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Abstract:

Synchronous Reluctance Motors (SynRMs) are increasingly utilized in high-performance drive systems due to their simple structure, high efficiency, and absence of permanent magnets; however, their control is challenging because of nonlinear behavior and significant parameter variations, particularly inductance changes caused by magnetic saturation. Conventional model-based control methods, including sliding mode control (SMC), depend on accurate system modeling, which limits their performance under uncertainties and disturbances. To overcome these limitations, this thesis proposes a novel model-free sliding mode control strategy based on a recurrent neural network (RNN-SMC), where the RNN is employed to learn and estimate the system dynamics online without requiring prior knowledge of motor parameters. In this approach, the equivalent control component is generated using the neural network approximation, while the discontinuous control term ensures robustness against external disturbances and model uncertainties. The integration of RNN with SMC enhances adaptability, reduces chattering, and improves tracking accuracy. The effectiveness of the proposed method is validated through simulation under various operating conditions, including parameter variations and load disturbances, demonstrating superior dynamic performance, reduced steady-state error, and enhanced robustness compared to conventional model-based SMC strategies, thereby confirming the potential of the proposed approach for reliable and high-performance control of SynRM drives.

Key words:

Synchronous Reluctance Motor (SynRM), Sliding Mode Control (SMC), Recurrent Neural Network (RNN), Model-Free Control, Extended Kalman Filter (EKF), Robust Control.

Résumé:

Les moteurs à réluctance synchrone (SynRM) sont de plus en plus utilisés dans les systèmes d'entraînement à haute performance en raison de leur structure simple, de leur rendement élevé et de l'absence d'aimants permanents ; toutefois, leur commande reste complexe en raison de leur comportement non linéaire et des variations significatives des paramètres, notamment les variations d'inductance dues à la saturation magnétique. Les méthodes de commande classiques basées sur des modèles, y compris la commande par mode glissant (SMC), dépendent fortement de la précision du modèle du système, ce qui limite leurs performances en présence d'incertitudes et de perturbations. Afin de surmonter ces limitations, cette thèse propose une nouvelle stratégie de commande par mode glissant sans modèle basée sur un réseau de neurones récurrent (RNN-SMC), dans laquelle le RNN est

utilisé pour apprendre et estimer en ligne la dynamique du système sans nécessiter de connaissance préalable des paramètres du moteur. Dans cette approche, la composante de commande équivalente est générée à l'aide de l'approximation du réseau de neurones, tandis que le terme de commande discontinu assure la robustesse face aux perturbations externes et aux incertitudes du modèle. L'intégration du RNN avec la SMC améliore l'adaptabilité, réduit le phénomène de broutement (chattering) et améliore la précision du suivi. L'efficacité de la méthode proposée est validée par simulation dans différentes conditions de fonctionnement, notamment en présence de variations paramétriques et de perturbations de charge, montrant des performances dynamiques supérieures, une réduction de l'erreur en régime permanent et une robustesse accrue par rapport aux approches SMC classiques basées sur un modèle, confirmant ainsi le potentiel de cette approche pour une commande fiable et performante des entraînements SynRM.

Mots clés:

Moteur à réluctance synchrone (SynRM), Commande par mode glissant (SMC), Réseau de neurones récurrent (RNN), Commande sans modèle, Filtre de Kalman étendu (EKF), Commande robuste.

المخلص :

تُستخدم المحركات ذات الممانعة المتزامنة (SynRM) بشكل متزايد في أنظمة القيادة عالية الأداء نظرًا لبنيتها البسيطة وكفاءتها العالية وعدم احتوائها على مغناطيس دائم، إلا أن التحكم فيها يظل معقدًا بسبب سلوكها غير الخطي والتغيرات الكبيرة في المعاملات، خاصة تغيرات المحاثّة الناتجة عن التشبع المغناطيسي. تعتمد طرق التحكم التقليدية المبنية على النماذج، بما في ذلك التحكم بالانزلاق (SMC)، بشكل كبير على دقة نموذج النظام، مما يحد من أدائها في ظل وجود عدم اليقين والاضطرابات. ولمعالجة هذه القيود، تقترح هذه الأطروحة استراتيجية جديدة للتحكم بالانزلاق بدون نموذج تعتمد على الشبكات العصبية المتكررة (RNN-SMC)، حيث يتم استخدام الشبكة العصبية لتعلم وتقدير ديناميكيات النظام أنيًّا دون الحاجة إلى معرفة مسبقة بمعاملات المحرك. في هذا النهج، يتم توليد مركبة التحكم المكافئة باستخدام تقريب الشبكة العصبية، بينما يضمن الجزء غير المستمر من التحكم تحقيق المتانة في مواجهة الاضطرابات الخارجية وعدم يقين النموذج. يساهم دمج الشبكات العصبية مع التحكم بالانزلاق في تحسين قابلية التكيف وتقليل ظاهرة الاهتزاز (Chattering) وتحسين دقة التتبع. وقد تم التحