case study in the previous sections. The stator parameters of the POC are illustrated in the Tab. 2.3.

Table 2.3 POC Stator Parameters

Outer diameter	200 mm	Turns per phase	120
Inner diameter	125 mm	Airgap	1.4 mm
Stack length	200 mm	Rated current	25 A _{pk}
Slots	36	Max current	100 A _{pk}
Pole pairs	3	Rated voltage	400 V _{rms}

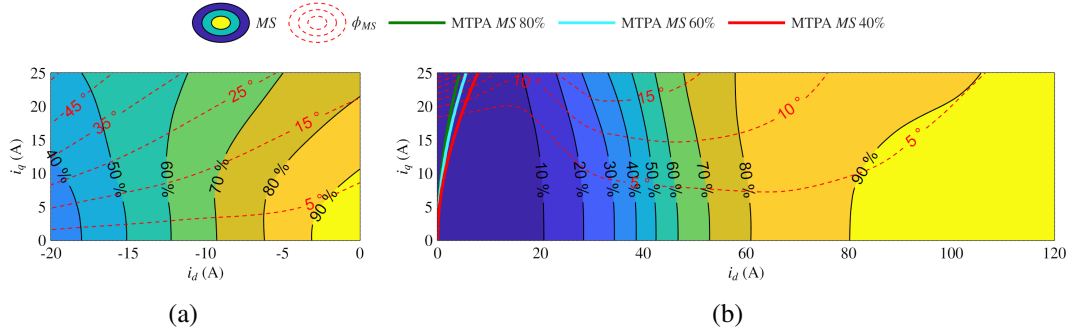


Fig. 2.16 MS amplitude and phase variation considering load current of the POC VFM. (a) demag map and (b) remag map.

Adopting the segmented PMs introduces two trade-offs compared to the ideal arc-shaped magnetic geometry. First, the PM volume is reduced by approximately 25%, resulting in lower λ_m and therefore lower torque per Ampere. Second, the segmentation results in a non-uniform airgap length and a larger equivalent airgap, further reducing the fundamental airgap flux density. These effects are visible in the FEA-calculated demagnetization and remagnetization maps shown in Fig. 2.16, by comparison with the ideal SPM case (Fig. 2.9.a and Fig. 2.9.b). Despite these quantitative differences, the maps maintain the same qualitative behavior and preserve favorable torque controllability and minimized current overrating characteristics.

The flux maps of the POC SPM are illustrated in Fig. 2.17.a and b, with the current operating limits for each MS corresponding to the demagnetization and remagnetization maps. The torque-speed characteristics and the iso-torque curves are presented in Fig. 2.17.c and d. Compared to the ideal SPM configuration, the maximum torque at each MS is reduced due to the lower PM flux linkage. Nevertheless, the iso-torque curves remain continuous across both the first and second quadrants of the dq -plane, ensuring smooth torque transitions during both remagnetization and demagnetization processes, under appropriate torque and MS control.

2.4 Traction Motor with FeN Magnets

A traction VFM prototype based on FeN PMs (Tab. 2.1) is designed for feasibility assessment, despite the current limitations in the availability of the FeN materials.

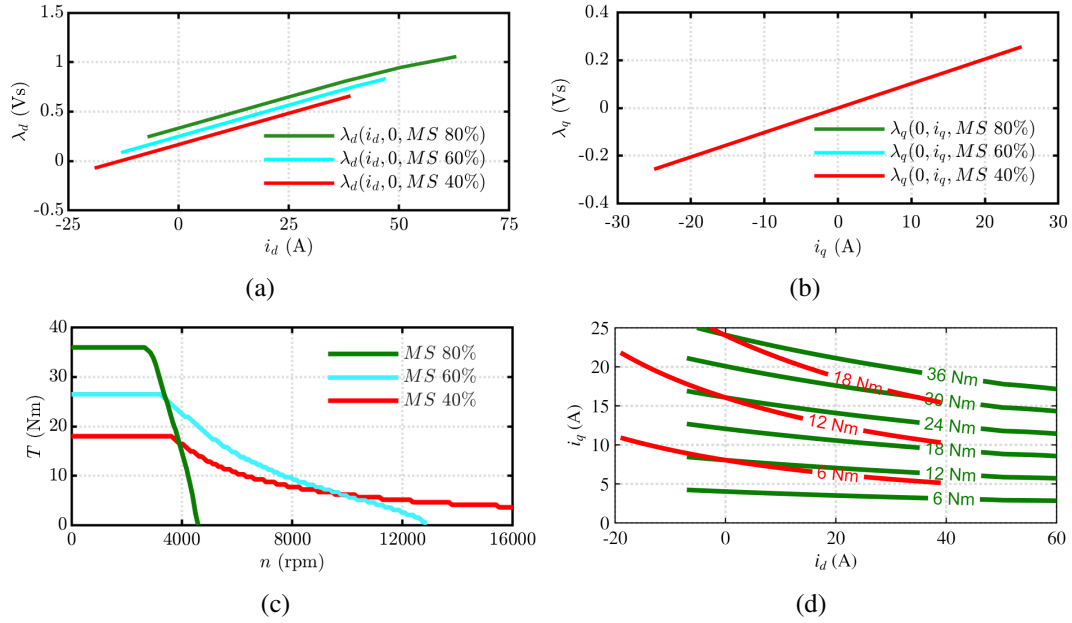


Fig. 2.17 Proof-of-concept motor profile: (a) λ_d map, (b) λ_q map, (c) torque-speed curve and (d) iso-torque curve.

The machine has four pole pairs and uses ideal arc-shaped SPM configuration with a magnet thickness of 8 mm. Its cross section is shown in Fig. 2.18. The corresponding stator parameters are presented in Tab. 2.4.

Table 2.4 Traction VFM Stator Parameters

Outer diameter	235 mm	Turns per phase	24
Inner diameter	155 mm	Airgap	0.7 mm
Stack length	155 mm	Rated current	300 A _{pk}
Slots	48	Max current	2000 A _{pk}
Pole pairs	4	Rated voltage	500 V _{rms}

The demagnetization and remagnetization characteristics are illustrated in Fig. 2.19, obtained from the FEA. Similar to the SPM machine analyzed in Section 2.2.2 and Section 2.3, the MTPA trajectories are located in the first quadrant. This ensures that unintentional demagnetization will not occur during normal operation. The machine is sensitive to phase variation, especially in the demagnetization quadrant.

The flux maps of the machine at different magnetization states are shown in Fig. 2.20. The corresponding iso-torque profiles are illustrated in Fig. 2.21. The zero torque lines are again aligned with the horizontal axis, confirming again the

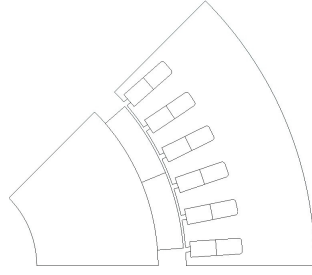


Fig. 2.18 Cross section of the traction VFM.

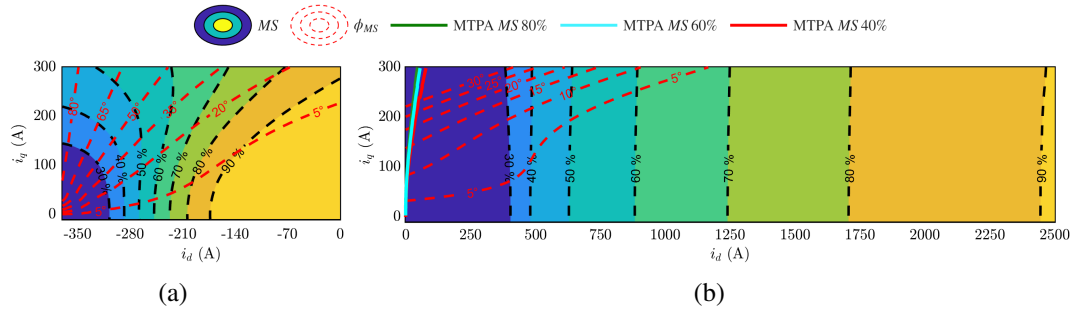
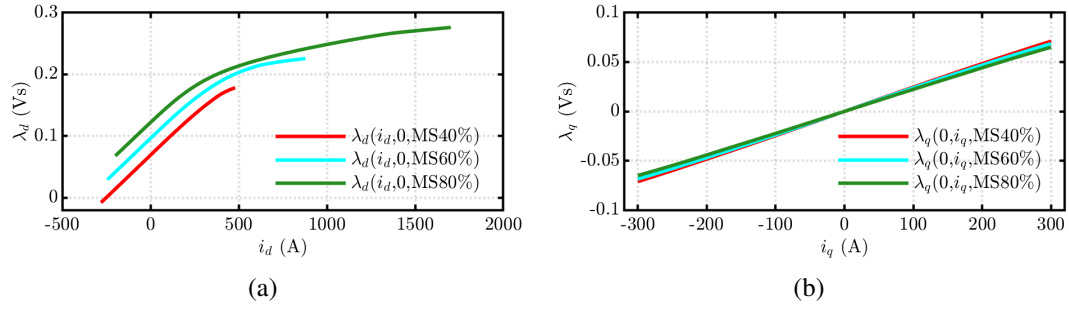
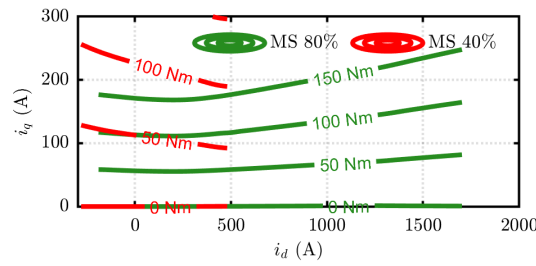
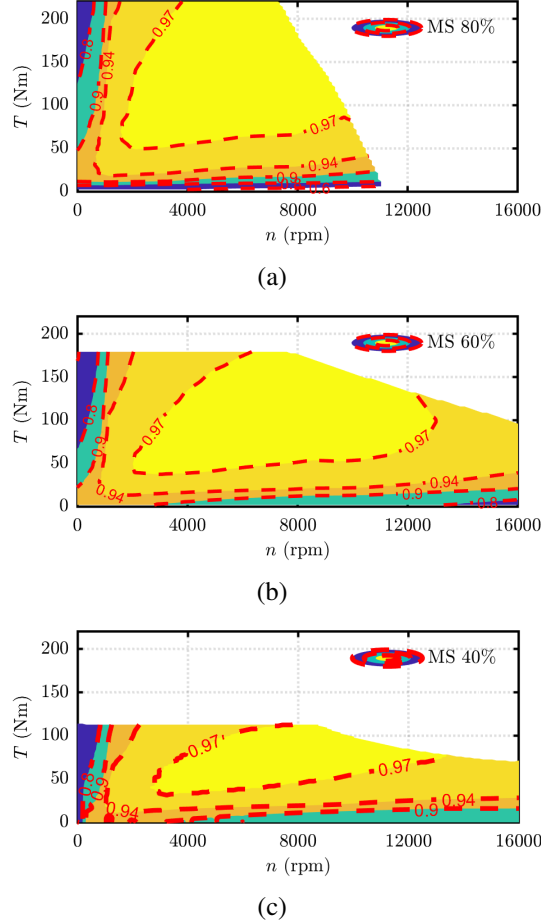


Fig. 2.19 MS amplitude and phase variation considering load current of the traction VFM: (a) demag map and (b) remag map.

SPM geometry can provide torque continuity during both demagnetization and remagnetization transients, as the current working points move from the MTPA curves to the corresponding magnetization regions without crossing the zero torque lines.

Fig. 2.20 Flux maps of the traction VFM: (a) λ_d and (b) λ_q map.Fig. 2.21 FEA calculated iso-torque curves for different MS levels of the traction motor prototype.Fig. 2.22 Efficiency maps of the traction machine with MS of (a) 80%, (b) 60% and (c) 40%.

The efficiency maps corresponding to different MS levels are presented in Fig. 2.22, derived from the flux maps. As expected, higher MS provides higher efficiency in the low speed and high torque region, while lower MS is more suitable in the FW region at higher speed.

2.5 Chapter Conclusion

This chapter demonstrated that rotor topology plays a fundamental role in determining the feasibility of VFMs based on low-coercivity magnets. Through a systematic comparison, it was shown that the SPM configuration provides the best compromise among the following aspects:

- minimization of the remagnetization current;
- robustness against unintentional demagnetization during normal operation, i.e., when the current vector follows the MTPA trajectories;
- capability to provide torque continuity during MS variations, i.e., moving the dq current vector from the MTPA curves to the remagnetization and demagnetization working points located in the first and second quadrant, respectively.

Therefore, the SPM topology is identified as the most suitable geometry for VFM applications. Moreover, this chapter introduced the methodology for generating demagnetization map, remagnetization map and flux map of different magnetization states for VFMs. **These findings establish the foundation for the modeling and control developments presented in the subsequent chapters.**

Chapter 3

Control Simulation Model

The development of accurate and computationally efficient models is essential for the control design and performance evaluation of VFMs. As discussed in chapter 2, the electromagnetic behavior of VFMs is strongly influenced by *MS* levels, which makes their modeling significantly more complex than that of conventional PMSMs.

Although FEA provides the most accurate representation of the electromagnetic behavior of LCF PMs, it requires intensive computations, making this approach unsuitable for VFM control development, where a large number of simulations must be executed within a reasonable time. This limitation motivates the development of control-oriented models that preserve the key physical characteristics of VFMs while ensuring fast execution.

In this context, this chapter presents a control-oriented modeling framework for VFMs. A baseline model is first introduced, capable of representing irreversible demagnetization and remagnetization. The model is then extended to include phase distortion of the magnetization state when the *MS* is varied under load, as introduced in Section 2.2.2. The developed models are suitable for both discrete-time simulation environments, such as MATLAB/Simulink and PLECS, and real-time Hardware-in-the-Loop (HiL) implementation. The proposed framework is based on the combination of:

- magnetization maps describing demagnetization and remagnetization dynamics (Section 2.2.2);
- flux maps for different magnetization states (Section 2.2.3);

- an update mechanism to track the evolution of the MS during the operation.

3.1 Machine Model Including Irreversible Demagnetization

The dynamic voltage equation of a VFM is the same as that of any PMSM:

$$\mathbf{v}_{dq} = R_s \mathbf{i}_{dq} + \frac{d\boldsymbol{\lambda}_{dq}}{dt} + \omega \mathbf{J} \boldsymbol{\lambda}_{dq} \quad (3.1)$$

where **bold** symbols denote the vector quantities. The vectors $\mathbf{v}_{dq}$, $\mathbf{i}_{dq}$ and $\boldsymbol{\lambda}_{dq}$ represent the voltage, current and flux linkage in the dq reference frame, respectively. R_s is the stator resistance, ω is the electrical angular speed and $\mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$ is the imaginary unit matrix.

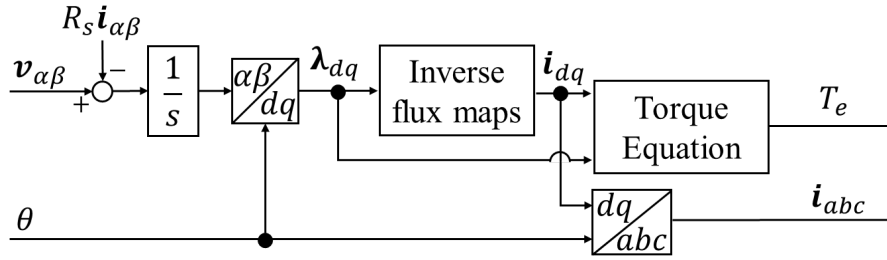


Fig. 3.1 Electromagnetic model of conventional PMSMs using CCG approach.

In conventional PMSMs, the electromagnetic model is typically built using a circuitual Controlled Current Generator (CCG) approach [93–95], as illustrated in Fig. 3.1. In this framework, the stator flux linkages are obtained by integrating the back-EMF. The relationship between flux linkages and stator currents is described through precomputed flux maps (Section 2.2.3), often stored as Look-Up-Tables (LUTs) to include the magnetic saturation and cross-coupling effects in the model. To directly determine the stator current from the flux linkages, the flux maps are manipulated to obtain the inverse flux maps:

$$\begin{cases} i_d = i_d(\lambda_d, \lambda_q) \\ i_q = i_q(\lambda_d, \lambda_q) \end{cases} \quad (3.2)$$

The electromagnetic torque is then computed from the dq current and flux linkages using the PMSM torque equation (see eq. (2.6)). Finally, the phase currents i_{abc} are obtained from i_{dq} through inverse Park and Clarke transformations.

3.1.1 VFM Electromagnetic Model

The fundamental difference between conventional PMSMs and VFMs is that the flux linkages depend not only on i_{dq} , but also on the MS :

$$\begin{cases} \lambda_d = \lambda_d(i_d, i_q, MS) \\ \lambda_q = \lambda_q(i_d, i_q, MS) \end{cases} \quad (3.3)$$

In this baseline model, the phase distortion phenomenon associated with MS variation under load is neglected. Under this assumption, the MS representation can be simplified as:

$$MS = \frac{\lambda_m}{\lambda_{m,max}} \quad (3.4)$$

where λ_m is the PM flux linkage and $\lambda_{m,max}$ corresponds to the fully magnetized condition.

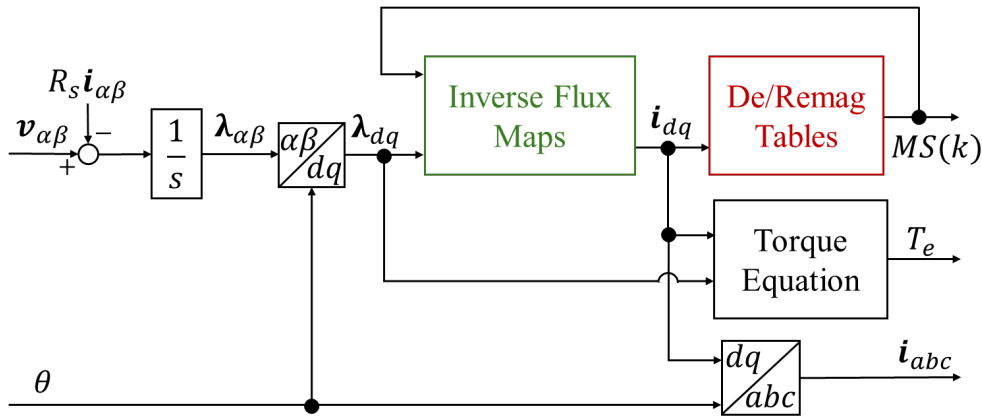


Fig. 3.2 Baseline VFM electromagnetic model.

A schematic view of the baseline VFM model is shown in Fig. 3.2. The structure is derived from the canonical PMSM electromagnetic model and extended by introducing the MS as an additional input to the inverse flux maps, with respect to the conventional scheme shown in Fig. 3.1.

Accordingly, the inverse flux maps become:

$$\begin{cases} i_d = i_d(\lambda_d, \lambda_q, MS) \\ i_q = i_q(\lambda_d, \lambda_q, MS) \end{cases} \quad (3.5)$$

In this way, i_{dq} is determined by both λ_{dq} and MS .

3.1.2 MS Updating Mechanism

The core function of the baseline VFM model is the De/Remag Tables block, which is used to update and memorize the MS . A detailed view of this block is shown in Fig. 3.3. The Demag map and Remag map blocks contain the magnetization map introduced in Section 2.2.2 and are used to determine whether the dq current operating point lies within the corresponding MS working region:

- if i_{dq} exceeds the corresponding contour in the demagnetization map, MS is reduced;
- if i_{dq} exceeds the corresponding contour in the remagnetization map, MS is increased;
- if i_{dq} remains within the corresponding contours in both maps, MS is unchanged.

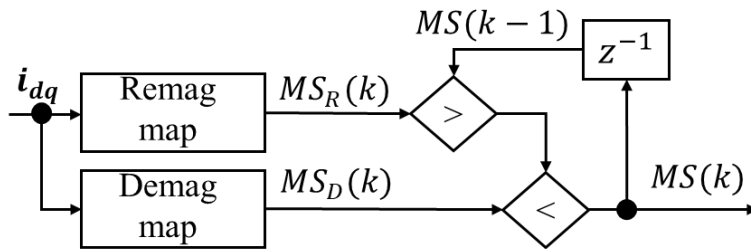


Fig. 3.3 MS updating mechanism.

The updated value of MS is then fed back into the inverse flux maps, which determine the dq current corresponding to the MS .

3.1.3 Model Validation against FEA Simulation

To validate the baseline model, the POC motor described in Section 2.3 is implemented in MATLAB/Simulink. The dynamic behavior of the model is assessed through comparison with FEA simulation performed in the JMAG Designer environment. Both demagnetization and remagnetization tests performed to evaluate the accuracy of the proposed VFM model. In the Simulink framework, the VFM operates under a closed-loop current control scheme, while the resulting current waveforms are directly imposed as inputs in the JMAG simulations.

It should be noted that the model is developed in the fundamental dq reference frame and therefore neglects space harmonics in the air-gap flux distribution. Moreover, eddy current effects are not included. As a consequence, harmonic ripple in the phase quantities is not explicitly reproduced. These simplifications are acceptable in the present work because the model is intended for control oriented simulation, where the main objective is to capture the dominant fundamental behavior of the VFM, including the voltage and flux response associated with irreversible demagnetization and remagnetization while preserving low computational cost.

The demagnetization and remagnetization processes are performed without load, i.e., with $i_q = 0$, since the baseline model does not account for PM phase distortion.

The results of the demagnetization test are presented in Fig. 3.4. The motor is first accelerated to 1000 rpm under fully magnetized conditions. A demagnetization current pulse with $i_d = -17$ A, $i_q = 0$ A is then applied and subsequently removed. The resulting dq current and flux linkages are shown for both simulation platforms.

The proposed model closely tracks the FEA results and accurately captures the flux linkage variations. The maximum peak error in λ_d is 6.9% of the nominal flux linkage before the demagnetization pulse. During the demagnetization transient, this error decreases to 3.0%, and further reduces to 2.8% once the demagnetization current is removed.

The initial discrepancy before the demagnetization pulse is attributed to eddy current effects in the FEA model, which slightly demagnetize the PM once the motor starts accelerating. The remaining errors during and after the transients are mainly due to space harmonics in the air-gap flux, which are captured by the FEA model but neglected in the baseline model. This modeling assumption is commonly adopted in

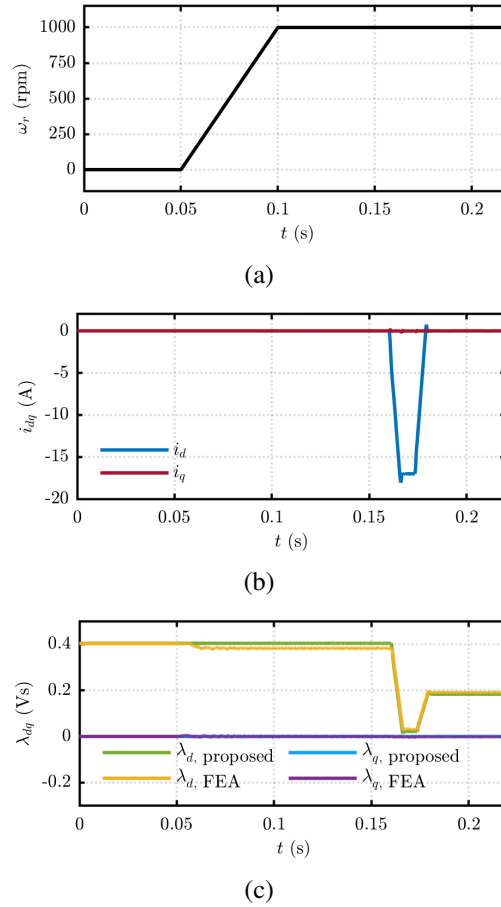


Fig. 3.4 Demagnetization test results with the proposed model compared to FEA: (a) rotor speed, (b) dq currents and (c) dq flux.

control-oriented simulations and is considered acceptable given the achieved level of accuracy.

The model accuracy is further confirmed by the EMF waveforms shown in Fig. 3.5, including detailed views before the demagnetization, during the transient, and after the current pulse release. The proposed model accurately tracks the fundamental component of the phase voltage. However, the back-EMF ripples shown in FEA simulation are not captured, due to the neglect of space harmonics in the modeling approach. This approximation is acceptable for motor control development.

The remagnetization test results are presented in Fig. 3.6. In this case, the VFM accelerated to 1000 rpm under fully magnetized conditions. A demagnetization pulse

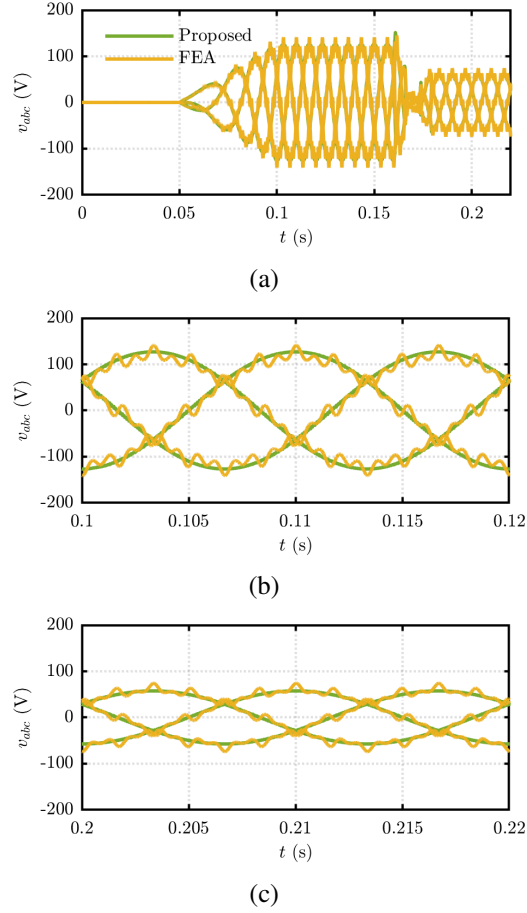


Fig. 3.5 Open circuit voltage waveforms at different stages of the demagnetization process: (a) Complete time waveform, (b) detail before the demagnetization pulse and (c) detail after the demagnetization pulse.

with $i_d = -22.5$ A and $i_q = 0$ A is first applied, followed by a remagnetization pulse with $i_d = 42.5$ A and $i_q = 0$ A. The maximum error in λ_d before remagnetization is 4.1% of the nominal flux linkage, then decreases to 1.9% during the remagnetization transient, and increases to 5.3% after the current pulse is released. These discrepancies are again attributed to space harmonics in the air-gap flux, which are not captured in the proposed model.

The corresponding back-EMF waveforms are shown in Fig. 3.7. Also in this case, the baseline model accurately reproduces the fundamental voltage component and closely follows the FEA results, while neglecting the harmonic ripple content.

The computational efficiency of the baseline model is summarized in Tab. 3.1, where both execution time and storage demand are compared against the FEA-based

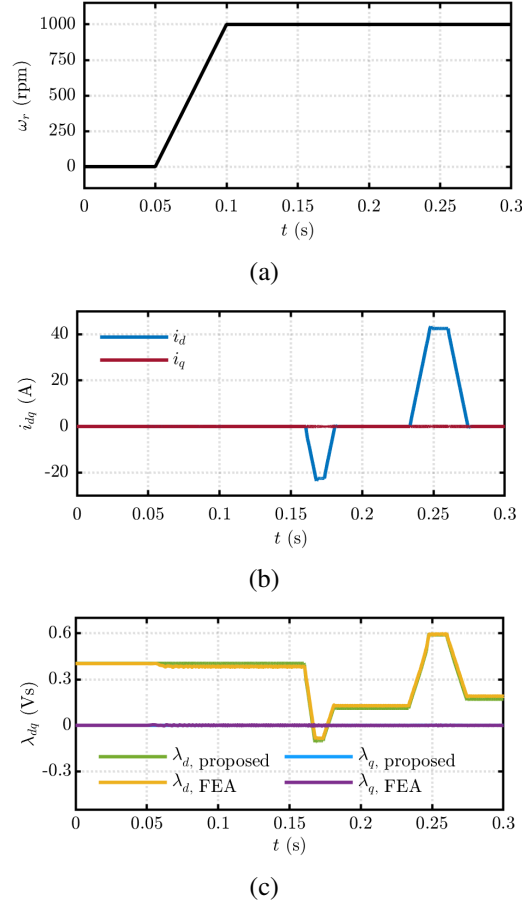


Fig. 3.6 Remagnetization test results with the proposed model compared to FEA: (a) rotor speed, (b) dq currents and (c) dq flux.

simulation. All simulations were performed on the same workstation equipped with a 32 GB RAM, an Intel Core i9-12900 processor @2.50 GHz and a 1 TB NVMe SSD.

Because of the high computational burden of FEA, which requires intensive electromagnetic calculations at each time step and produces large datasets, larger step sizes are implemented in JMAG than in Simulink. Simulations in Simulink are executed with a step time of $2 \mu s$, whereas JMAG simulations use $20 \mu s$ for the demagnetization test and $50 \mu s$ for the remagnetization test. Even under these conditions, the proposed model reduces the computational time by approximately 500 times and lowers the storage demand from the GB range to the MB range. These results confirm that the baseline VFM model enables fast simulation while maintaining high accuracy with respect to the FEA-based methods.

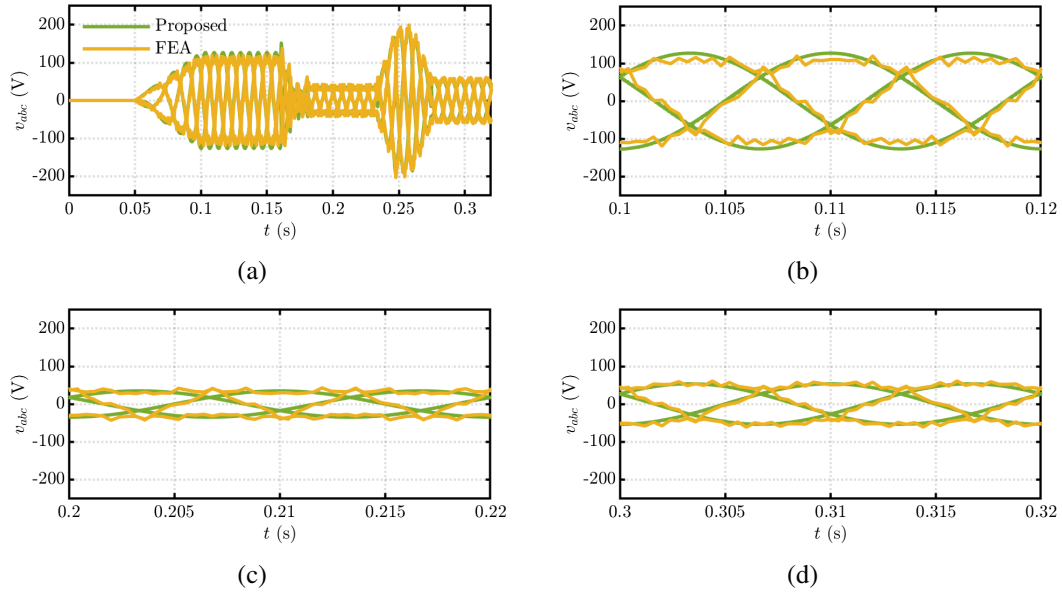


Fig. 3.7 Open circuit voltage waveforms at different stages of the remagnetization process: (a) complete time waveform, (b) detail before the demagnetization pulse, (c) detail after the demagnetization pulse and (d) detail after the remagnetization pulse.

Table 3.1 Model performance assessment: baseline VFM model VS FEA

Modeling Approach	Computational Time		Storage Demand	
	Demag Test	Remag Test	Demag Test	Remag Test
Proposed Simulink Model	≈ 10 s	≈ 14 s	≈ 24 MB	≈ 24 MB
FEA-based (JMAG 23.1)	≈ 75 min	≈ 124 min	≈ 46 GB	≈ 23 GB

It should also be noted that, in the FEA simulations, the phase currents are directly imposed. If the inverter and control system models were also included, as in the Simulink implementation, the computational burden of the FEA-based approach would be even higher.

3.2 Machine Model Including Asymmetric Magnetization

The baseline VFM model introduced in Section 3.1 captures the irreversible demagnetization and remagnetization behavior of LCF PMs. However, when the MS is

manipulated under load conditions ($i_q \neq 0$), the magnetization direction is no longer aligned with the PM reference direction, i.e., the d -axis. This misalignment results in asymmetric magnetization of the PM.

This phenomenon is described by the definition of the complex magnetization state ($\mathbf{MS}$) previously introduced in eq. (2.1) of Section 2.2.2, which accounts for both magnitude (MS) and phase (ϕ_{MS}) variations.

More specifically, MS describes the amplitude of the fundamental PM flux linkage variation and the ϕ_{MS} describes the angular deviation of this fundamental component with respect to the d -axis. The limitation of this representation is that, $\mathbf{MS}$ does not reproduce the detailed local magnetization distribution point by point inside the PMs. Instead, it provides an approximate global equivalent representation of the magnetization distribution, which is acceptable for control oriented model development.

To accurately describe the machine electromagnetic behavior, the flux linkage model must account for PM saturation, cross-saturation effects and asymmetric magnetization:

$$\begin{cases} \lambda_d = \lambda_d(i_d, i_q, \mathbf{MS}) \\ \lambda_q = \lambda_q(i_d, i_q, \mathbf{MS}) \end{cases} \quad (3.6)$$

Starting from the baseline model in Fig. 3.2, an improved model is presented in Fig. 3.8, with the highlighted core blocks accounting for the phase component ϕ_{MS} .

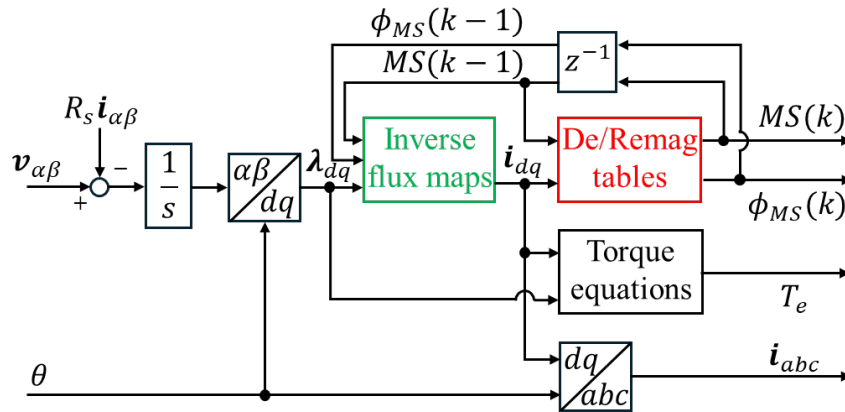


Fig. 3.8 VFM electromagnetic model including asymmetric magnetization.

3.2.1 Inverse Flux Maps

As highlighted by the green block in Fig. 3.8, the inverse flux maps are used to determine the current vector $\mathbf{i}_{dq}$ as a function of the flux linkage vector $\boldsymbol{\lambda}_{dq}$, the magnitude MS and the phase ϕ_{MS} :

$$\begin{cases} i_d = i_d(\lambda_d, \lambda_q, MS, \phi_{MS}) \\ i_q = i_q(\lambda_d, \lambda_q, MS, \phi_{MS}) \end{cases} \quad (3.7)$$

A direct implementation of this relationship would require 4D LUTs, which are not practical for control-oriented models due to their high memory requirements and computational burden, especially in real-time simulation and embedded applications.

To overcome this limitation, the concept of armature flux is introduced. The main idea is to separate the PM contribution from the total flux linkage. In this way, the asymmetric magnetization behavior can be represented more efficiently with reduced dimensionality.

The armature flux vector is defined as:

$$\boldsymbol{\lambda}_{L,dq} = \boldsymbol{\lambda}_{dq} - \mathbf{MS} \cdot \lambda_{m,max} \quad (3.8)$$

More explicitly, the armature flux components can be expressed as:

$$\begin{cases} \lambda_{L,d} = \lambda_d(i_d, i_q, MS) - \lambda_{m,d} \\ \lambda_{L,q} = \lambda_q(i_d, i_q, MS) - \lambda_{m,q} \end{cases} \quad (3.9)$$

Under asymmetric magnetization conditions, the PM contribution is no longer aligned with the d -axis. Therefore, the PM flux has components along both the d and q axes:

$$\begin{cases} \lambda_{m,d} = MS \cdot \lambda_{m,max} \cdot \cos(\phi_{MS}) \\ \lambda_{m,q} = MS \cdot \lambda_{m,max} \cdot \sin(\phi_{MS}) \end{cases} \quad (3.10)$$

This shows that when $\phi_{MS} \neq 0$, the PM contribution introduces a coupling between the d and q axes. By separating this contribution, the armature flux $\boldsymbol{\lambda}_{L,dq}$ captures only the effects related to the stator currents, including magnetic saturation and cross-saturation.

The current vector i_{dq} is then computed using nonlinear inverse armature flux maps:

$$\begin{cases} i_d = i_d(\lambda_{L,d}, \lambda_{L,q}, MS) \\ i_q = i_q(\lambda_{L,d}, \lambda_{L,q}, MS) \end{cases} \quad (3.11)$$

These maps are obtained by manipulating the flux maps introduced in Section 2.2.3 according to (3.9). With this approach, the phase component ϕ_{MS} can be eliminated from the LUTs, reducing the inverse flux map to 3D LUTs and making the model suitable for practical implementation. The internal structure of the inverse flux map block, highlighted in green in Fig. 3.8, is illustrated in Fig. 3.9.

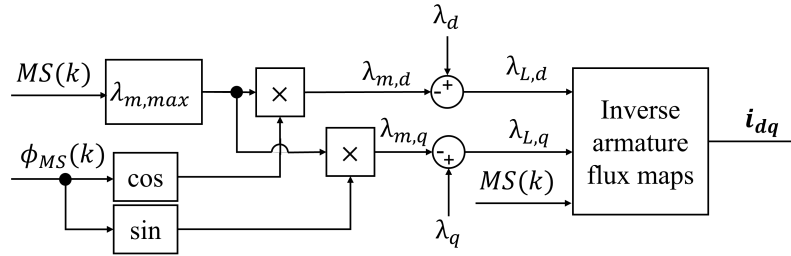
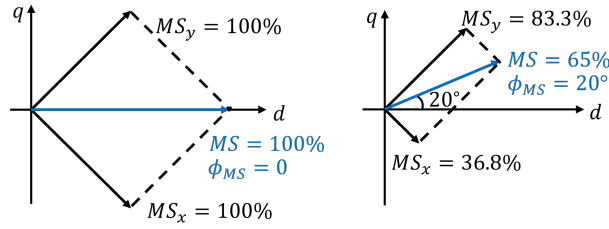


Fig. 3.9 Detailed inverse flux maps block

3.2.2 Two Virtual PM Representation

To describe asymmetric magnetization analytically, a representation based on two virtual PMs is introduced. *It should be noted that these virtual PMs do not correspond to physically separated magnets, they are used as an auxiliary mathematical representation of the fundamental PM magnetization state for control oriented modeling.* Two auxiliary axes, denoted as x and y , are defined at angles of $\pm\frac{\pi}{4}$ with respect to the d -axis. The total PM flux linkage is then decomposed into its projections along these two axes, corresponding to two virtual magnet, denoted as PM_x and PM_y , as shown in Fig. 3.10. Each virtual PM is characterized by its own magnetization state, MS_x and MS_y .

Fig. 3.10 Vector representation of the PM_x, PM_y virtual magnets.

Under this representation, the complex $\mathbf{MS}$ can be expressed as:

$$\begin{aligned} \mathbf{MS} &= \frac{1}{\sqrt{2}} \left(MS_x e^{-j\frac{\pi}{4}} + MS_y e^{j\frac{\pi}{4}} \right) \\ &= \frac{1}{2} \begin{bmatrix} MS_y + MS_x \\ MS_y - MS_x \end{bmatrix} \end{aligned} \quad (3.12)$$

from which the magnitude and phase of the complex MS are represented as:

$$MS = \frac{1}{2} \sqrt{(MS_x + MS_y)^2 + (MS_y - MS_x)^2} \quad (3.13a)$$

$$\phi_{MS} = \tan^{-1} \left(\frac{MS_y - MS_x}{MS_y + MS_x} \right) \quad (3.13b)$$

This representation provides a clear description of asymmetric magnetization. In the symmetric case, shown on the left side of Fig. 3.10, when MS_x and MS_y are both 100 %, the complex $\mathbf{MS}$ is 100% $\angle 0$, i.e., aligned with the d -axis. Conversely, in the asymmetric case, shown on the right side of Fig. 3.10, using eq. (3.13b), different values of MS_x and MS_y equal to 36.8% and 83.3% lead to a complex $\mathbf{MS}$ of 65% $\angle 20^\circ$.

Each virtual PM is associated with its own demagnetization and remagnetization profiles, derived from the complex magnetization maps shown in Fig. 2.9 by applying the inverse of eq. (3.13b). The magnetization maps of PM_x and PM_y can be expressed as:

$$\begin{aligned} MS_x &= \frac{MS \cdot [\cos(\phi_{MS}) + \sin(\phi_{MS})]}{\sqrt{2}} \\ MS_y &= \frac{MS \cdot [\cos(\phi_{MS}) - \sin(\phi_{MS})]}{\sqrt{2}} \end{aligned} \quad (3.14)$$

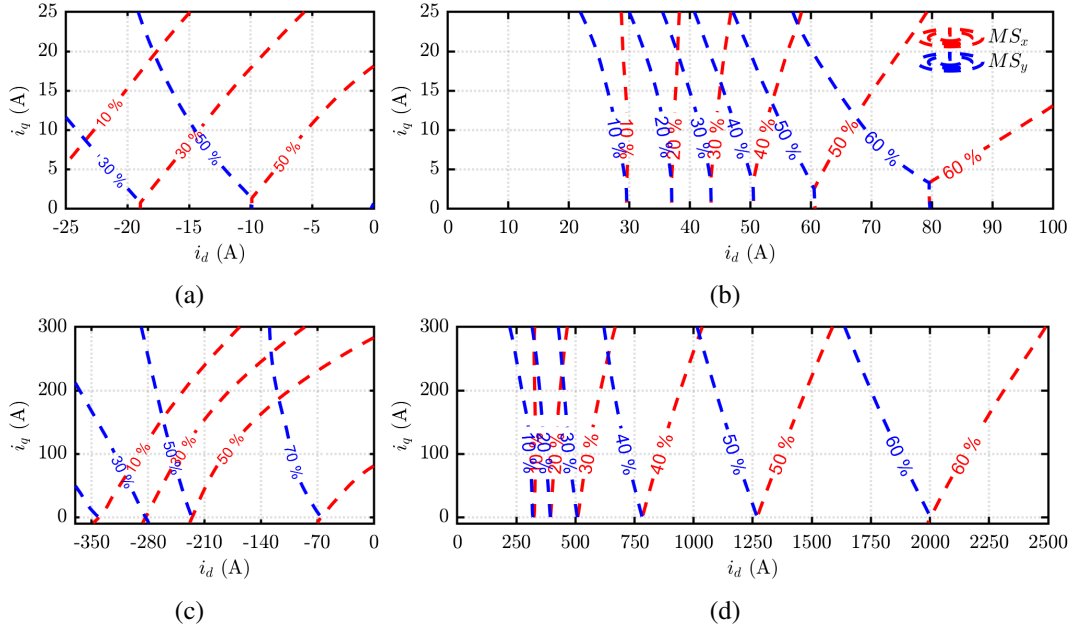


Fig. 3.11 Magnetization maps for the x and y virtual magnets: (a) demagnetization map of the POC motor, (b) remagnetization map of the POC motor, (c) demagnetization map of the traction motor and (d) remagnetization map of the traction motor.

The resulting demagnetization and remagnetization maps of PM_x and PM_y for the POC and traction motor, introduced in Section 2.3 and Section 2.4, are reported in Fig. 3.11.

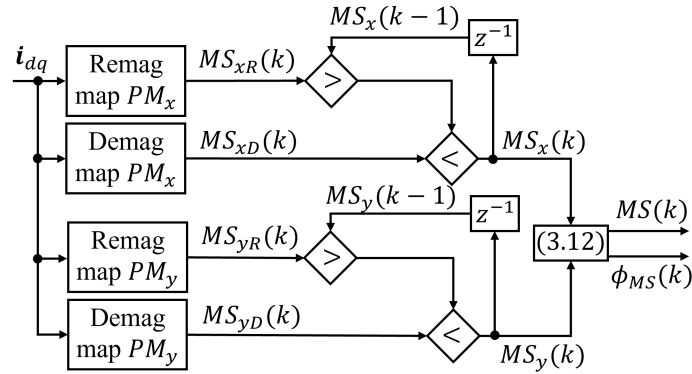


Fig. 3.12 Complex MS updating system.

The corresponding update mechanism of the complex MS is illustrated in Fig. 3.12. The proposed approach extends the MS update system of the baseline model (Fig. 3.3) to account for both x and y components of the magnetization state. Considering the demagnetization process for PM_x , at each time step k , the component

$MS_x(k)$ is updated by comparing its previous value $MS_x(k-1)$ with the output of the demagnetization map $MS_{xD}(k)$:

- if i_{dq} exceeds the corresponding demagnetization contour shown in Fig. 3.11, then $MS_{xD}(k) < MS_x(k-1)$, and therefore $MS_x(k) = MS_{xD}(k)$;
- if i_{dq} remains within the corresponding demagnetization contour shown in Fig. 3.11, then $MS_{xD}(k) \geq MS_x(k-1)$, and therefore $MS_x(k) = MS_x(k-1)$.

A dual mechanism is adopted for remagnetization, again considering PM_x . At each time step k , the remagnetization map provides a value $MS_{xR}(k)$:

- if i_{dq} exceeds the corresponding remagnetization contour shown in Fig. 3.11, then $MS_{xR}(k) > MS_x(k-1)$, and therefore $MS_x(k) = MS_{xR}(k)$;
- if i_{dq} remains within the corresponding demagnetization contour shown in Fig. 3.11, then $MS_{xR}(k) \leq MS_x(k-1)$, and therefore $MS_x(k) = MS_{xR}(k)$.

The same update principle is applied to $MS_y(k)$. The updated $MS_x(k)$ and $MS_y(k)$ are then combined according to eq. (3.13b) to obtain the magnitude and phase of the complex $\mathbf{MS}(k)$.

3.2.3 Model Validation against FEA Simulation

The accuracy of the proposed VFM model including asymmetric magnetization is assessed through comparison with both the baseline model and FEA model developed in JMAG. Both the baseline and proposed models are implemented in MATLAB/Simulink under closed-loop current control. The resulting phase currents from the proposed model are directly imposed to the FEA model in order to reduce the computational burden of the electromagnetic simulations.

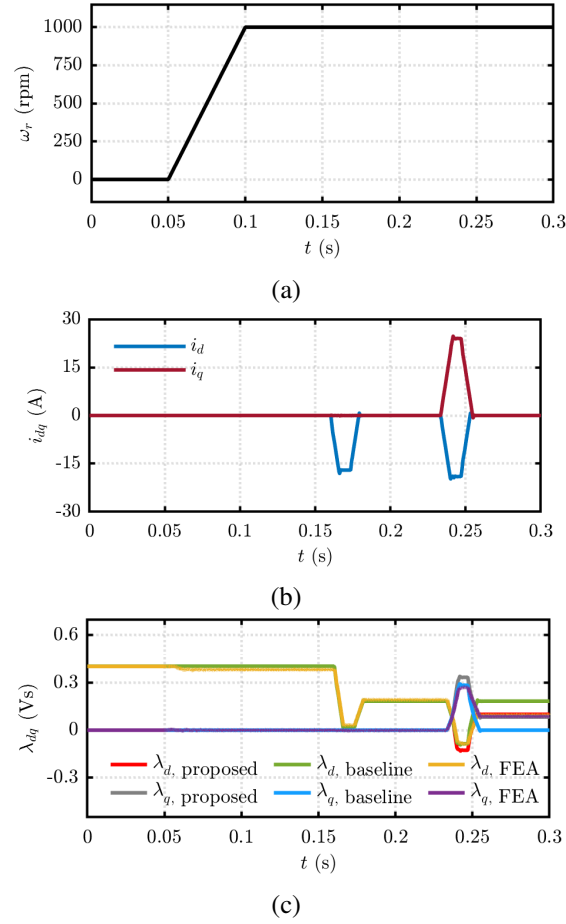


Fig. 3.13 Demagnetization test results with the proposed model compared to FEA: (a) rotor speed, (b) dq currents and (c) dq flux.

The POC motor introduced in Section 2.3 is used for validation. The first validation scenario focuses on demagnetization and is reported in Fig. 3.13. The simulation starts with $\mathbf{MS} = 100\% \angle 0$ and the machine is accelerated to 1000 rpm. A symmetrical demagnetization pulse with $i_d = -17$ A and $i_q = 0$ is first applied, followed by an asymmetrical pulse with $i_d = -19$ A and $i_q = 24$ A.

During the symmetrical demagnetization case, both the baseline and proposed models exhibit similar results, as both models can accurately represent the symmetrical magnetization conditions. Significant differences arise under asymmetric demagnetization. After the second demagnetization current pulse, the baseline model shows substantial deviations, with errors of 52.5% and 43.5% in λ_d and λ_q , respectively, with respect to their nominal values. In contrast, the proposed model

maintains high accuracy, with corresponding errors in λ_d and λ_q limited to 1.5% and 1.8%, respectively.

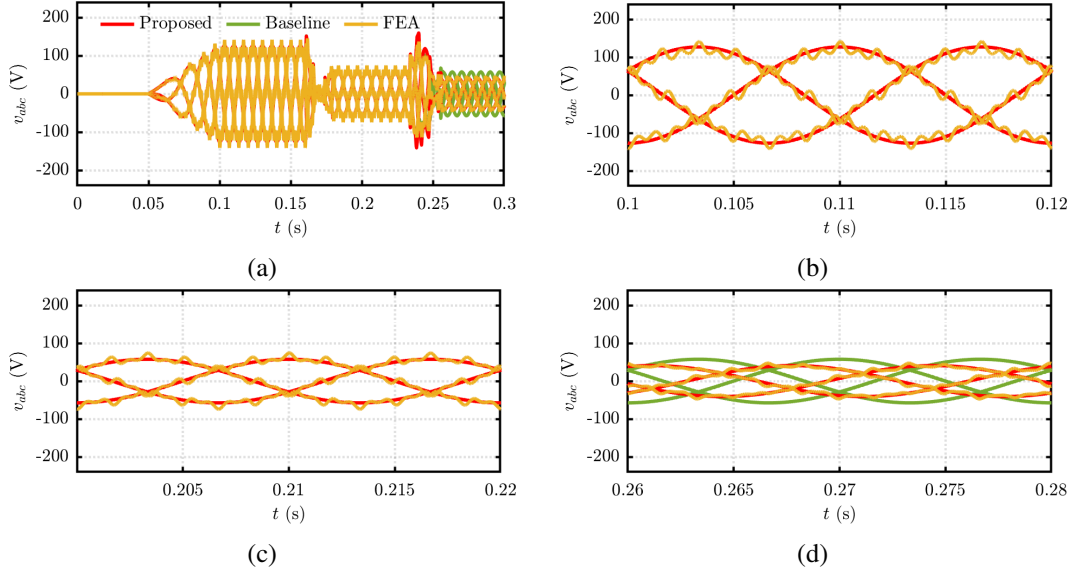


Fig. 3.14 Open circuit voltage waveforms at different stages of the demagnetization process: (a) complete time waveform, (b) detail before the symmetrical demagnetization pulse, (c) detail after the demagnetization pulse and (d) detail after the asymmetrical remagnetization pulse.

Further evidence is provided by the back-EMF waveforms shown in Fig. 3.14. The baseline model accurately reproduces the EMF during the symmetrical condition, but fails to capture the asymmetric demagnetization effect, leading to a phase shift of approximately 40° and a magnitude overestimation of about 40%. Conversely, the proposed model closely tracks the FEA results throughout the entire simulation. Minor discrepancies are observed only in the harmonic content, which is neglected in both Simulink models. This simplification remains acceptable for control-oriented modeling.

A second validation scenario focuses on remagnetization and is shown in Fig. 3.15. The machine is initialized with $\mathbf{MS} = 100\% \angle 0$ and accelerated to 1000 rpm. A symmetrical demagnetization pulse with $i_d = -23$ A and $i_q = 0$ is first applied, followed by two remagnetization pulses: a symmetrical pulse with $i_d = 43$ A, $i_q = 0$ and an asymmetrical pulse with $i_d = 50$ A, $i_q = 13$ A.

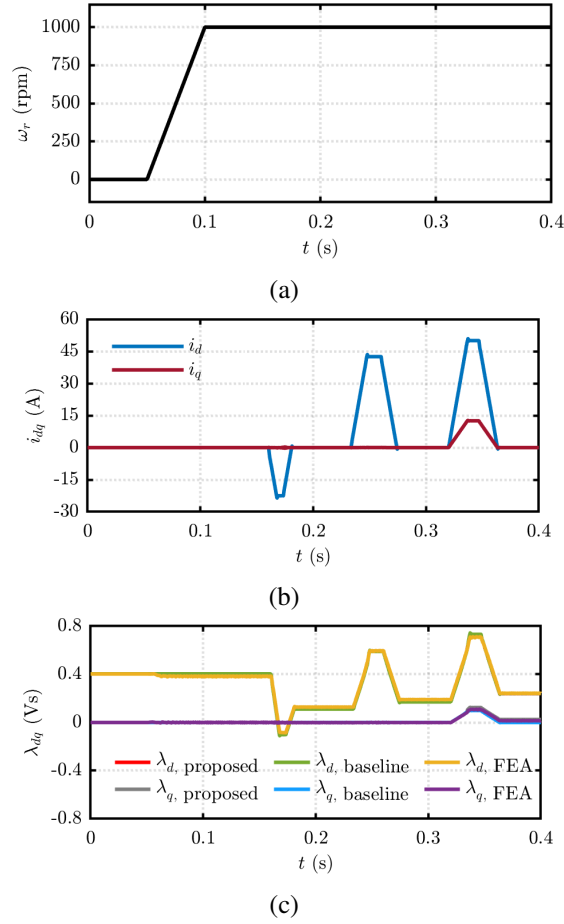


Fig. 3.15 Remagnetization test results with the proposed model compared to FEA: (a) rotor speed, (b) dq currents and (c) dq flux.

As in the demagnetization case, both models show comparable performance under symmetrical conditions. After the asymmetrical remagnetization, the maximum error in λ_d remains below 1% for both models. For λ_q , the baseline model exhibits an error of 7.0%, while the proposed model reduces this error to 5.6%.

The smaller discrepancy between the two models in the remagnetization case is attributed to the reduced phase variation of the complex **MS** maps in the first quadrant, compared to the second quadrant associated with demagnetization (see Fig. 2.16). The back-EMF waveforms shown in Fig. 3.16 confirm that only limited phase differences appear after the asymmetrical remagnetization pulse.

The discrepancies between the proposed model and FEA results are again mainly due to space harmonics in the air-gap flux, which are included in the FEA model

but neglected in the proposed model. This assumption is generally acceptable for control-oriented simulations.

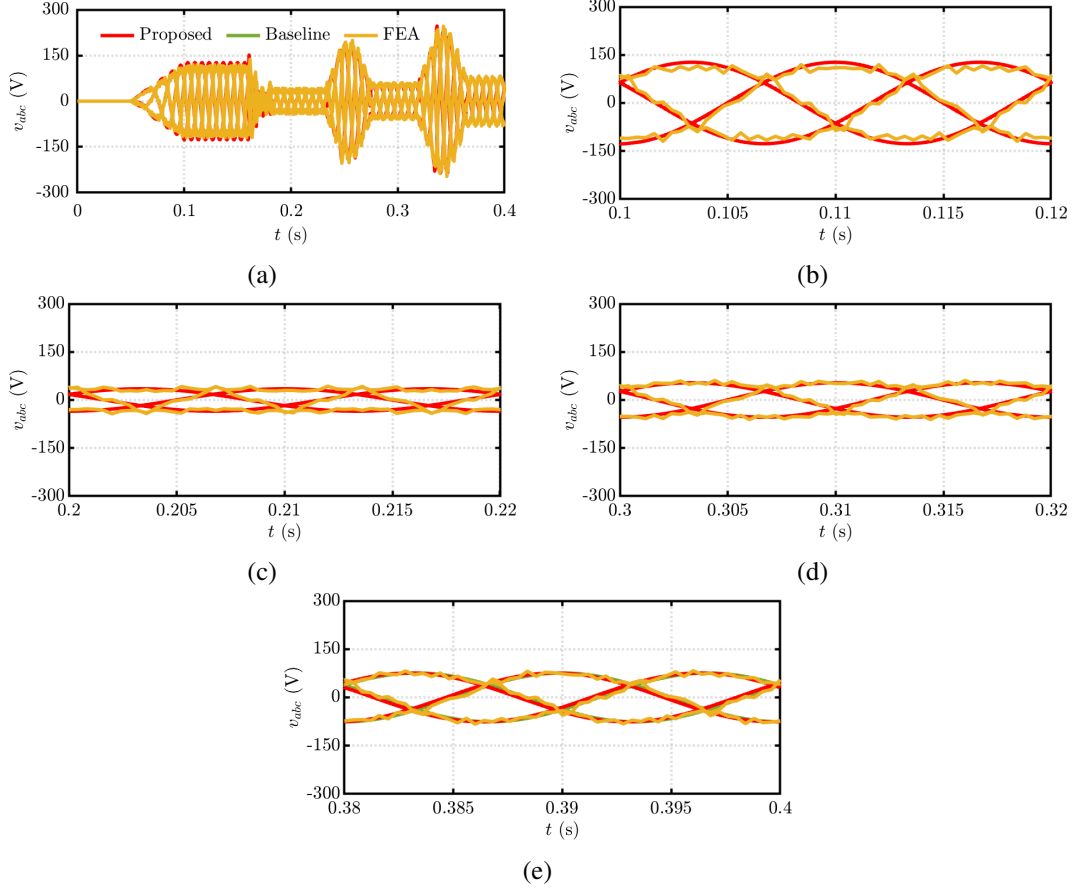


Fig. 3.16 Open circuit voltage waveforms at different stages of the remagnetization process: (a) complete time waveform, (b) detail before the demagnetization pulse, (c) detail after the demagnetization pulse, (d) detail after the symmetrical remagnetization pulse and (e) detail after the asymmetrical remagnetization pulse.

The computational burden of the proposed model is summarized in Tab. 3.2. All simulations are performed on the same workstation used in the Section 3.1.3. The Simulink model was executed ten times, and the reported simulation time corresponds to the average over all runs. Because of the large time and memory demand of the FEA model, larger step sizes are used in JMAG than in Simulink. Specifically, Simulink uses a fixed step of $2 \mu\text{s}$, while JMAG uses $20 \mu\text{s}$ for the demagnetization test and $50 \mu\text{s}$ for the remagnetization test.

Despite the larger time step used in the FEA simulations, the proposed model reduces the computational time by approximately 500 times and lowers the storage

Table 3.2 Model performance assessment: improved VFM model VS FEA

Modeling Approach	Computational Time		Storage Demand	
	Demag Test	Remag Test	Demag Test	Remag Test
Proposed Simulink Model	≈ 11 s	≈ 14 s	≈ 27 MB	≈ 27 MB
FEA-based (JMAG 23.1)	≈ 75 min	≈ 124 min	≈ 46 GB	≈ 23 GB

requirement from the GB range to the MB range. These results confirm that the proposed model significantly improves computational efficiency while preserving high fidelity with respect to FEA.

3.3 Model Validation in Real-Time

To assess the practical applicability of the proposed VFM model, its performance is further evaluated in a real-time environment. [In this chapter, the HiL implementation serves three purposes. First, it verifies the real-time computational feasibility of the proposed control oriented model on the HiL hardware platform. Second, it provides an additional validation level by comparing the real-time and offline model responses against FEA results during *MS* variation transients. Third, it enables system-level demonstration, such as vehicle acceleration and driving cycle simulations.](#) While the previous validation against FEA establishes the model accuracy, HiL testing further enables computationally demanding scenarios, such as long-duration drive tests and full driving cycles, which are impractical with offline or FEA-based simulations.

In the HiL setup, the inverter and motor models are implemented on a PLEXIM RT Box2, while the control algorithm is executed on an STM32 G474 nucleo board. The hardware setup is shown in Fig. 3.17.

The RT box includes three real-time Cortex-A53 CPUs. The proposed VFM electromagnetic model is sufficiently light to be executed entirely on a single CPU. A second CPU is assigned to the PWM switching signals and to the circuital models of the voltage source inverter and electrical machine, with the circuital part executed on the FPGA at a step size of 4 ns. The third CPU handles the reference signals transmitted to the controller, the output signals of the electromagnetic model, and data export units. The complete real-time model layout is shown in Fig. 3.18.

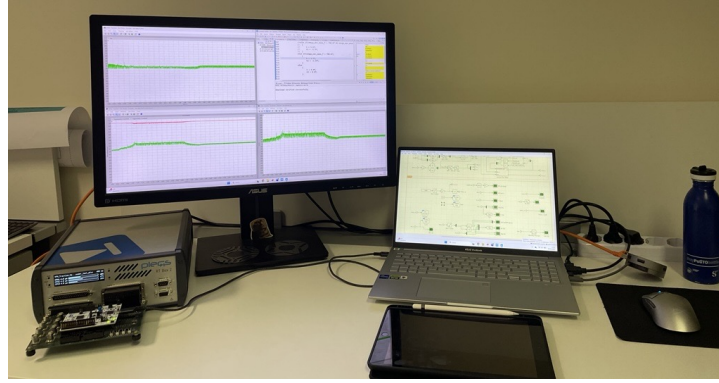


Fig. 3.17 Hardware setup of the real-time environment, with PLEXIM RT Box 2 and STM32 G4 MCU.

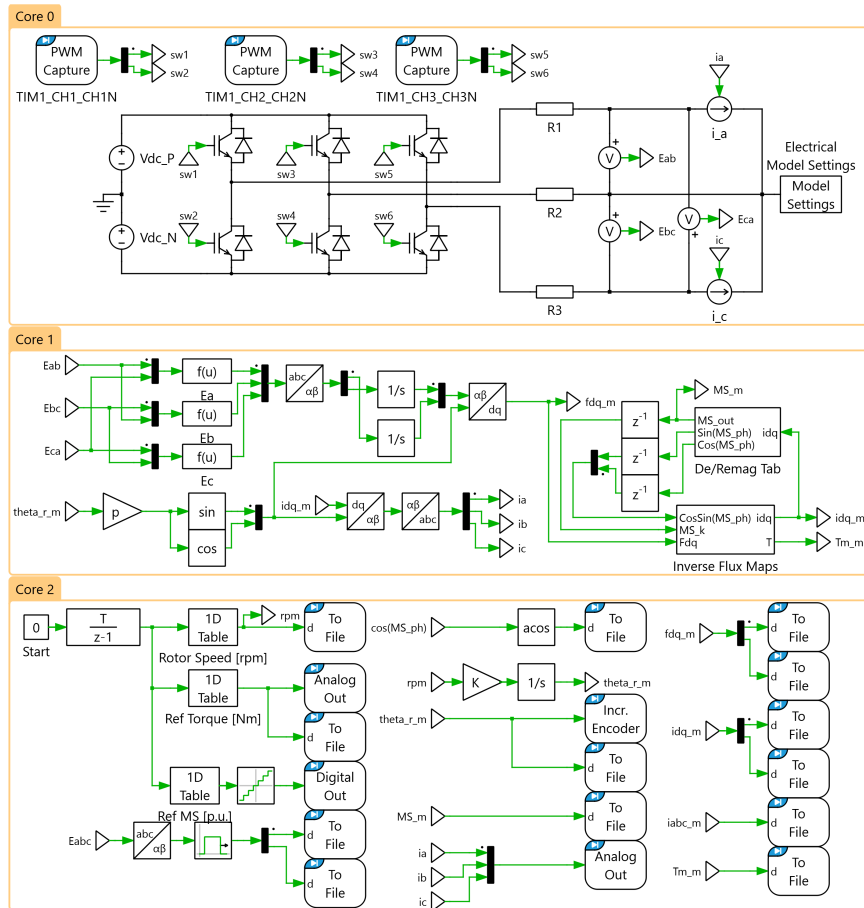


Fig. 3.18 Real-time model layout in PLECS.

The HiL environment emulates a realistic drive system. Three analog output channels of the RT box generate the phase current signals i_{abc} . These channels are

physically connected to the ADC inputs of the STM32 MCU, reproducing realistic current acquisition including sampling, quantization, and conversion delays. Rotor position feedback is emulated through the RT box encoder block, which generates two quadrature encoder signals together with the index signal. The inverter switching commands are received through PWM input blocks connected to the MCU outputs. The control frequency is set as 20 kHz and the dead-time is configured in the MCU.

Both the POC motor described in Section 2.3 and traction motors described in Section 2.4 are implemented in the real-time environment. Demagnetization and remagnetization transients are first evaluated against offline PLECS and JMAG FEA models. A WLTC Class 3 driving cycle is then simulated in real-time using the traction motor integrated within a full vehicle model, with *MS* managed through a dedicated switching strategy.

3.3.1 Real-time Model Validation of POC Motor

The POC motor is first used to verify that the proposed model can be executed in real-time while preserving the same electromagnetic behavior observed in the offline simulation and in the FEA benchmark. For this purpose, the motor is implemented in the real-time environment and validated against the offline PLECS model and the JMAG FEA model. In both the HiL and offline simulations, the motor is operated under identical closed-loop current control. The phase currents obtained from the offline simulation are imposed on the FEA model to ensure a fair comparison.

A demagnetization test is shown in Fig. 3.19. The simulation starts with $\mathbf{MS} = 100\% \angle 0$ and the machine is accelerated to 1000 rpm. A symmetrical demagnetization pulse with $i_d = -10$ A and $i_q = 0$ is first applied, followed by an asymmetrical pulse with $i_d = -10$ A and $i_q = 10$ A. The HiL model accuracy can be verified by observing λ_{dq} . The maximum peak error is 0.5% of the nominal flux linkage in λ_d under symmetrical conditions, increases to 5.0% and 7.5% in λ_d and λ_q during the asymmetrical pulse, and reduces to 2.6% and 0.7% at steady state.

A remagnetization test is presented in Fig. 3.20. The VFM is initialized at $\mathbf{MS} = 100\% \angle 0$ and accelerated to 1000 rpm, then demagnetized symmetrically with $i_d = -25$ A and $i_q = 0$. This is followed by two remagnetization current pulses: a symmetrical pulse with $i_d = 45$ A and $i_q = 0$, and an asymmetrical pulse with $i_d = 55$ A, $i_q = 18$ A. The model accuracy is again verified through λ_{dq} . The peak error is 5.1%

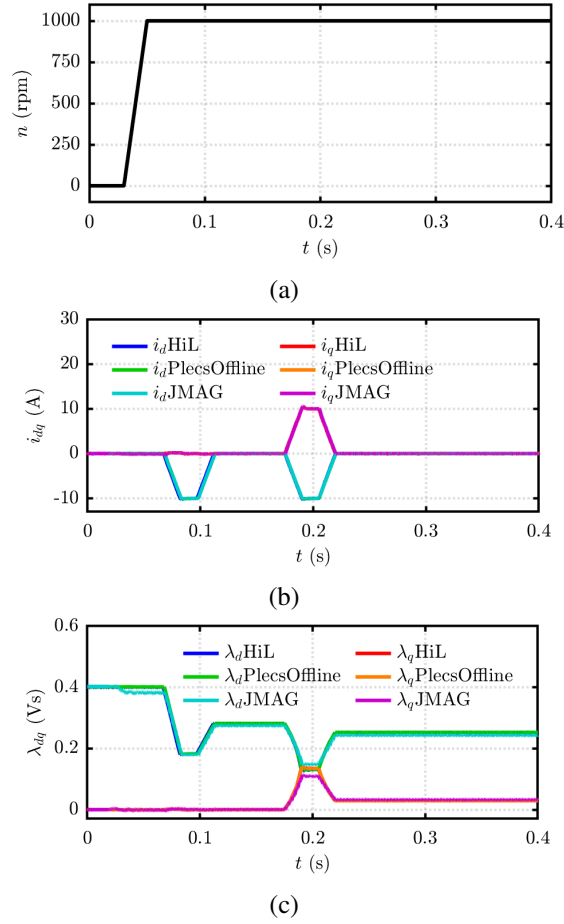


Fig. 3.19 Demagnetization test simulated with HiL, offline and FEA: (a) rotor speed, (b) dq currents and (c) dq fluxes.

in λ_d for the symmetrical pulse, reaches 9.5% and 3.9% in λ_d and λ_q during the asymmetrical pulse, and decreases to 2.6% and 1.3% at steady state.

In both tests, the λ_{dq} obtained in HiL coincide with those of the offline model and closely match the FEA results, with negligible steady-state error and limited transient discrepancy. These deviations are attributed to space harmonics in the air-gap flux, which are captured by the FEA model but neglected in the real-time formulation. This assumption is common and acceptable in control-oriented real-time simulations.

The computational time required by the different models is shown in Tab. 3.3. The PLECS offline model and the JMAG FEA simulations are performed on the same workstation described in Section 3.1.3. The HiL and offline models operate with a step time of $2.5 \mu s$, whereas the JMAG model uses a step time of $250 \mu s$

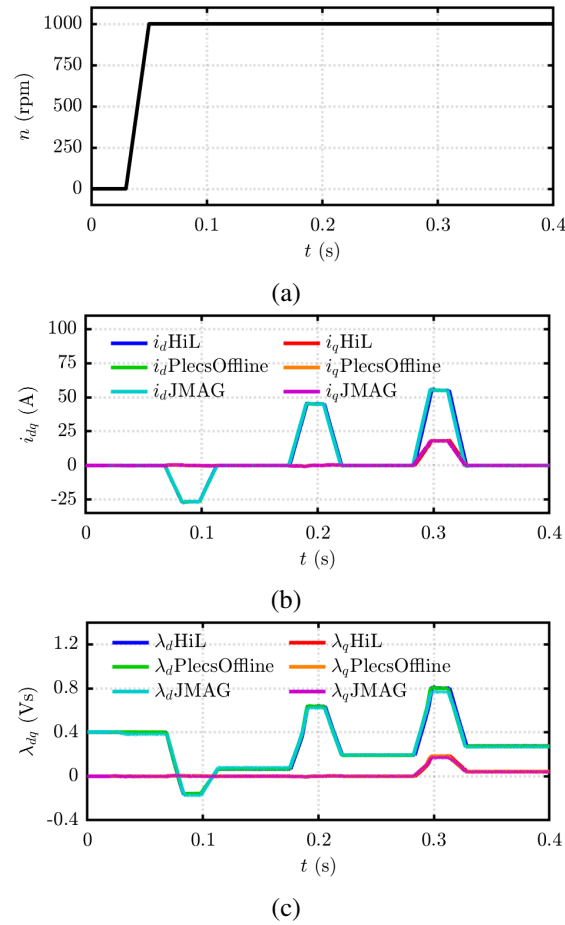


Fig. 3.20 Remagnetization test simulated with HiL, offline and FEA: (a) rotor speed, (b) dq currents and (c) dq fluxes.

to reduce the computational burden. Even with a step time 100 times larger, the FEA model requires 52 minutes to complete a 0.4 s simulation, while the offline model completes the same test in a few seconds. The HiL model further reduces the effective computation time by running in real time. For longer simulations, the cumulative cost of offline simulations becomes impractical.

3.3.2 Real-time Model Validation of Traction Motor

The same real-time validation of Section 3.3.1 is then repeated on the traction motor to further confirm the computational feasibility and model accuracy also for a machine of high current and torque rating. Again, the HiL results are compared with

Table 3.3 Computational time comparison with different modeling approaches for a 0.4 s test using POC motor.

Modeling Approach	Computational Time
HiL	Real Time
Offline	≈ 4 s
FEA	≈ 52 min

those of the offline model and JMAG FEA model. The HiL model and the offline simulations both operate under identical closed-loop current control settings.

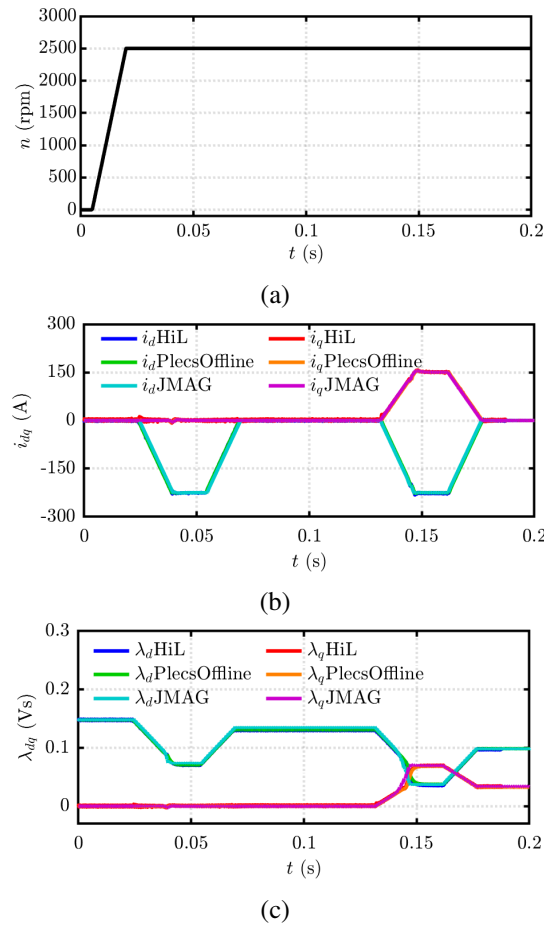


Fig. 3.21 Demagnetization test simulated with HiL, offline and FEA: (a) rotor speed, (b) dq currents and (c) dq fluxes.

A demagnetization test is shown in Fig. 3.21. The simulation is initialized with $MS = 80\% \angle 0$ and the machine is accelerated to 2500 rpm. Two consecutive demag-

netization current pulses are applied: first, a symmetrical pulse with $i_d = -225$ A and $i_q = 0$, followed by an asymmetrical pulse with $i_d = -225$ A and $i_q = 150$ A. The dq -axis flux linkages obtained from the HiL experiment are indistinguishable from those of the offline model and closely match the FEA results. This confirms that the proposed HiL model accurately reproduces both symmetrical and asymmetrical demagnetization transients.

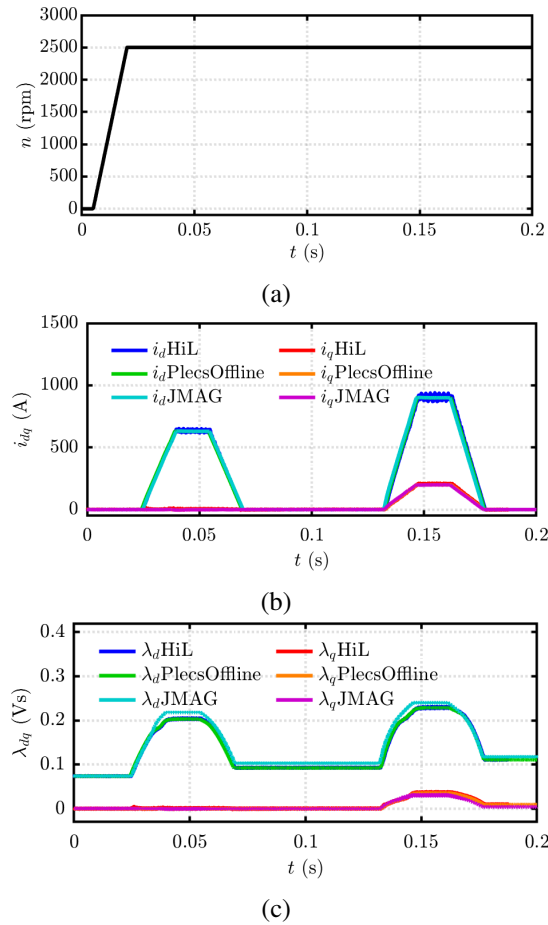


Fig. 3.22 Remagnetization test simulated with HiL, offline and FEA: (a) rotor speed, (b) dq currents and (c) dq fluxes.

A remagnetization test is reported in Fig. 3.22. In this case, the VFM is initialized at $\mathbf{MS} = 40\% \angle 0$ and accelerated to 2500 rpm. Two remagnetization current pulses are sequentially injected: a symmetrical pulse with $i_d = 625$ A and $i_q = 0$, followed by an asymmetrical pulse with $i_d = 900$ A and $i_q = 200$ A. The dq currents measured in the HiL experiment exhibit slightly higher noise than in the offline simulations, due to the hardware current acquisition process combined with the large current

amplitudes. Despite this, the resulting dq -axis flux linkages from the real-time HiL test overlap with those obtained from the offline model.

A limited discrepancy is observed with respect to the FEA results. The maximum transient deviation is 8.9% of the nominal flux in λ_d for the symmetrical pulse, while during the asymmetrical pulse the peak errors are 6.4% in λ_d and 3.5% in λ_q . These errors reduce to 4.5% and 3.2% at steady state after the two pulses. As in the previous cases, these differences are mainly due to space harmonics in the airgap flux distribution, which are captured by the FEA model but neglected in the HiL formulation.

Table 3.4 Computational time comparison with different modeling approaches for a 0.2 s test using traction motor.

Modeling Approach	Computational Time
HiL	Real Time
Plecs Offline	≈ 2 s
FEA	≈ 19 min

The computational time required by the different models is reported in Tab. 3.4. The HiL and offline models operate with a step time of $3.333 \mu\text{s}$, whereas the JMAG model uses a step time of $333.3 \mu\text{s}$. Even with this step size, the FEA model requires 19 minutes to complete a 0.2 s simulation, while the offline model finishes in a few seconds. The proposed HiL model bridges this gap and extends the validation capability to long-duration tests, such as full driving cycles, which are impractical or impossible to perform in conventional offline environments.

3.3.3 System-Level Demonstration Using Passenger Vehicle

In this section, the HiL environment is used for system-level demonstration, showing that the proposed VFM can be integrated within a full vehicle simulation framework and used for realistic applications. An acceleration test and driving cycle analysis with online MS variations are performed in the HiL environment, using the traction motor integrated within a full vehicle model. The main vehicle parameters are summarized in Tab. 3.5.

Table 3.5 Vehicle data

Vehicle total mass	2000 kg
Max Vehicle speed	180 km/h
Wheel diameter	0.6934 m
Rolling resistance coefficient	0.006
Drag coefficient	0.28
Gear efficiency	0.97
Gear ratio	11

The acceleration test is performed under maximum torque conditions, with a target maximum speed of 180 km/h. Three different *MS* management strategies are considered:

- vehicle with fixed *MS* at 80%;
- vehicle with fixed *MS* at 60%;
- vehicle with variable *MS* switching between 60% and 80%, based on the maximum available torque according to the torque–speed characteristics shown in Fig. 2.22.

The vehicle initially accelerates at *MS* = 80%, and a demagnetization to *MS* = 60% is performed above 8000 rpm, i.e., when *MS* = 60% provides a higher torque limit than the *MS* = 80% condition. This test aims to evaluate the acceleration performance of the vehicle equipped with the traction VFM under different *MS* conditions. The results are presented in Fig. 3.23.

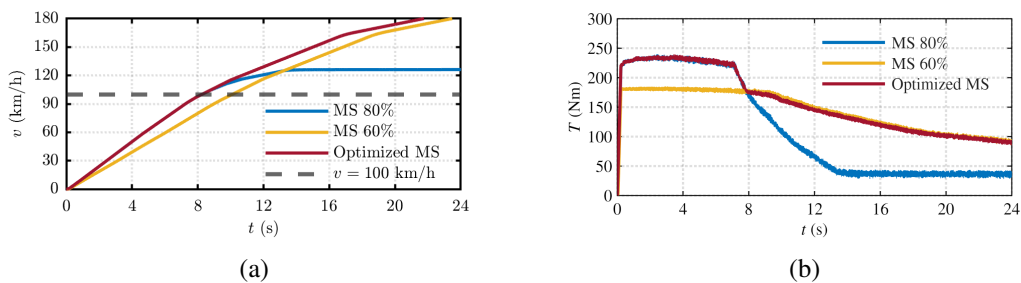


Fig. 3.23 Results of real-time 0-180 km/h acceleration test. Blue: *MS* fixed at 80%; Yellow: *MS* fixed at 60%; Red: *MS* switches between 60% and 80% for optimized peak torque. (a) vehicle speed and (b) e-motor torque.

The 0–100 km/h acceleration time is 10.1 s for the $MS = 60\%$ case, while it is reduced to 8.1 s when operating at $MS = 80\%$, due to the higher torque capability at low speed. However, at $MS = 80\%$, the vehicle acceleration becomes limited above 100 km/h, and the target maximum speed of 180 km/h is not reached. In contrast, the $MS = 60\%$ configuration enables the vehicle to reach the maximum speed in 23.5 s.

By adopting the variable MS strategy, the vehicle benefits from both the high torque capability at $MS = 80\%$ and the extended speed range at $MS = 60\%$. As a result, the vehicle reaches 100 km/h in 8.1 s and achieves the maximum speed of 180 km/h in 21.7 s.

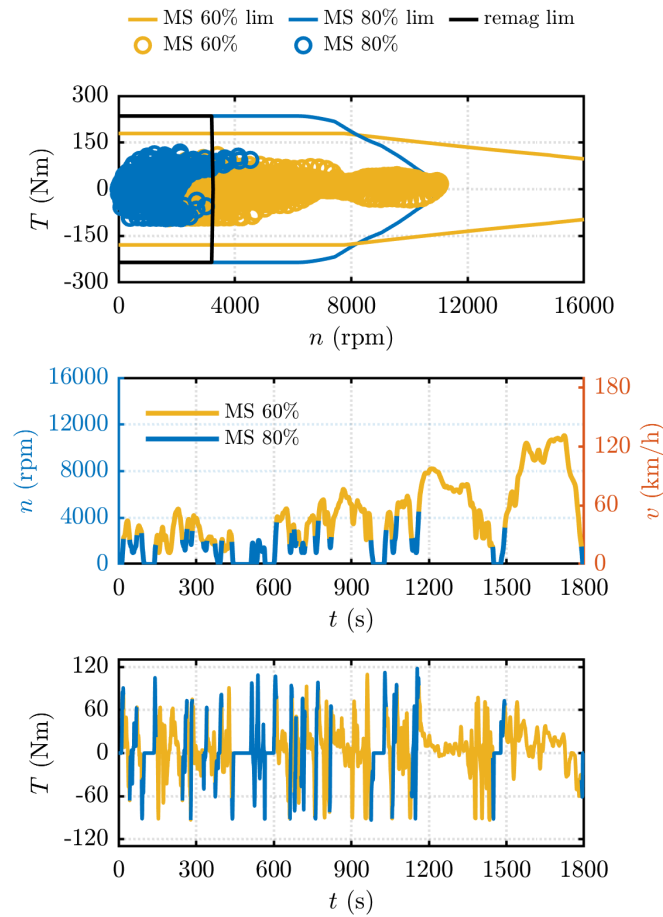


Fig. 3.24 Reference WLTC driving cycle torque and speed with MS switching between 60% and 80%.

The second test consists of a WLTC Class 3 driving cycle. Also in this case, three MS strategies are compared:

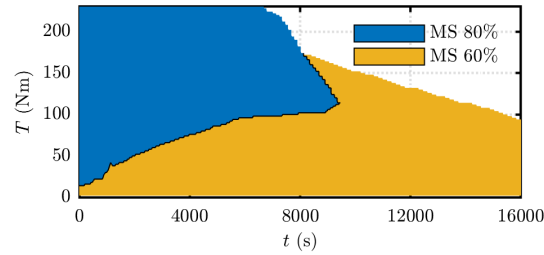


Fig. 3.25 Torque-speed profile of the *MS* selection based on the highest efficiency between *MS* 60% and 80%.

- vehicle with fixed *MS* at 80%;
- vehicle with fixed *MS* at 60%;
- vehicle with variable *MS* switching between 60% and 80%, where the optimal *MS* is selected based on maximum efficiency according to the efficiency maps shown in Fig. 2.22.

The reference torque and speed trajectories of the driving cycle are shown in Fig. 3.24, obtained offline using the variable *MS* strategy. The corresponding *MS* selection is illustrated in Fig. 3.25, where the operating points at *MS* = 80% and *MS* = 60% are highlighted. As expected, *MS* = 80% is preferred in the high-torque, low-speed region, whereas *MS* = 60% is more efficient in the medium-to-high speed and low-to-medium torque region.

The black line in Fig. 3.24 indicates the region where remagnetization to *MS* = 80% is feasible, considering the inverter voltage constraint. This limitation arises from the significantly higher current required for remagnetization compared to demagnetization. As a result, remagnetization can only be performed below approximately 3200 rpm. The methodology used to determine this feasible region will be presented in chapter 5.

The real-time simulation results for the variable *MS* strategy are reported in Fig. 3.26. The motor torque closely follows the reference profile, while the *MS* is dynamically adjusted between 60% and 80%. Limited torque deviations are observed during remagnetization transients, remaining within acceptable ranges.

The motor energy losses obtained under the different *MS* strategies are compared in Fig. 3.27. When *MS* is fixed at 80%, the WLTC cycle cannot be fully completed due to the limited capability in the high speed region, and the simulation is interrupted

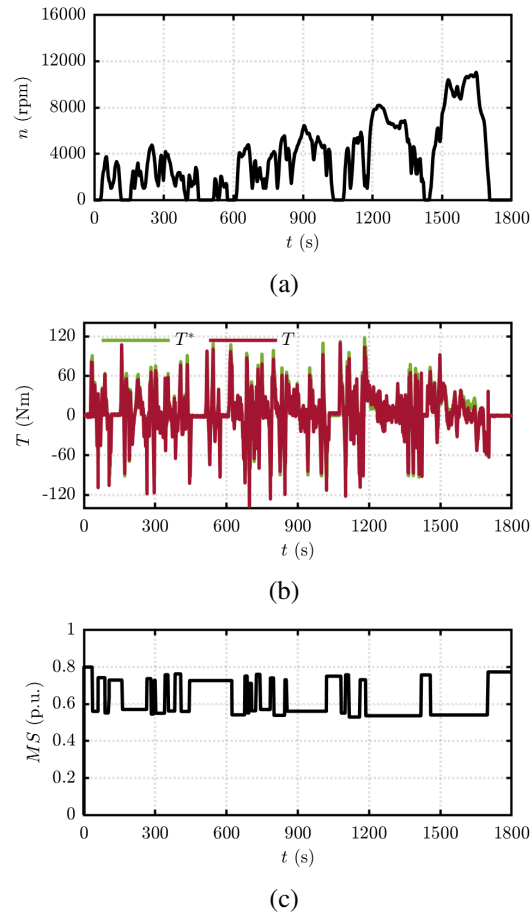


Fig. 3.26 Results of real-time WLTC test: (a) motor speed, (b) torque and (c) MS .

at 1651 s. Up to this point, the total energy loss is 989 kJ. In contrast, operating at fixed $MS = 60\%$ reduces the total energy loss to 948 kJ, while allowing the full cycle to be completed.

The variable MS strategy further reduces the total energy loss to 925 kJ, corresponding to a reduction of approximately 7% with respect to the fixed $MS = 80\%$ case and about 2.5% with respect to the fixed $MS = 60\%$ case. This improvement is mainly due to the fact that most operating points in the WLTC cycle fall within the medium-speed and medium-to-low torque region, where $MS = 60\%$ provides higher efficiency.

This case study demonstrates a realistic HiL application of the proposed VFM model, highlighting its capability to be integrated within a full vehicle simulation framework. This implementation would not be feasible using offline or FEA models,

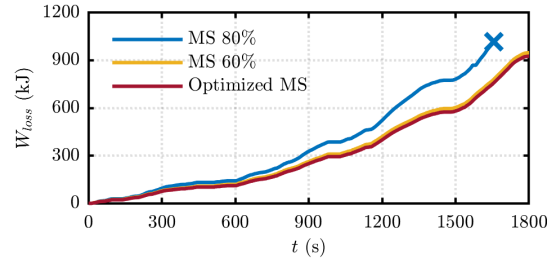


Fig. 3.27 Motor total loss during the WLTC test. Blue: *MS* fixed at 80%; Yellow: *MS* fixed at 60%; Red: *MS* switches between 60% and 80% for optimized efficiency.

due to their prohibitive computational burden and lack of real-time execution capability. The proposed model enables fast and stable simulations while preserving the essential nonlinear characteristics of the machine, making it suitable for system-level analysis, control development and HiL validation.

3.4 Chapter Conclusion

This chapter presented control-oriented modeling frameworks for VFMs suitable for both offline simulation and real-time implementation. A baseline model was first introduced, capable of reproducing irreversible demagnetization and remagnetization phenomena by combining inverse flux maps with demagnetization and remagnetization tables. The model was validated against transient FEA simulations and shown to achieve high accuracy with a computational effort several orders of magnitude lower than that of FEA.

The baseline model was then extended to account for asymmetric magnetization under load conditions. By introducing the complex magnetization state, the armature-flux formulation, and the two virtual PM representation, the proposed model was able to capture both the magnitude and phase variations of the magnetization state while preserving a control-oriented structure suitable for discrete-time implementation. The improved model showed a clear accuracy advantage over the baseline formulation in the presence of asymmetric demagnetization and remagnetization scenarios.

Finally, the proposed model was implemented in a real-time HiL environment and validated on both the POC motor and the traction motor. The HiL results closely matched the offline and FEA simulations, while enabling long-duration and system-level tests that are impractical with conventional offline or FEA-based approaches.

In particular, the full-vehicle acceleration and WLTC case studies demonstrated that the proposed model can be effectively used for realistic drive-cycle analysis and *MS* management strategies in traction applications.

The main outcomes of this chapter can be summarized as follows:

- a baseline VFM model was developed to represent irreversible demagnetization and remagnetization with low computational cost;
- the baseline model was validated against FEA and shown to provide high accuracy with a computational effort several orders of magnitude lower than that of FEA-based simulations;
- the model was extended to include asymmetric magnetization under load by introducing the complex magnetization state, the armature flux formulation and the two virtual PM representation;
- the improved model shows a clear accuracy advantage over the baseline one in the presence of asymmetric demagnetization and remagnetization transients;
- the proposed modeling framework was successfully implemented in a real-time HiL environment and validated on both the POC motor and the traction motor;
- the HiL implementation enabled long-duration and system-level simulations, including vehicle acceleration and WLTC case studies, which would be impractical with offline or FEA-based approaches.

Overall, the results of this chapter confirm that the proposed VFM control simulation model provides an effective compromise between physical fidelity and computational efficiency, thereby establishing a suitable platform for the control design developments presented in the following chapter.

Chapter 4

Control Strategies

The regulation of the MS is a key functionality of VFMs. By intentionally injecting dedicated dq current pulses, the machine can be demagnetized or remagnetized, thereby modifying the PM flux linkage. This capability allows the maximum torque-speed profile to be modified and enables efficiency optimization over a wide operating range. However, when MS regulation is performed during machine operation, it can introduce significant torque perturbations, causing user discomfort and affecting controllability of the drive. In traction applications, such disturbances would be perceived by passengers as undesirable jerks.

As previously introduced in Section 2.2.4, in order to maintain torque during MS variation under load, the machine must exhibit nearly horizontal, smooth and continuous iso-torque lines across both the first and second quadrant. Once this condition is satisfied, a dedicated control strategy is required to move the dq current vector:

- during remagnetization transients, i_{dq} moves from the MTPA curve to the corresponding remagnetization operating point in the first quadrant, according to the remagnetization map shown in Fig. 2.9;
- during demagnetization transients, i_{dq} moves from the MTPA curve to the corresponding demagnetization operating point in the second quadrant, according to the demagnetization map shown in Fig. 2.9.

An example of the ideal dq current vector trajectory required to vary MS at a constant torque of 15 Nm is shown in Fig. 4.1, considering the POC motor. The

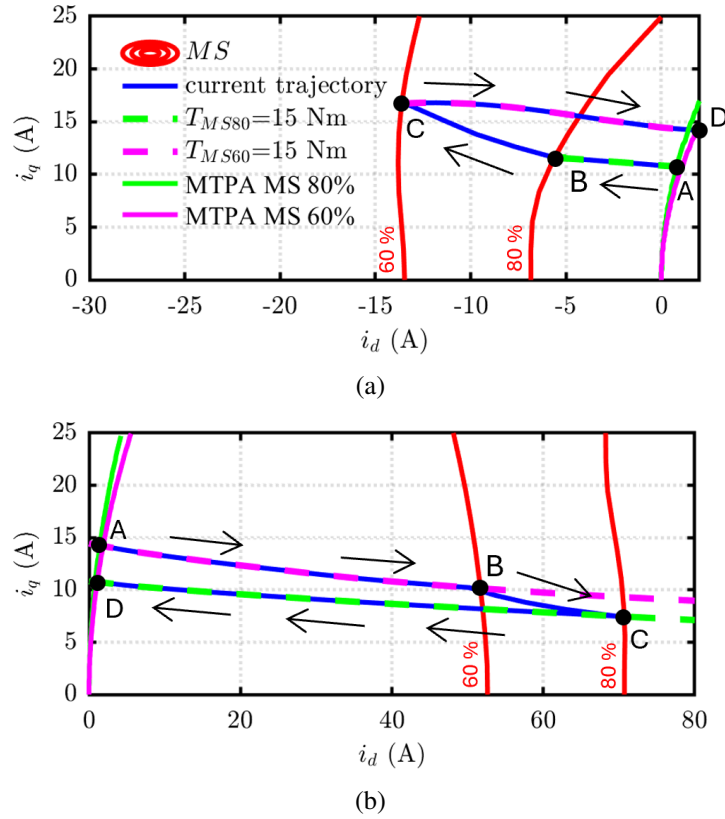


Fig. 4.1 Current vector trajectory for MS manipulation between 60% and 80% at 15 Nm: (a) demagnetization and (b) remagnetization.

red solid contours indicate the operating boundaries associated with $MS = 80\%$ and $MS = 60\%$, obtained from the magnetization maps introduced in Fig. 2.9. The dashed green and magenta curves represent the 15 Nm iso-torque lines corresponding to $MS = 80\%$ and $MS = 60\%$, respectively. The solid green and magenta curves denote the corresponding MTPA trajectories. The blue curve represents the current trajectory adopted to perform the MS transition while keeping the torque at 15 Nm.

The demagnetization process from $MS = 80\%$ to $MS = 60\%$, shown in Fig. 4.1.a, can be described as follows:

1. **Initial operating point at $MS = 80\%$ (point A):**
the process starts from point A, which is the MTPA operating point corresponding to $MS = 80\%$ at $T = 15$ Nm.
2. **Movement toward the demagnetization boundary (point A→B):**
the current vector moves from point A to point B along the 15 Nm iso-torque

line corresponding to $MS = 80\%$. Point B is the intersection between this iso-torque line and the $MS = 80\%$ contour in the Demag map. During this stage, MS remains fixed at $MS = 80\%$.

3. Demagnetization stage (point B→C):

from point B, the current vector enters the demagnetization region and moves toward point C. During this stage, MS decreases progressively from 80% to 60%. The trajectory passes through the sequence of intermediate points defined by the iso-torque lines at each intermediate MS . For graphical simplicity, only the initial point B and the final point C of this stage are shown.

4. Arrival at the $MS = 60\%$ boundary (point C):

at point C the machine reaches the desired $MS=60\%$ boundary in the Demag plane while still following the constant-torque trajectory.

5. Convergence to the final MTPA operating point (point C→D):

after the demagnetization is completed, the current vector moves from point C to point D along the 15 Nm iso-torque line corresponding to $MS=60\%$. Point D is the final MTPA operating point at $MS=60\%$. During this stage, MS remains fixed at 60%.

Symmetrically, the remagnetization i_{dq} trajectory at a constant torque of 15 Nm is shown in Fig. 4.1.b, describing the transition from $MS = 60\%$ to $MS = 80\%$. This remagnetization process can be described as follows:

1. Initial operating point at $MS = 60\%$ (point A):

the process starts from point A, which is the MTPA at $MS=60\%$, 15 Nm.

2. Movement toward the remagnetization boundary (point A→B):

the current vector moves along the 15 Nm iso-torque line of $MS = 60\%$. Point B is the intersection between this iso-torque line and the $MS = 60\%$ contour in the Remag map. During this stage, MS remains fixed at 60%.

3. Remagnetization stage (point B→C):

the i_{dq} vector enters the remagnetization region. MS increases progressively from 60% to 80%. The trajectory passes through the sequence of intermediate points defined by the iso-torque lines at each intermediate MS .

4. **Arrival at the $MS = 80\%$ boundary (point C):**

At point C the motor is remagnetized at $MS = 80\%$.

5. **Convergence to the final MTPA operating point (point C→D):**

i_{dq} moves to point D along the 15 Nm iso-torque line corresponding to $MS = 80\%$. Point D is the final MTPA operating point at $MS = 80\%$, 15 Nm.

As can be noted, the main control challenge is to maintain torque continuity during MS variation without explicitly tracking the full sequence of intermediate MS trajectories. Designing a control system capable of forcing the current vector i_{dq} to follow the ideal path is particularly difficult during the **demagnetization stage** and **remagnetization stage**. In these two stages, the current reference should ideally be determined from the sequence of intermediate MS s between the initial and final MS levels. This requires access to the corresponding operating boundaries and iso-torque curves for all intermediate MS values. Storing all these data in LUTs would lead to impractically high memory requirements, whereas computing them online would introduce a significant computational burden. Moreover, the flux and demag/remag maps are prone to measurement and model uncertainty. Any inaccuracy in the determination of these maps would reflect in significant torque undulation during the MS transient.

To address this issue, this chapter introduces two dedicated control strategies aimed at mitigating the torque disturbances during MS variation, without requiring explicit estimation of the current trajectory. The first control strategy is developed within a Direct Vector Flux Control (DFVC) framework, while the second control strategy is developed based on a modified Field Oriented Control (FOC) framework.

4.1 *MS* Regulation Under Direct Flux Vector Control

The DFVC strategy [96] exploits the $d_s q_s$ reference frame shown in Fig. 4.2, where the d_s axis is aligned with the stator flux vector.

In the $d_s q_s$ coordinates, the voltage equations are represented as:

$$\begin{cases} v_{ds} = R_s i_{ds} + \frac{d\lambda}{dt} \\ v_{qs} = R_s i_{qs} + \lambda \left(\omega + \frac{d\delta}{dt} \right) \end{cases} \quad (4.1)$$

calculated as:

$$i_{qs}^* = \frac{T^*}{\frac{3}{2}p\lambda^*} \quad (4.3)$$

This means that, in the DFVC, the torque reference T^* and the flux amplitude reference λ^* are sufficient to determine the current reference i_{qs}^* . The i_{qs}^* is saturated according to the inverter current limit i_{max} .

During constant *MS* operating conditions, the reference flux amplitude λ^* is set according to the MTPA law. The flux reference is then saturated by the Flux Weakening (FW) block according to the motor speed.

4.1.1 *MS* Manipulation Strategy

When the control switches to the *MS* manipulation mode, as highlighted in Fig. 4.3, the generation of the reference flux amplitude is no longer based on the MTPA law. Instead, it is generated from dedicated LUTs. In particular, the Demag and Remag maps $MS(i_d, i_q)$ reported in Fig. 2.9 are manipulated and inverted and written as $\lambda(MS, T)$. During the demagnetization process, the reference flux amplitude is selected based on the reference torque T^* and the target demagnetization state MS_{demag}^* :

$$\lambda^* = \lambda(MS_{demag}^*, T^*) \quad (4.4)$$

Similarly, during the remagnetization process, the reference flux amplitude is determined based on the reference torque and the target remagnetization state MS_{remag}^* :

$$\lambda^* = \lambda(MS_{remag}^*, T^*) \quad (4.5)$$

The demagnetization and remagnetization LUTs used for generating the flux reference during the *MS* variation process are shown in Fig. 4.4. These LUTs are derived from the magnetization maps introduced in Fig. 2.9. During their generation process, not only the required *dq* current vector to achieve a certain *MS* is recorded, but also the corresponding *dq* flux linkages. Then, by using the torque equation in the *dq* frame (2.6), the corresponding torque for each i_{dq} operating point is evaluated. The flux amplitude is calculated as:

$$\lambda = \sqrt{\lambda_d^2 + \lambda_q^2} \quad (4.6)$$

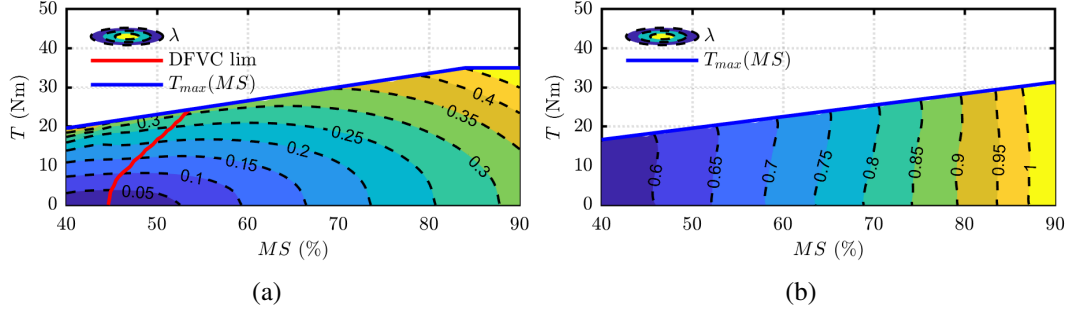


Fig. 4.4 (a) Inverse demagnetization and (b) inverse remagnetization maps defined in the torque and MS domain for the DFVC control strategy.

By manipulating these data, the demagnetization and remagnetization characteristics can be expressed in the torque and flux domain, thus defining the LUTs $\lambda(MS_{demag}^*, T^*)$ and $\lambda(MS_{remag}^*, T^*)$. In this way, the control system can directly generate the flux reference needed for the desired MS target at a given torque level, without explicitly estimating the full dq current trajectory through all intermediate MS levels.

Apart from the generation of the reference flux amplitude, the rest of the DFVC structure remains unchanged during MS regulation mode. The quadrature current reference i_{qs}^* is still computed according to eq. (4.3). Therefore, even in the case of inaccurate Demag/Remag LUTs or flux maps, smooth torque is maintained during the MS variation transient. Once the target MS^* is reached, the controller returns to the normal operating mode and the reference flux amplitude is again based on the MTPA law.

This control strategy permits to use the DFVC both during the normal drive operation and magnetization transients, with clear advantages in terms of smooth controllability. However, the demagnetization region accessible under DFVC is limited. In Fig. 4.4.a, the red line represents the lower feasibility boundary of the DFVC-based demagnetization process. The flux amplitude contours have a circular shape, and gradually become horizontal as MS decreases and T increases. This behavior occurs when the operating point approaches the MTPV region. In this condition, the sign of the feedback signal reverses, causing the control input to diverge and the control to become unstable.

This limitation does not affect the remagnetization process. As can be noticed in Fig. 4.4.b, the flux contours are vertical or pseudo-vertical, meaning that the sign of

the feedback signal does not change. Therefore, stable control is maintained during remagnetization.

4.1.2 Flux Observer

The flux observer is a key building block of DFVC-based motor control. The proposed control strategy for *MS* manipulation with reduced torque disturbance can be associated with any type of flux observer suitable for VFMs, i.e., including the *MS* variability in the flux estimation.

A modified flux observer with embedded *MS* estimation is proposed, according to the block diagram represented in Fig. 4.5. This observer is based on a classical hybrid flux observer, which combines the flux estimated through EMF integration (voltage model) with the flux obtained from the magnetic saturation characteristic, also called flux maps (current model). Compared to the classical hybrid observer, the present observer is extended in two ways:

- by implementing the 3D flux maps introduced in Section 2.2.3 to include *MS* variation;
- by adding an online estimator of the magnetization state.

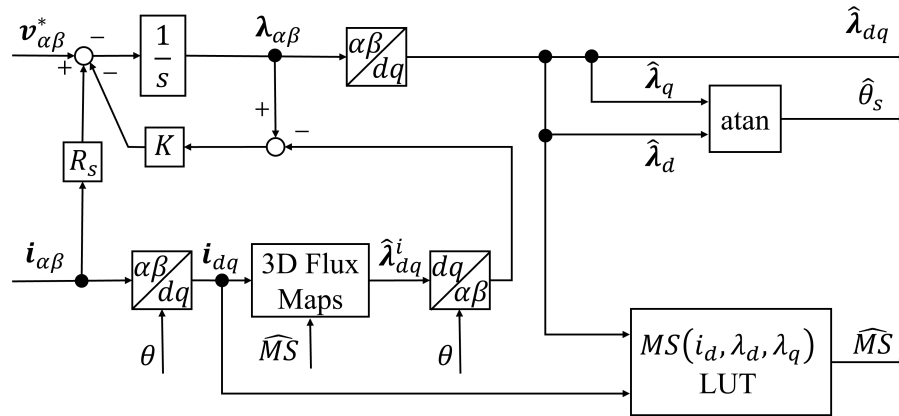


Fig. 4.5 Integrated flux observer with *MS* estimation.

The online *MS* estimator returns the estimated $\widehat{MS}$ based on the measured i_d and the observed dq flux linkages, using the dedicated $MS(i_d, \lambda_d, \lambda_q)$ LUT. Again, this

LUT is derived from the flux maps introduced in Section 2.2.3, since they contain the relationship of $\lambda_{dq} = \lambda_{dq}(i_d, i_q, MS)$.

Finally, the estimated flux angle $\hat{\theta}_s$ is calculated as:

$$\hat{\theta}_s = \tan^{-1} \frac{\lambda_q}{\lambda_d} \quad (4.7)$$

4.1.3 Performance Validation

The effectiveness of the developed DFVC control strategy for providing smooth torque transition during MS manipulation is evaluated in MATLAB/Simulink, using a 600 V DC-link inverter and the POC VFM introduced in Section 2.3.

The simulation results are represented in Fig. 4.6. The machine initially operates at $MS = 80\%$, accelerates to 1000 rpm, and is loaded with a constant reference torque of 12 Nm throughout the entire test. A demagnetization command to $MS = 50\%$ is then applied, followed by a remagnetization command back to $MS = 80\%$.

As can be observed, the controlled variables closely track their corresponding references throughout the entire test. In particular, MS accurately follows its target MS^* , while the residual torque perturbation only presents short-duration spikes within ± 2 Nm during both demagnetization and remagnetization transients.

The flux behavior shown in Fig. 4.6.c further clarifies the control action. The machine starts with a constant reference flux amplitude of 0.30 Vs. During the demagnetization transient, the flux reference is reduced to 0.16 Vs in order to drive the machine toward the target $MS = 50\%$. Once the demagnetization process is completed, the reference flux returns to 0.23 Vs, corresponding to the normal operating condition at the new magnetization state. Subsequently, during remagnetization, a flux pulse of 0.95 Vs is applied, after which the reference settles again to 0.31 Vs when the machine returns to $MS = 80\%$. As can be noted, i_{qs}^* is correspondingly updated based on (4.3), thus maintaining the torque control during the whole process, without requiring explicit prediction of the full current trajectory in the dq plane.

Fig. 4.7 shows the results of deep demagnetization test performed at the same speed of 1000 rpm, but with torque reference $T^* = 16$ Nm and target $MS^* = 40\%$. The results show that the control becomes unstable and is unable to reach the target MS^* . Indeed, this working point lies beyond the red limit boundary shown

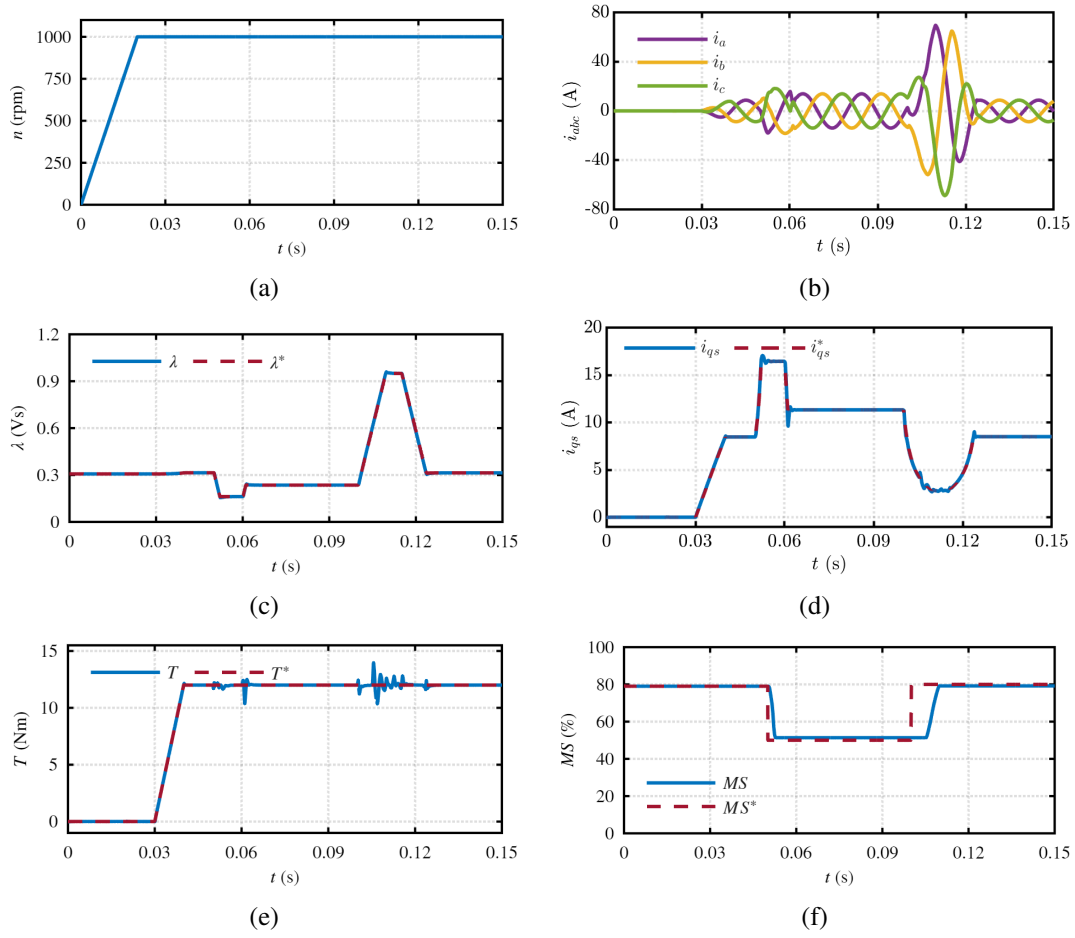


Fig. 4.6 Simulation test results using the DFVC control strategy: (a) rotor mechanical speed, (b) phase currents, (c) flux magnitude, (d) stator q_s -axis current, (e) electromagnetic torque and (f) magnetization state.

in Fig. 4.4.a. Therefore, the DFVC-based strategy is unsuitable for deep demagnetization. For this reason, another control strategy capable of achieving deep demagnetization is developed in the next section.

4.2 MS Regulation Under Modified Field Oriented Control

Another control strategy for achieving smooth torque transition during MS variation is developed within a modified FOC framework. This approach, referred to as the

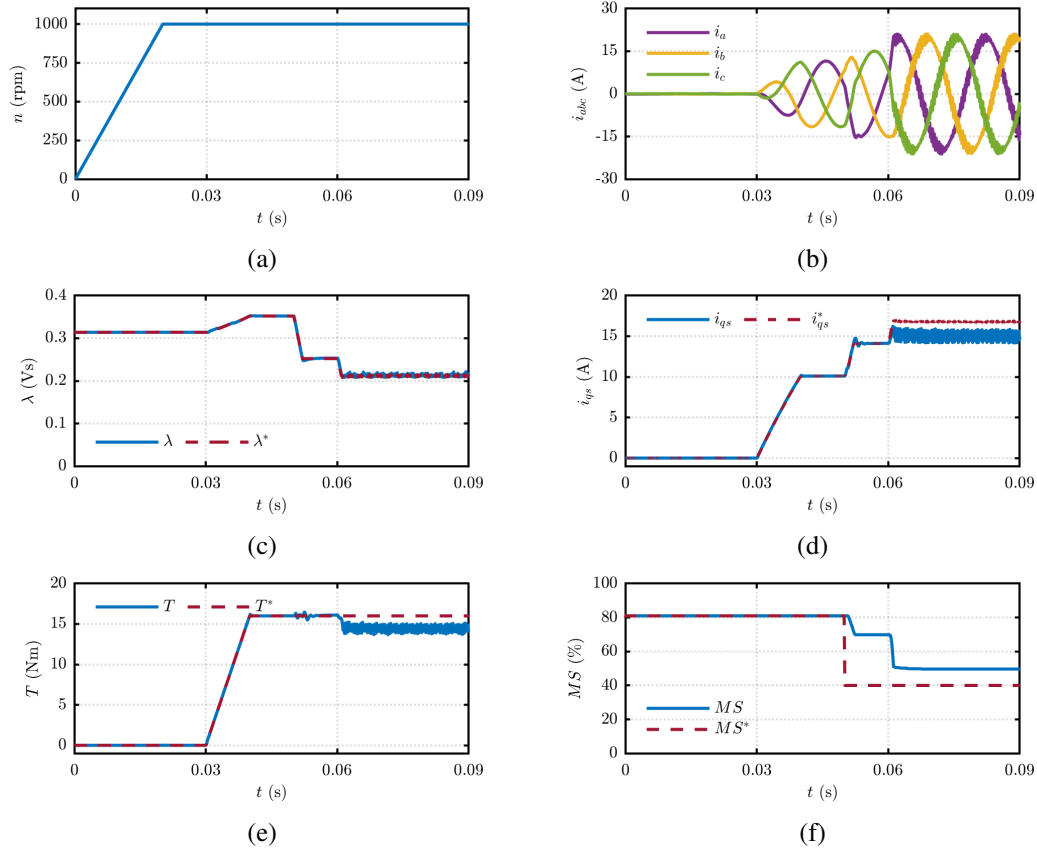
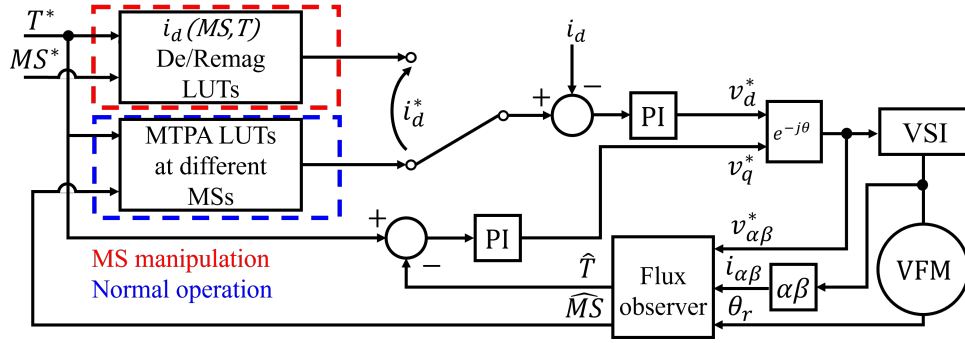


Fig. 4.7 Deep demagnetization test results using the DFVC control strategy: (a) rotor mechanical speed, (b) phase currents, (c) flux magnitude, (d) stator q_s -axis current, (e) electromagnetic torque and (f) magnetization state.

($i_d - T$) control strategy, uses the d -axis current and the electromagnetic torque as the two closed-loop controlled variables. The overall structure of this proposed control strategy is shown in Fig. 4.8.

In this control strategy, the d -axis current loop is used to drive the demagnetization and remagnetization process, while the torque loop ensures tracking of the torque reference during the MS transition. In this way, the MS variation can be performed while reducing the torque disturbances associated with demagnetization and remagnetization transients.

During normal operation, when the MS is constant, the d -axis current reference i_d^* is obtained from a LUT implementing the MTPA law as a function of the torque

Fig. 4.8 Proposed $i_d - T$ control scheme.

reference T^* and the estimated magnetization state $\widehat{MS}$:

$$i_d^* = i_d(T^*, \widehat{MS}) \quad (4.8)$$

When the controller switches to the MS manipulation mode, the same structure is preserved, but the MTPA LUT is replaced by dedicated demagnetization and remagnetization tables:

$$i_d^* = i_d(T^*, MS^*) \quad (4.9)$$

In this case, the reference i_d^* is generated as a function of the torque reference T^* and the desired target magnetization state MS^* , so that the required MS transition can be performed while maintaining smooth torque behavior.

Torque estimation is also required to close the torque loop and can be obtained using any suitable flux observer, according to:

$$\hat{T} = \frac{3}{2}p \left(\hat{\lambda}_d i_q - \hat{\lambda}_q i_d \right) \quad (4.10)$$

Meanwhile, the estimated $\widehat{MS}$ can be obtained using the same flux observer introduced in Section 4.1.2.

4.2.1 Torque Loop Calibration

The torque loop of the proposed control system is illustrated in Fig. 4.9, where the torque gain $K_T = T/i_q$ is derived from the conventional PMSM torque equation:

$$T = \frac{3}{2}p(\lambda_d i_q - \lambda_q i_d) = \frac{3}{2}[(L_d - L_q)i_d + \lambda_m]i_q \quad (4.11)$$

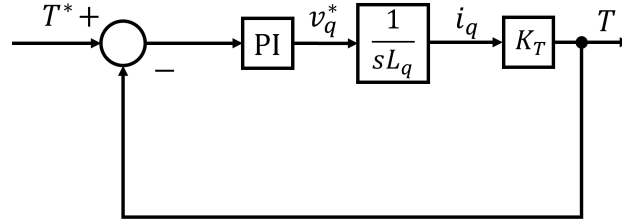


Fig. 4.9 Detailed torque loop.

Accordingly, the torque gain K_T can be expressed as:

$$K_T = \frac{3}{2}[(L_d - L_q)i_d + \lambda_m] \quad (4.12)$$

The torque gain K_T depends on the machine inductances L_d , L_q , the PM flux linkage λ_m , and the d -axis current i_d . In the proposed control implementation, L_d , L_q and λ_m are taken as their nominal values, while i_d is continuously updated from the measured current. Therefore, the torque gain K_T is updated online and used for the calibration of the PI regulator.

The PI controller of the torque loop can be calibrated according to:

$$\begin{cases} k_p = \omega_b L_q / K_T \\ k_i / k_p < \omega_b \end{cases} \quad (4.13)$$

where ω_b is the desired bandwidth of the torque loop.

This calibration allows the torque controller to maintain a consistent dynamic response despite the variation of the torque gain caused by changes in i_d . Therefore, the proposed formulation is suitable for the present control strategy, where the d -axis current is intentionally modified during the MS manipulation process.

4.2.2 MS Manipulation Strategy

When the controller operates in the MS manipulation mode shown in Fig. 4.8, the reference d -axis current i_d^* is generated from dedicated demagnetization and remagnetization tables. During demagnetization, the reference current i_d^* is selected based on the torque reference T^* and the target demagnetization state MS_{demag}^* :

$$i_d^* = i_d \left(MS_{demag}^*, T^* \right) \quad (4.14)$$

Similarly, during remagnetization, the reference d -axis current i_d^* is generated based on the torque reference and the target remagnetization state MS_{remag}^* :

$$i_d^* = i_d \left(MS_{remag}^*, T^* \right) \quad (4.15)$$

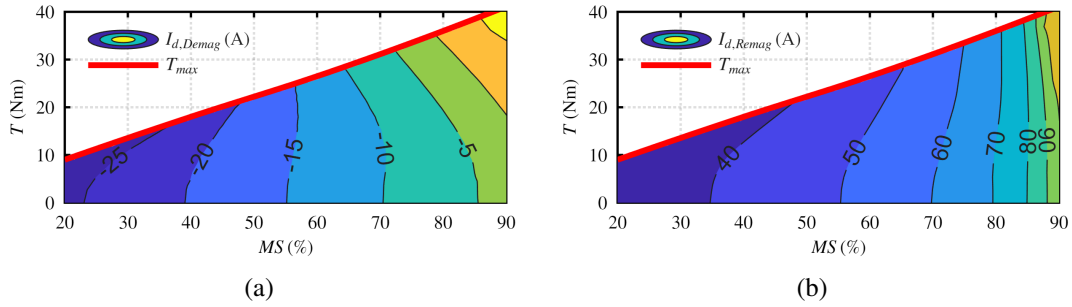


Fig. 4.10 (a) Inverse demagnetization and (b) inverse remagnetization maps defined in the torque and MS domain for the $i_d - T$ control strategy.

The demagnetization and remagnetization LUTs used for generating the reference d -axis current in MS manipulation mode are represented in Fig. 4.10. These tables are derived from the magnetization maps introduced in Fig. 2.9, where the dq current vector required to achieve any target MS is provided. For each i_{dq} operating point, the corresponding dq flux linkages are also recorded, so that the associated torque can be calculated through the PMSM torque equation.

By manipulating these data, the demagnetization and remagnetization characteristics are represented in the i_d and T domain, thus defining the LUTs $i_d^*(MS_{demag}^*, T^*)$ and $i_d^*(MS_{remag}^*, T^*)$. In this way, the control system is able to directly generate the reference d -axis current required to reach the desired target MS at a given torque level, without explicitly estimating the full dq current trajectory through all interme-

diate MS levels. Meanwhile, the torque continuity is guaranteed by the closed loop torque controller operating in q -axis. Again, in the case of inaccurate Demag/Remag LUTs a smooth torque transient is guaranteed.

Apart from the i_d^* generation, the control structure remains identical in MS manipulation and constant MS modes. Moreover, this control strategy allows the machine to reach any target MS^* . As can be noticed in Fig. 4.10 where the i_d contours are vertical or pseudo-vertical in the (T, MS) domain. This means the sign of i_d loop feedback will not change, providing stable control.

4.2.3 Performance Validation

The effectiveness of the developed $i_d - T$ control strategy for providing smooth torque transition during MS manipulation is evaluated in MATLAB/Simulink, using a 600 V DC-link inverter and the POC VFM introduced in Section 2.3.

The simulation results are illustrated in Fig. 4.11. The machine initially operates at $MS = 80\%$, accelerates to 1000 rpm, and is loaded with a constant reference torque of 16 Nm throughout the entire test. A demagnetization command to $MS = 40\%$ is then applied, followed by a remagnetization command back to $MS = 60\%$.

As can be observed, the controlled variables closely track their corresponding references throughout the entire test. In particular, the MS accurately follows its reference MS^* . Moreover, Fig. 4.11.c shows that, under the same simulation conditions as the deep demagnetization test performed with the DFVC-based strategy (see Fig. 4.7), the $i_d - T$ control strategy successfully demagnetizes the machine to the desired value of $MS^* = 40\%$.

In terms of the torque response, the error during the MS transients remains within ± 1.0 Nm.

The d -axis current behavior shown in Fig. 4.11.b clearly reflects the action of the proposed control strategy. During normal operation, i_d remains close to zero, as dictated by the MTPA operating condition at the initial MS . When the demagnetization command is applied, a negative i_d pulse is generated to reduce the MS from 80% to 40%. Later, a positive i_d pulse is injected to remagnetize the machine to MS of 60%. In both cases, the actual i_d closely follows its reference, confirming the effectiveness of the current regulation.

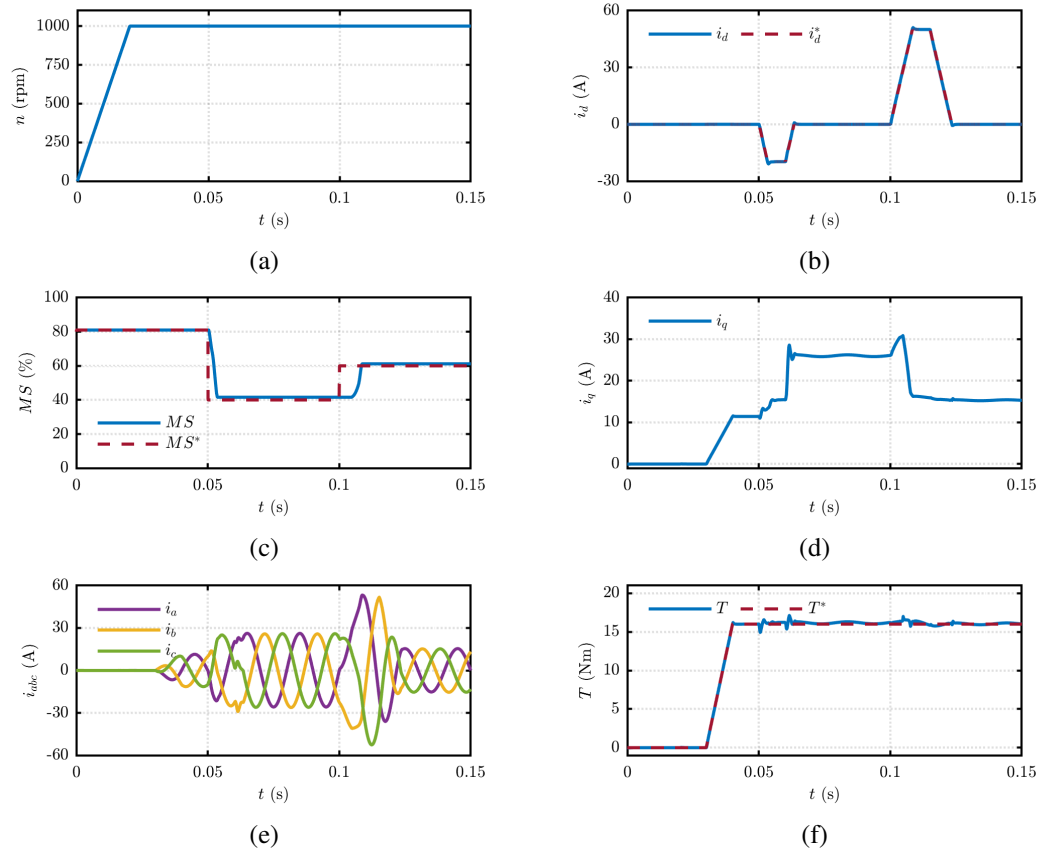


Fig. 4.11 Simulation test results using the $i_d - T$ control strategy: (a) rotor mechanical speed, (b) d -axis current, (c) magnetization state, (d) q -axis current, (e) phase currents and (f) electromagnetic torque.

The q -axis current shown in Fig. 4.11.d is automatically adjusted by the torque control loop in order to maintain the required torque during the magnetization transients. This is a key point of the proposed strategy: while the i_d is intentionally modified to manipulate the MS , i_q is continuously adapted so that the torque remains close to the desired value. As a result, the torque disturbance is significantly small despite the large i_d current pulse is applied.

These results confirm that the proposed $i_d - T$ control strategy is capable of achieving the desired MS transitions while maintaining a smooth torque response, and permitting the deep demagnetization. Therefore, the method provides an effective solution for reducing the unwanted torque pulsation associated with demagnetization and remagnetization transients in VFMs.

4.3 Comparison of the Proposed Strategies

This chapter introduced two dedicated control strategies aimed at reducing the torque disturbance during magnetization state variation in VFMs. Both strategies are developed to avoid the explicit estimation of the full current vector trajectory across the intermediate MS levels, which would otherwise require either large LUTs or significant online computational effort and would be highly sensitive to inaccurate LUTs determination. Instead, both methods generate dedicated references for MS variation while preserving smooth torque transition.

The first strategy is developed within the DFVC framework, with the stator flux amplitude λ and the quadrature current component i_{qs} being the control variables. The second strategy, referred to as $i_d - T$, is developed within a modified Field Oriented Control (FOC) framework strategy controlling the d -axis current and the electromagnetic torque estimate in closed-loop.

Although both control strategies have been shown to effectively reduce torque pulsation during demagnetization and remagnetization transients, they present different strengths and limitations.

A first important difference is related to the Field-Weakening (FW) operation. In the DFVC-based strategy, FW is naturally embedded in the control framework. Once the reference flux amplitude is saturated according to the voltage constraint, the dq current vector rotates automatically toward the appropriate operating point. Therefore, no additional FW LUTs or dedicated supervisory logic are required. This represents a practical advantage of the DFVC approach. In contrast, the $i_d - T$ strategy retains a more conventional FOC-based structure. FW is not automatically obtained from the control itself, and a dedicated LUT-based strategy is required, similar to classical FOC implementations. Therefore, from the viewpoint of FW integration, the DFVC-based approach is simpler and more direct.

A second major difference is related to the feasible achievable MS range, especially during demagnetization. In the DFVC-based strategy, the reference flux amplitude for MS manipulation is generated from inverse demagnetization and remagnetization LUTs defined in the (MS, T) domain, as reported in Fig. 4.4. However, the DFVC demagnetization capability is limited by the MTPV of the machine. When the VFM is required to operate beyond MTPV region during demagnetization transient, the controller will not work anymore. As a consequence, deep demagnetization

cannot be achieved within the DFVC framework. This limitation does not affect the remagnetization process.

The $i_d - T$ strategy does not suffer from this limitation. Since MS manipulation is directly imposed through the d -axis current. The full demagnetization region can be accessed, and deeper MS variation can be achieved. This makes the $i_d - T$ strategy more flexible when a wide MS operating range is required.

The main characteristics of the DFVC-based strategy can be summarized as follows:

- it provides automatic FW capability through direct saturation of the flux reference;
- it allows direct generation of the torque-producing current reference from the torque and flux references;
- it is naturally suitable for smooth remagnetization control;
- it is more suitable when deep demagnetization is not required.

And the main characteristics of the $i_d - T$ strategy can be summarized as follows:

- it directly regulates the d -axis current and the electromagnetic torque in closed-loop;
- it provides higher flexibility for deep MS manipulation, especially during demagnetization;
- it requires a dedicated FW strategy.

Table 4.1 [Comparison between the proposed DFVC and \$i_d - T\$ control strategies](#)

Aspect	DFVC	$i_d - T$
Category	Flux control	FOC
Controlled variables	λ, i_{qs}	i_d, T
MS variation reference	$\lambda^*(MS, T)$ LUT	$i_d^*(MS, T)$ LUT
Main strength	Automatic FW	Deep demagnetization achievable
Main limitation	Limited demagnetization range	Dedicated FW required
Recommended use	Simple FW integration required	Wide MS range access required

For clarity, the main differences between the two proposed control strategies are summarized in Tab. 4.1.

Overall, the choice between the two strategies depends on the target application and on the design priority. If simple integration of FW is the main objective and deep demagnetization is not required, the DFVC strategy is particularly attractive. Conversely, if the application requires wider accessibility of the demagnetization region over the MS range, the $i_d - T$ strategy becomes preferable.

Chapter 5

Remagnetization Speed Limit under Constrained Voltage

Remagnetization under load is one of the most critical operating conditions in VFMs. Unlike steady-state operation at a fixed MS , the remagnetization process requires the injection of a high-amplitude current pulse in order to increase the PM flux linkage level. While MS variation can be performed relatively easily at non-load conditions, the same operation becomes significantly more challenging when the machine is operating at non-zero speed. In particular, the large current required for remagnetization produces a substantial increase in stator flux linkage during the transient. As a consequence, the inverter voltage limit can be rapidly reached, which may prevent the remagnetization process from being completed successfully.

This limitation has an important practical consequence: the feasible operating region of the VFM during remagnetization can be much smaller than the steady-state torque-speed envelopes corresponding to different MS levels. Therefore, the conventional torque-speed characteristics alone are not sufficient to assess the actual operating capability of the VFM when online MS regulation is required. Instead, the remagnetization voltage constraint must also be taken into account, since it can substantially restrict the portion of the torque-speed plane from which remagnetization is feasible.

This concept is illustrated in Fig. 5.1, where the torque-speed profiles of the POC motor are reported for MS of 80% and 40%, considering a 600 V DC link. The blue and green curves represent the steady-state torque-speed limits at these two MS levels,

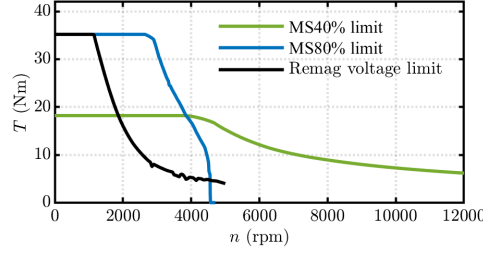


Fig. 5.1 Torque-speed profiles of the POC motor at $MS = 40\%$ and $MS = 80\%$. The black curve indicates the remagnetization feasibility boundary under inverter voltage constraint.

whereas the black curve indicates the remagnetization feasibility boundary under inverter voltage constraint. More specifically, only the operating points located on the left side of the black curve allow remagnetization from $MS = 40\%$ to $MS = 80\%$ to be performed successfully. It can be observed that this constraint significantly reduces the actually accessible operating range of the machine, highlighting the need for a dedicated analysis of the remagnetization voltage limit.

This chapter introduces a computationally efficient method for determining the feasible torque-speed operating range under inverter voltage constraints, including remagnetization transients. The proposed method relies on the manipulation of pre-calculated flux maps and magnetization maps introduced in chapter 2, enabling a rapid evaluation of the remagnetization voltage limit and providing a practical tool for VFM development.

5.1 Remagnetization Voltage Limit

5.1.1 $MS - \lambda_r$ Remagnetization Characteristic

The derivation of the feasible operating range considering remagnetization voltage limit starts from the remagnetization characteristics introduced in Section 2.2.2, which describe the current vector $\mathbf{i}_{dq}$ required to remagnetize the motor to a desired MS . These maps can be obtained either from FEA simulations or from experimental measurements. In addition, each operating point of the remagnetization map is associated with a corresponding dq flux linkage and therefore with a corresponding electromagnetic torque.

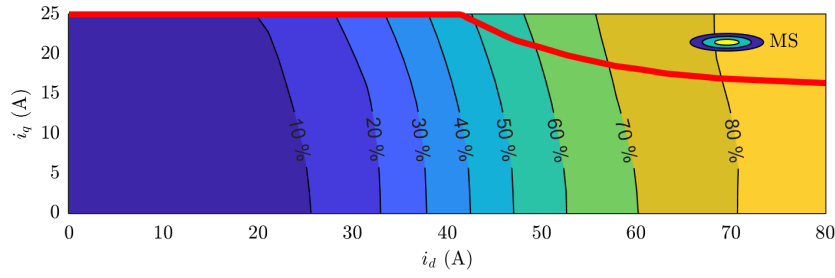


Fig. 5.2 Remagnetization characteristics of the POC VFM. The red curve represents the current trajectory for obtaining maximum achievable torque at the lowest MS .

For the purpose of evaluating the remagnetization operating range, a specific trajectory is selected within the remagnetization map. This trajectory is highlighted by the red curve in Fig. 5.2. It represents the i_{dq} trajectory to remagnetize the machine while maintaining the maximum torque at the lowest considered MS . In the present case, for the POC motor, the minimum operating MS is 40%, and the corresponding maximum torque is 19 Nm (see Fig. 5.1). Therefore, the red trajectory identifies all the remagnetization current vectors capable of increasing the MS while preserving a constant torque of 19 Nm.

For each point along this trajectory, the corresponding stator flux linkage during the remagnetization pulse, denoted as λ_r , can be evaluated from the flux maps. The subscript r indicates that the quantity refers to the remagnetization transient. By associating each remagnetization operating point with the corresponding target MS , the $MS - \lambda_r$ characteristic is obtained, as illustrated in Fig. 5.3.

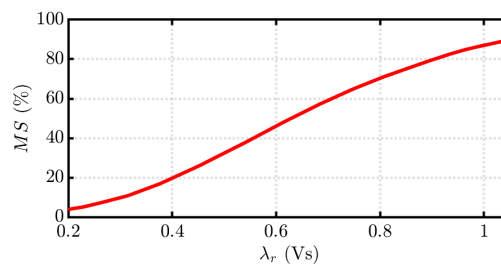


Fig. 5.3 λ_r vs MS remagnetization characteristic.

The $MS - \lambda_r$ remagnetization characteristic describes the flux linkage amplitude that must be generated during the remagnetization current pulse while maintaining the maximum torque corresponding to $MS = 40\%$. This flux linkage amplitude is higher than the nominal flux of the machine (0.374 Vs), thus approaching the inverter voltage limit.

5.1.2 Remagnetization Speed Calculation

The $MS - \lambda_r$ remagnetization characteristic introduced in Fig. 5.3 can be used to determine the maximum speed at which the remagnetization current pulse can be applied without exceeding the inverter voltage limitation. The proposed procedure is illustrated in Fig. 5.4. When the machine operates close to the voltage limit, the voltage equation can be approximated as:

$$\mathbf{v}_{dq} \approx \omega \mathbf{J} \boldsymbol{\lambda}_{dq} \rightarrow |\mathbf{v}_{dq}| \approx \omega \lambda \quad (5.1)$$

where λ is the flux linkage amplitude. The inverter voltage constraint is then imposed as:

$$|\mathbf{v}_{dq}| \leq V_{max} \quad (5.2)$$

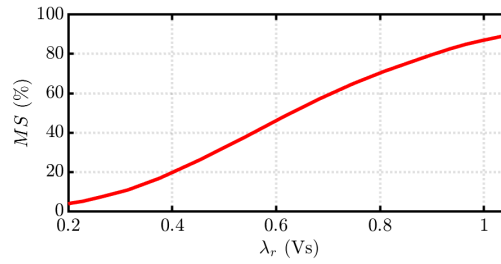


Fig. 5.4 Block diagram showing the procedure for calculating the remagnetization voltage limit curve.

It should be noted that the simplified relation in eq. (5.1) neglects the resistive voltage drop and the transient flux derivative $d\boldsymbol{\lambda}_{dq}/dt$. In addition, the ideal inverter voltage limit does not account for inverter nonlinearities. In practice, the actual voltage required during the remagnetization transient is expected to be slightly higher, while the effective voltage available for control is slightly lower. For this reason, in the simulation validation the voltage limit is reduced to $0.9 \cdot V_{dc} / \sqrt{3}$. As a consequence, the actual remagnetization speed limit is expected to be slightly lower than the one computed by the simplified calculation. However, this approximation is considered suitable for the present analysis, since it provides a computationally efficient estimation of the remagnetization limitation.

Considering standard PWM modulation with third-harmonic injection in the linear modulation region, the maximum available stator voltage is:

$$V_{max} = v_{dc}/\sqrt{3} \quad (5.3)$$

where v_{dc} is the DC-link voltage.

For a given rotor speed n , the corresponding electrical pulsation is calculated as:

$$\omega = n \cdot p \cdot \frac{\pi}{30} \quad (5.4)$$

Based on (5.1), the maximum flux linkage amplitude λ_{Vmax} that can be sustained at that speed without exceeding the voltage limit is calculated as:

$$\lambda_{Vmax} = \frac{V_{max}}{\omega} \quad (5.5)$$

This flux amplitude is then used as input to the $MS-\lambda_r$ remagnetization characteristic in order to determine the maximum magnetization state MS_{Vmax} that can be reached at the given speed while respecting the inverter voltage limit. Once MS_{Vmax} is determined, the corresponding torque is calculated by interpolating the torque–speed characteristics of the motor at different MS levels. This provides the maximum achievable torque T_r that can be maintained during the remagnetization transient under the voltage constraint V_{max} .

Repeating this procedure over different test speeds, a trajectory is obtained in the torque-speed profile, as shown in Fig. 5.1. This trajectory defines the boundary operating condition for remagnetization under load considering the inverter voltage limit.

5.1.3 Hysteresis Behavior in the Torque Vs Speed Profile

The analysis in Section 5.1.2 shows the remagnetization can only be performed if the machine speed is below the threshold imposed by the remagnetization voltage limit at the corresponding torque level, which lies to the left of the black curve shown in Fig. 5.1. This constraint introduces a speed-dependent hysteresis in the

MS variation during operation, which further modifies the actually accessible torque-speed operating region of the VFM.

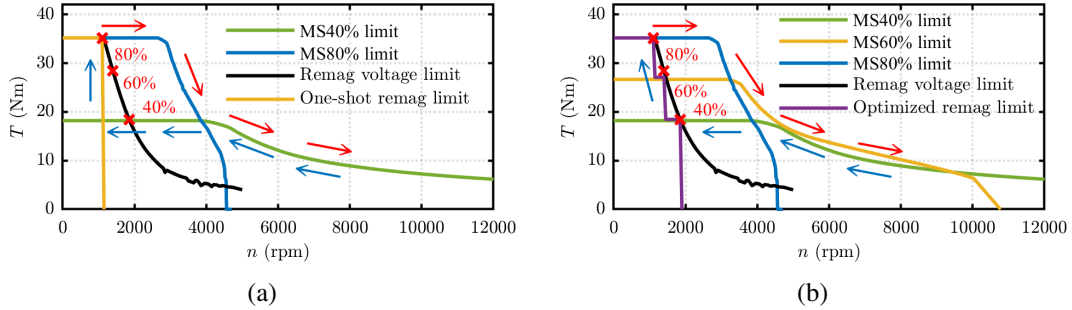


Fig. 5.5 VFM torque-speed profiles when adopting a) two MS levels, 40% and 80% (one-step remagnetization) and (b) three MS levels, 40%, 60% and 80% (two-steps remagnetization).

This effect can further reduce the practically feasible remagnetization operating region. Fig. 5.5.a illustrates the case in which the vehicle operates with only two MS levels, namely 40% and 80%. For visual clarity, the figure reports the absolute value of the torque. The red arrows indicate acceleration trajectory, whereas the blue arrows indicate deceleration scene. In this example, the vehicle starts to accelerate at $MS = 80\%$, then it is demagnetized to $MS = 40\%$ for FW operation above 4000 rpm, and subsequently decelerates. However, remagnetization back to $MS = 80\%$ during braking is infeasible for speeds higher than 1000 rpm because of the voltage constraint. Therefore, in order to recover the high MS condition, the vehicle must first decelerate below this threshold and then apply the remagnetization current pulse. As a consequence, the operating region from which remagnetization to $MS = 80\%$ is feasible is restricted to the left of the yellow vertical line.

The effect of this limitation can be mitigated by using finer granularity of MS variation, i.e., by using a multi-step remagnetization strategy. This case is illustrated in Fig. 5.5.b, where the vehicle operates at three MS levels: 40%, 60% and 80%. The vehicle again starts to accelerate with $MS = 80\%$, then is demagnetized to $MS = 60\%$ above 3200 rpm, and further demagnetized to $MS = 40\%$ above 4500 rpm. During deceleration, the machine is first remagnetized from $MS = 40\%$ to $MS = 60\%$, which is feasible below 1450 rpm, and afterwards remagnetized from $MS = 60\%$ to $MS = 80\%$ below 1100 rpm. As can be observed, in this two-step remagnetization scenario the speed threshold for remagnetization is higher than in the one-step case, therefore unblocking higher torque limit earlier using $MS = 60\%$. Consequently, the feasible remagnetization operating region is extended up to the purple line.

If the remagnetization is performed with finer granularity, the feasible remagnetization operating region progressively approaches the theoretical remagnetization voltage limit curve shown in black. Therefore, the hysteresis introduced by the remagnetization voltage constraint plays an important role in the actual torque-speed capability of the VFM. However, the MS granularity cannot be increased indefinitely, since each remagnetization step requires the injection of a high-amplitude current pulse. This leads to increased control complexity, additional energy consumption, and higher thermal stress. As a result, a trade-off must be found between the granularity of the MS variation and the severity of the speed limitations imposed by the voltage constraint.

It should be noted that, for the VFM, the remagnetization voltage limit occurs at relatively low speed because of the high current amplitude required for remagnetization. Demagnetization does not introduce a comparable limitation, since it requires much lower current and the associated flux linkage during the MS manipulation transient remains below the nominal value, thus not incurring in voltage limitation.

5.2 Simulation Results

5.2.1 Verification of Remagnetization Voltage Limit

To verify the remagnetization voltage limit described in Section 5.1, dedicated simulation of the VFM drive are carried out in MATLAB/Simulink, using the POC VFM model introduced in Section 2.3, the $i_d - T$ control strategy introduced in Section 4.2 and a 600 V DC-link inverter. In these tests, the motor is operated under torque control, while the rotor speed is externally imposed in order to evaluate the feasibility of the remagnetization transient at specific operating points in the torque-speed plane.

The objective of these simulations is to assess whether the remagnetization process can be successfully completed at operating points located close to the theoretical remagnetization voltage limit curves shown in Fig. 5.5.

Fig. 5.6 reports the simulation results of a one-step remagnetization transient from $MS = 40\%$ to $MS = 80\%$. The test is carried out at a constant rotor speed of 1000 rpm and with a reference torque of 16 Nm. This operating condition is

intentionally selected close to the remagnetization speed threshold for reaching $MS = 80\%$, indicated by the yellow line of Fig. 5.5.a, while the torque reference is chosen close to the maximum torque achievable at $MS = 40\%$.

The voltage limit is imposed as $V_{max} = 0.9 \cdot V_{dc} / \sqrt{3}$, in order to account for inverter nonlinearities, resistive voltage drops, and to preserve a suitable control margin. For the considered DC-link voltage, this corresponds to 310 V.

As shown in Fig. 5.6.d, the phase voltage waveforms indicate that, during the remagnetization transient, the voltage amplitude increases significantly and approaches the voltage limit V_{max} , represented by the dashed line. At the same time, the MS is successfully increased to $MS = 80\%$, as shown in Fig. 5.6.e, while the torque remains regulated close to its reference value of 16 Nm, as shown in Fig. 5.6.b. This test confirms the validity of the remagnetization speed limit derived in Section 5.1, indicating that voltage saturation would occur if the same remagnetization transient were attempted at higher speed.

Fig. 5.7 reports the simulation results of a two-step remagnetization process. In the first step, the machine is remagnetized from $MS = 40\%$ to $MS = 60\%$ at a constant rotor speed of 1300 rpm and with a reference torque of 16 Nm. In the second step, the machine is further remagnetized from $MS = 60\%$ to $MS = 80\%$ at 1000 rpm and with a reference torque of 25 Nm. Both operating points are intentionally selected close to the theoretical remagnetization voltage limit curve shown in Fig. 5.5.b, in order to verify the accuracy of the proposed calculation method also in the case of two-step remagnetization.

The phase voltage waveforms shown in Fig. 5.7.d confirm that both remagnetization steps are performed close to the voltage boundary. In both cases, the voltage amplitude rises significantly during the remagnetization pulse and reaches a peak value of about 290 V, which remains slightly below the voltage limit of 310 V. Therefore, both remagnetization transients are feasible, but only with a limited voltage margin. This again confirms the accuracy of the remagnetization speed limit derived in Section 5.1.

Moreover, the adoption of a two-step remagnetization strategy not only increases the speed threshold at which remagnetization becomes feasible, but also improves the available torque capability, the intermediate state $MS = 60\%$ provides a higher torque limit than $MS = 40\%$, thereby enlarging the practically accessible operating region before full remagnetization to $MS = 80\%$ is completed.

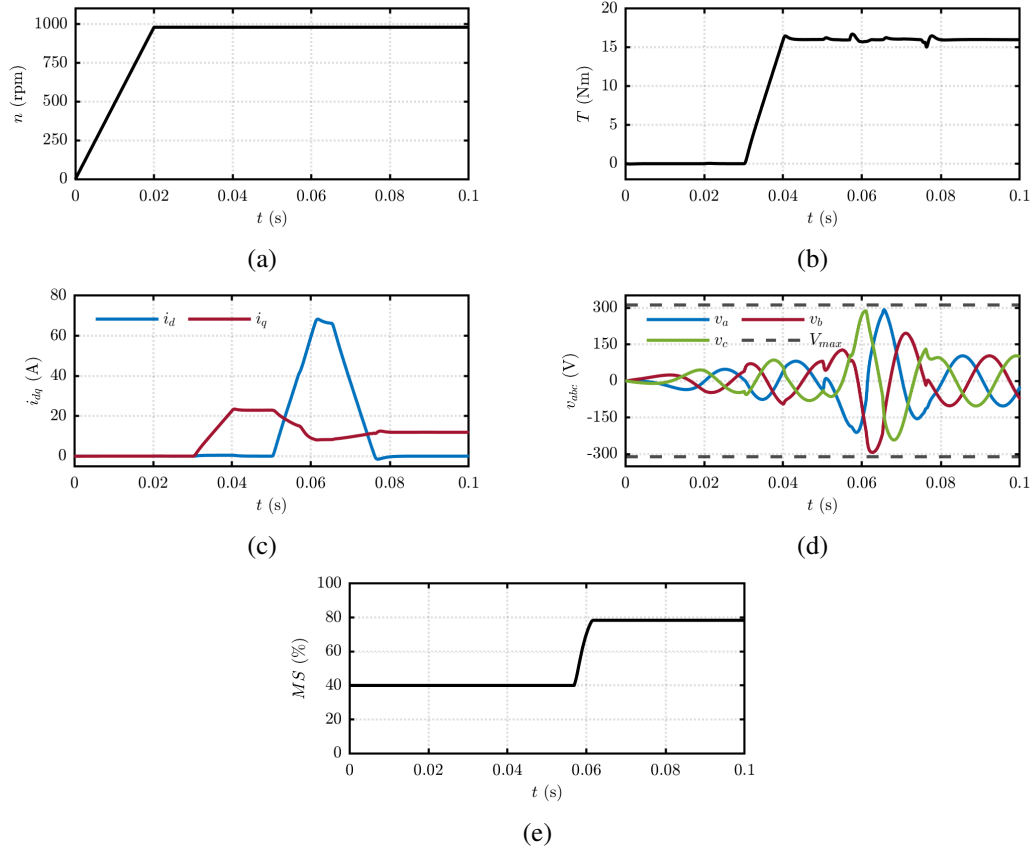


Fig. 5.6 Simulation results: remagnetization transient MS 40% $\rightarrow$ 80% at 1000 rpm, 16 Nm. (a) rotor speed, (b) electromagnetic torque, (c) stator dq -axis currents, (d) three phase voltage with its voltage limit and (e) MS .

5.2.2 Simple Acceleration and Deceleration Driving Cycle Test

The developed method is further adopted to verify the capability of the VFM drive to complete a driving cycle when the remagnetization voltage limit is taken into account. For a given speed profile, the corresponding torque demand is computed based on a light electric vehicle model. This choice is consistent with the POC VFM introduced in Section 2.3, which is designed as a down-scaled traction motor for preliminary validation purposes. The main vehicle parameters used in this analysis are shown in Tab. 5.1.

A custom driving cycle is considered, composed of two consecutive acceleration and deceleration routines:

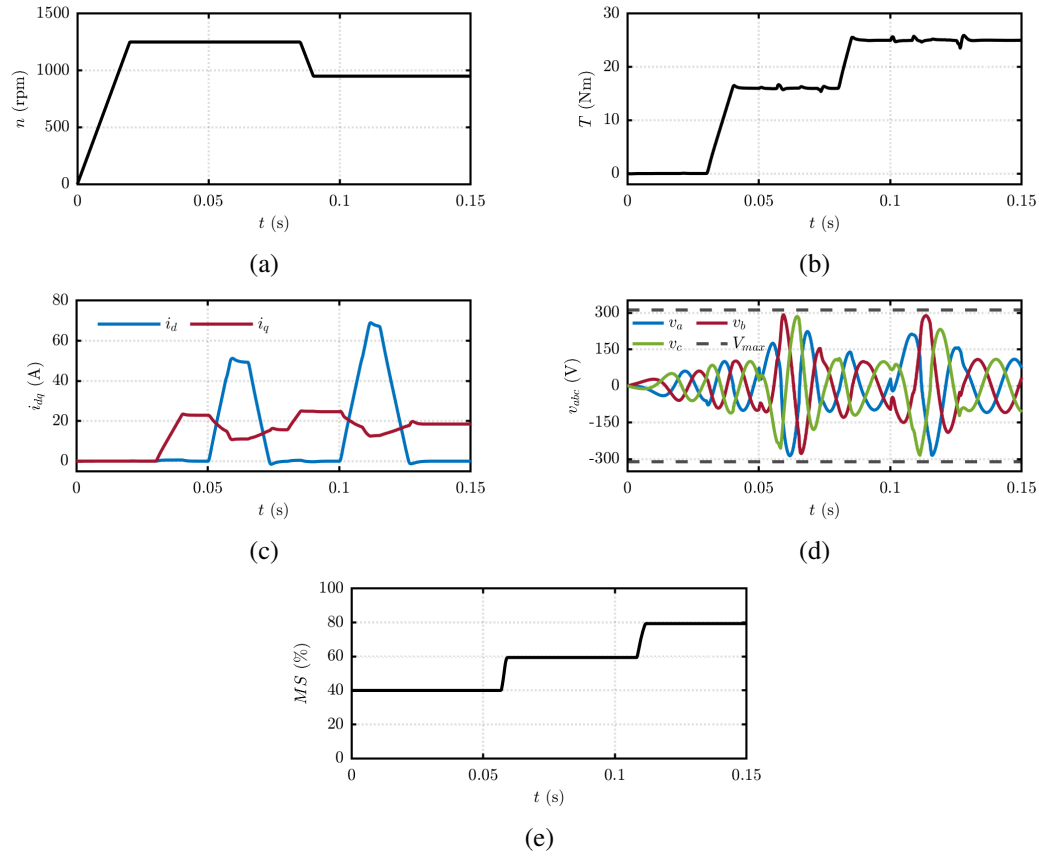


Fig. 5.7 Simulation results: two remagnetization transients $MS\ 40\% \rightarrow 60\%$ at 1300 rpm, 16 Nm and $MS\ 60\% \rightarrow 80\%$ at 1100 rpm, 25 Nm. (a) speed, (b) torque, (c) stator currents, (d) phase voltage with its voltage limit and (e) MS .

- in the first routine, the motor accelerates from 0 to 10000 rpm and then decelerates to 1150 rpm.
- in the second routine, starting from 1150 rpm, the motor accelerates again up to 4300 rpm and finally decelerates to standstill.

Two different MS regulation strategies are considered in order to assess the effect of the remagnetization voltage limit on the feasibility of the driving cycle. In the first case, the vehicle operates with only two MS levels of 40% and 80%. In the second case, an intermediate MS is also introduced, and the vehicle operates with $MS = 40\%$ $MS = 60\%$ and $MS = 80\%$.

The results of the first test are shown in Fig. 5.8. The vehicle initially accelerates with $MS = 80\%$, and is then demagnetized to $MS = 40\%$ once the speed exceeds

Table 5.1 Main parameters of the light vehicle

Vehicle total mass	1200 kg
Wheel diameter	0.406 m
Rolling resistance coefficient	0.02
Drag coefficient	0.5
Gear efficiency	0.97
Gear ratio	12

4000 rpm, since the required torque can no longer be sustained within the torque-speed limit corresponding to $MS = 80\%$. The vehicle then completes the remaining part of the first acceleration-deceleration routine at $MS = 40\%$

During the second acceleration and deceleration routine, a critical condition occurs at about $t = 150$ s, when a torque demand of 25 Nm is required at a speed above 1150 rpm. This operating point lies outside the feasible torque-speed region associated with $MS = 40\%$. In principle, the same point would become feasible at $MS = 80\%$, since the higher MS provides a larger torque capability. However, the remagnetization from $MS = 40\%$ to $MS = 80\%$ cannot be performed at that speed because of the inverter voltage constraint. As a consequence, the required torque cannot be delivered, even though it would be compatible with the steady-state torque-speed limit of the higher MS .

The infeasible operating points are indicated by the black points in Fig. 5.8. Therefore, the vehicle cannot complete the driving cycle when only $MS = 80\%$ and $MS = 40\%$ are used. This result clearly illustrates the practical impact of the remagnetization voltage limit on the feasible operating region of the VFM drive.

The results of the second test are shown in Fig. 5.9. In this case, the vehicle starts the first acceleration with $MS = 80\%$, and is then demagnetized to $MS = 60\%$ once the speed exceeds 4000 rpm. A further demagnetization to $MS = 40\%$ is applied above 9750 rpm, where the lower MS becomes advantageous for extending the speed range.

During the subsequent deceleration, the speed decreases to 1150 rpm. This speed is low enough to allow the motor to be remagnetized to $MS = 60\%$ under the inverter voltage constraint. Therefore, at the end of the first routine, the machine does not remain at $MS = 40\%$, but recovers the intermediate magnetization state $MS = 60\%$.

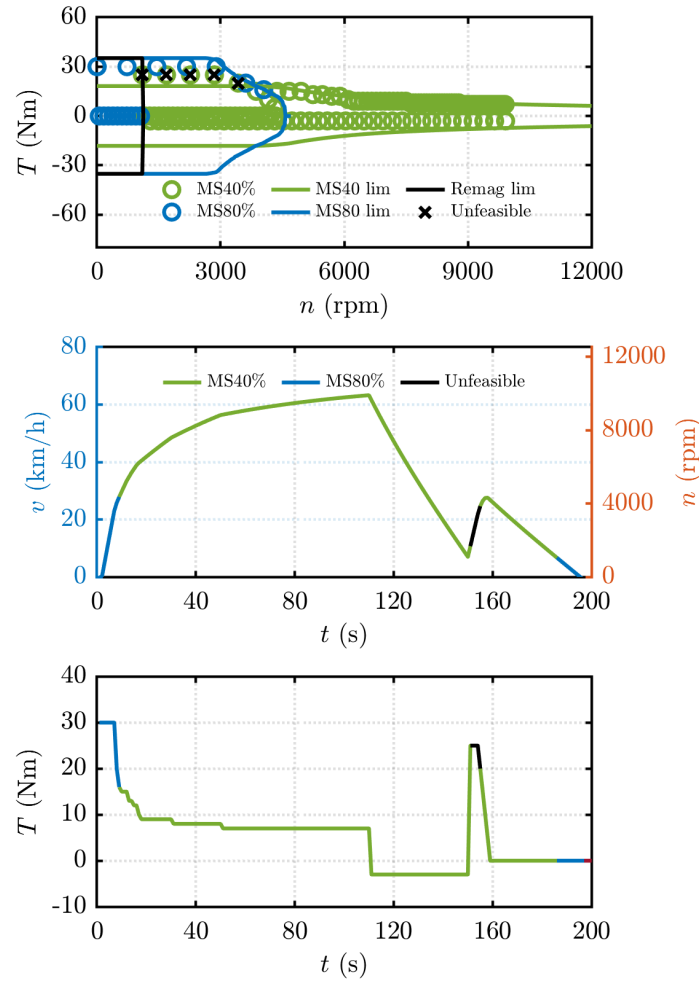


Fig. 5.8 Simple driving cycle test with feasible and unfeasible operating points are colored, with MS switching between 40% and 80%.

The second acceleration and deceleration routine is completed without encountering infeasible operating points, since the torque-speed capability associated with $MS = 60\%$ is sufficient to cover all the required operating points of the second routine.

Compared to the first case where only $MS = 80\%$ and $MS = 40\%$ are used, the introduction of the intermediate $MS = 60\%$ mitigates the constraint imposed by the remagnetization voltage limit and improves the capability of the vehicle to follow the required driving cycle. In other words, the use of finer MS granularity not only increases the speed threshold at which remagnetization becomes feasible, but also increases the available torque capability.

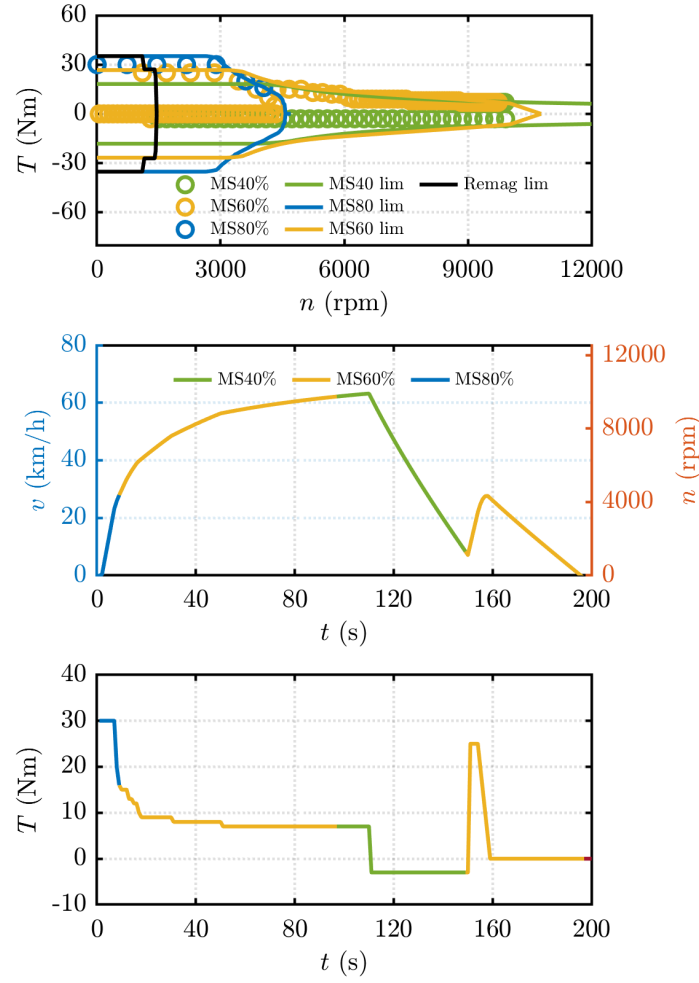


Fig. 5.9 Simple driving cycle test with MS switching between 40%, 60% and 80%.

This result confirms that a finer granularity in MS variation can effectively reduce the hysteresis-related limitation introduced by remagnetization under voltage constraint.

5.2.3 WLTC Class 3 Test

The proposed method is further applied to a more realistic driving mission, considering the low and medium speed regions of a WLTC Class 3 driving cycle. The same vehicle model adopted in Section 5.2.2 is used here, together with the same two MS regulation strategies.

The results of the first test using only $MS = 80\%$ and $MS = 40\%$ are shown in Fig. 5.10. As can be observed, the POC VFM is unable to complete the considered driving cycle. In particular, a large number of operating points fall outside the feasible torque-speed envelope and are therefore marked in red. In addition, several operating points are indicated by black points, meaning that they would appear feasible when considering only the steady-state torque-speed characteristics, but they are not actually reachable because the remagnetization process cannot be completed under the inverter voltage constraint. This result confirms again that, when only $MS = 40\%$ and $MS = 80\%$ are used, the remagnetization voltage limit can significantly restrict the actual operating capability of the VFM.

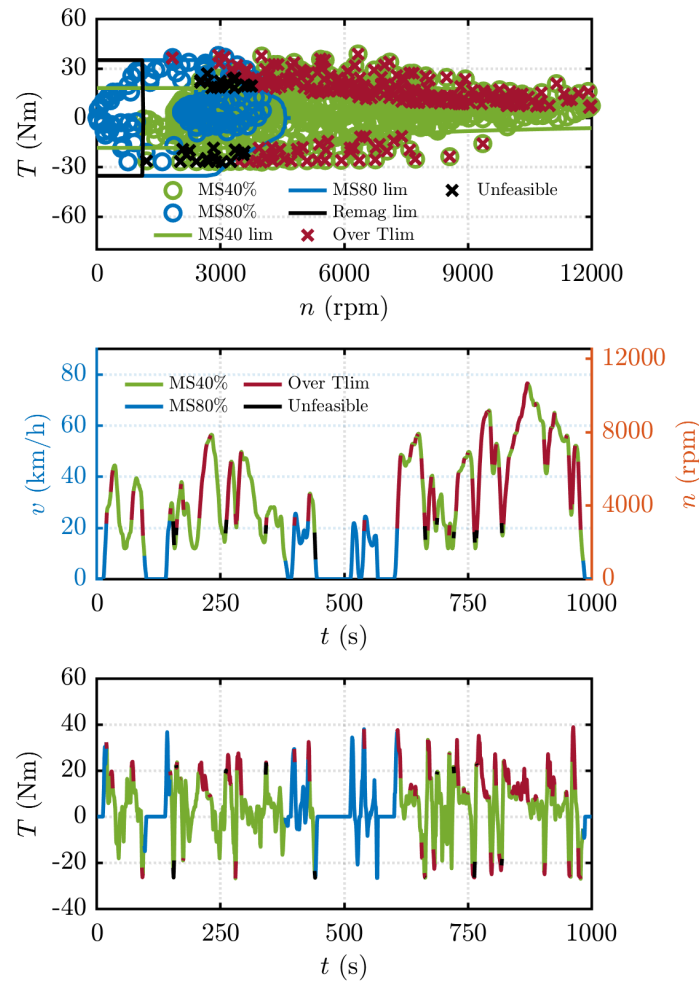


Fig. 5.10 WLTC driving cycle test with feasible and unfeasible operating points colored, MS switches between 40% and 80%.

The results of the second test performed using $MS = 80\%$, $MS = 60\%$ and $MS = 40\%$ are shown in Fig. 5.11. Although the driving cycle still cannot be fully completed because of the limited torque capability of the POC motor, the number of unfeasible operating points caused by remagnetization failure is significantly reduced compared with the first test. This improvement is due to the use of finer MS granularity, which mitigates the constraints imposed by the remagnetization voltage limit.

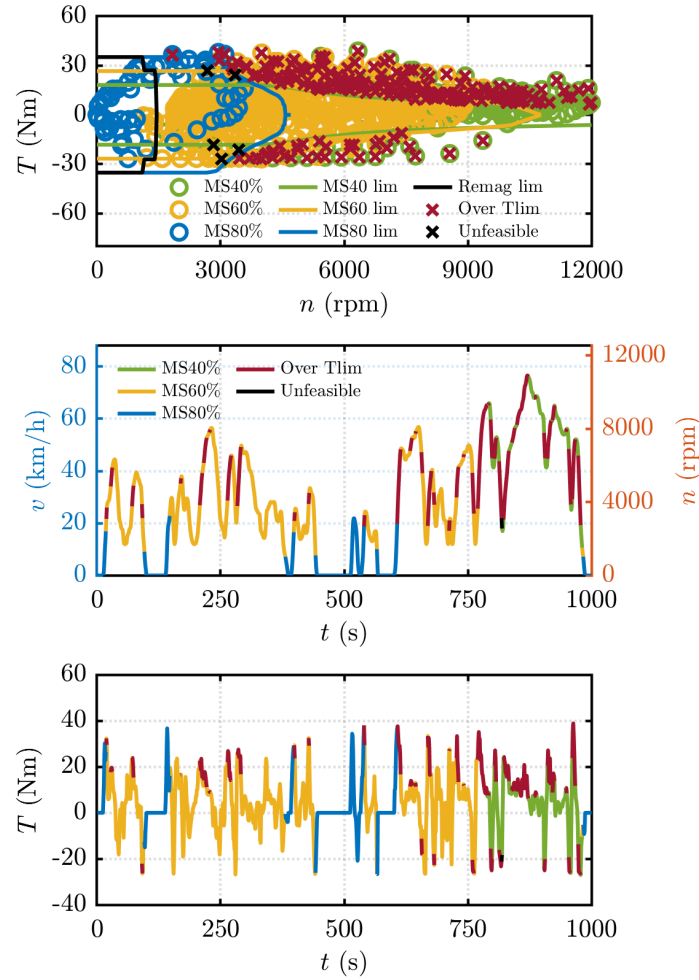


Fig. 5.11 WLTC driving cycle test with feasible and unfeasible operating points colored, MS switches between 40%, 60% and 80%.

These two tests illustrate that the proposed method for calculating the remagnetization voltage limit can be applied to any driving cycle conditions. Furthermore, the proposed method can be applied to any vehicle equipped with any VFM, as far as the corresponding flux maps, magnetization maps and vehicle model are available.

As an example, the WLTC driving cycle test previously presented for the traction motor in a heavier passenger car, introduced in Section 3.3.3, already accounts for this voltage limitation, as indicated by the black line shown in Fig. 3.24.

5.3 Chapter Conclusion

This chapter addressed the problem of remagnetization under inverter voltage constraint, showing that the actual operating capability of a VFM cannot be evaluated only from the steady-state torque-speed profiles at different MS levels. During remagnetization, the high-amplitude current pulse required to increase the PM flux linkage produces a substantial rise in stator flux, which may drive the machine to the inverter voltage limit even when the operating points would appear feasible within steady-state torque-speed profiles. For this reason, remagnetization introduces an additional limit in the torque-speed plane that must be explicitly considered in the definition of the feasible operating region.

To address this issue, a method to determine the remagnetization voltage limit from pre-calculated remagnetization maps and flux maps is proposed. The method is based on the derivation of the $MS - \lambda_r$ remagnetization characteristic, which relates the target MS to the flux linkage amplitude required during the remagnetization transient. By combining this characteristic with the inverter voltage limit and the torque-speed profiles at different MS levels, the remagnetization feasibility boundary can be directly obtained in the torque-speed plane.

The proposed analysis also showed that the remagnetization voltage limit introduces a hysteresis effect in the actual torque-speed operating region of the VFM, which can further restrict the operating area. The adoption of finer MS granularity, i.e., multi-step remagnetization, can mitigate this hysteresis effect by increasing the feasible remagnetization speed threshold and enlarging the available torque capability. However, this benefit must be balanced against the increased control complexity, additional energy consumption and thermal stress associated with repeated high-amplitude remagnetization current pulses.

The main outcomes of this chapter can be summarized as follows:

- the steady-state torque-speed profiles alone are not sufficient to define the actual feasible operating region of a VFM when remagnetization under load is required;
- the proposed $MS - \lambda_r$ remagnetization characteristic provides an effective intermediate step for evaluating the remagnetization voltage limit;
- the remagnetization feasibility boundary can be directly obtained in the torque-speed plane by combining remagnetization maps, flux maps and inverter voltage constraint;
- the remagnetization voltage limit introduces a hysteresis effect, which further reduces the practically accessible torque-speed operating region;
- multi-step remagnetization can mitigate this limitation by improving both the feasible remagnetization speed range and the available torque capability.

The validity of the proposed method was verified through simulations of the POC VFM drive. Both one-step and two-step remagnetization tests were performed at operating points close to the calculated theoretical remagnetization voltage boundary, confirming the accuracy of the calculated limit. Moreover, the method was applied to a custom acceleration-deceleration driving cycles and to the low and medium speed region of a WLTC Class 3 driving cycle. These case studies showed that the remagnetization voltage limit can strongly affect the capability of the VFM drive to follow a required vehicle mission, and that the use of intermediate MS level can effectively reduce the associated feasibility limitations.

This chapter demonstrated that the remagnetization voltage limit is a key aspect of VFM operation under load and must be explicitly considered in VFM development. The proposed method provides a practical way to complete the VFM torque-speed profile, and can be applied to any VFM drive system that the corresponding flux maps, magnetization maps and vehicle model are available.

Chapter 6

Conclusions and Future Works

This dissertation addressed the design, modeling and control of variable flux machines for sustainable electric drive applications, focusing on the use of low-coercivity PMs thus reducing the dependence on heavy rare earth materials. The research was motivated by the increasing need for high-efficiency traction systems capable of combining wide speed range operation, reduced use of critical raw materials and compatibility with practical inverter and control constraints.

Starting from the machine design perspective, the work first investigated the suitability of different rotor topologies for VFMs equipped exclusively with low-coercivity magnets. The comparative analysis showed that rotor geometry has a decisive impact on remagnetization current demand, robustness against unintended demagnetization, and torque controllability during *MS* transitions. Three IPM and SPM configurations were considered, and the SPM topology proved to be the most suitable solution. In particular, it offers the lowest remagnetization current requirement, avoids torque discontinuity during *MS* variation transients, and preserves favorable controllability under load. On the other hand, SPM machine is also the motor type most prone to undesired asymmetric demagnetization. On this basis, a proof-of-concept prototype using SPM rotor geometry and AlNiCo magnets was designed and used as the main reference machine throughout the dissertation. In addition, a traction-oriented VFM based on FeN magnets was developed to assess the applicability of the proposed concepts to full-scale traction systems.

A second major contribution of this dissertation concerns the development of control-oriented VFM simulation models. Since FEA-based approaches are too

computationally demanding for extensive control development and real-time applications, a computationally efficient electromagnetic model was proposed. A baseline model was first introduced, capable of reproducing irreversible demagnetization and remagnetization by combining inverse flux maps with a novel MS updating mechanisms. The model was then extended to include asymmetric magnetization under load by introducing the complex magnetization state, the armature-flux formulation of the magnetic model, and the two-virtual-PM representation. This improved model was shown to reproduce both the magnetization and phase variation of the PM flux linkage with high fidelity. The proposed formulation was validated against FEA simulations and then implemented in real-time on a HiL platform, demonstrating that it can reproduce complex VFM behavior with accuracy suitable for control design, while reducing computational burden by several orders of magnitude compared to FEA.

Building on this modeling framework, two dedicated control strategies were proposed to achieve smooth torque transitions during MS variation. The first strategy was developed within a DFVC framework, where the stator flux magnitude λ and the quadrature current i_{qs} are directly regulated. The MS manipulation is obtained through a dedicated flux reference, while the closed loop i_{qs} control permits maintaining torque continuity. This method, which has been jointly patented with Volvo Cars Corporation, has the advantage of permitting a unique control strategy during normal operation and during MS transients. On the other hand, it becomes unstable if a deep demagnetization is required, as the MTPV locus might be encountered. The second strategy was developed within a modified FOC framework, referred to as the $i_d - T$ strategy, where the d -axis current and torque are regulated in closed-loop. The MS variation is controlled through i_d , while the torque loop is fed by a torque estimate. This method provides extended stability even at very low MS , although $i_d - T$ is not optimal during normal operation. Both approaches were specifically designed to avoid the explicit estimation of the full dq current trajectory through intermediate MS levels, which would require either large LUTs or high online computational effort. The results showed that both control strategies are effective in significantly reducing torque pulsation during demagnetization and remagnetization transients. At the same time, their comparison highlighted pros and cons of each strategy: the DFVC approach provide simpler field-weakening integration but restricted demagnetization capability, whereas the $i_d - T$ strategy offers wider accessibility of the demagnetization region when deep MS variation

is required, but it requires precomputed field-weakening data stored in LUTs as conventional FOC strategy.

The final part of the dissertation addressed the impact of inverter voltage limit on the feasible remagnetization transients, which represents one of the most critical practical constraints for VFMs operating under load. Although this was overlooked in the previous literature, the thesis shows that the actual operating capability of a VFM cannot be evaluated only from its steady-state torque-speed profiles at different MS levels. During remagnetization, the high-amplitude current required to increase the PM flux linkage may raise the stator flux to a level that causes voltage saturation, even when the operating point appears feasible in steady-state conditions. To capture this effect, a method was proposed to determine the remagnetization feasibility boundary directly from precomputed flux maps and magnetization maps. The method is based on the $MS - \lambda_r$ remagnetization characteristic and enables rapid evaluation of the feasible torque-speed region. The analysis also showed that the remagnetization voltage constraint introduces a hysteresis effect in the practically accessible operating region, and this limitation can be mitigated through multi-step remagnetization strategies using finer MS granularity. The validity of the method was verified through simulations on the POC drive and through the analysis of driving cycle performance.

The proposed modeling and control methodologies were validated at different levels:

- **FEA-based validation:** used to assess the electromagnetic accuracy of the proposed control oriented models, including irreversible demagnetization, remagnetization and asymmetric magnetization effects;
- **Control oriented simulation validation through MATLAB/Simulink and PLECS:** used to evaluate the effectiveness of the proposed control strategies and to verify the remagnetization voltage limit under different operating conditions and driving cycles;
- **HiL validation:** used to verify the real-time computational feasibility of the proposed model and to demonstrate its suitability for controller testing and system-level vehicle simulations.

Overall, the main contributions of this dissertation can be summarized as follows:

- identification of the SPM rotor geometry as the most suitable topology for VFMs employing only low-coercivity magnets, due to its reduced remagnetization current and superior torque controllability during MS variation transients;
- definition and practical extraction methodology of demagnetization maps, remagnetization maps and flux maps for VFMs;
- development of computationally efficient VFM electromagnetic model representing irreversible and asymmetric magnetization phenomena under load;
- implementation of the proposed VFM model in a real-time HiL environment, enabling long-duration and system-level simulations;
- development of two control strategies for reducing torque disturbance during MS variations, one based on DFVC and one based on modified FOC;
- definition of a method for determining the remagnetization limit and the feasible torque-speed operating range under inverter voltage constraint;

However, due to severe delays in prototype manufacturing, the experimental activities related to the developed VFM prototype are still ongoing, and the corresponding results were not yet available at the time of writing this dissertation. As a consequence, the proposed methodologies could not be validated experimentally within the present work.

Moreover, the initial objective of this research was to focus on a VFM equipped with FeN magnets. However, since the FeN material was not available during the project, AlNiCo magnets were adopted as an alternative solution to permit the development and validation of the proposed design, modeling and control methods. This change of PM materials is responsible for the low performance of the designed VFM prototype compared to conventional PMSMs.

The main directions for future work can be summarized as follows:

- experimental validation of the proposed electromagnetic model on the POC motor, together with the proposed control strategies;
- investigation of thermal effects on MS control, with particular attention to the high-amplitude current pulses required for remagnetization;

- further optimization of the machine geometry;
- systematic benchmark comparisons with other REE-free synchronous machines;
- assessment of NVH behavior during demagnetization and remagnetization transients;
- development of a more suitable flux observer for the proposed control strategies, capable of estimating the MS online while preserving a simple structure, possibly taking into account also the PM temperature variation;
- extension of the proposed electromagnetic model to account for space harmonics in the machine.

In conclusion, this dissertation provides a coherent framework for the geometry selection, electromagnetic modeling, control development, and operating range evaluation of VFMs. The proposed methods contribute to the advancement of variable flux drive technology and support its future adoption in high performance and more sustainable electric traction systems.

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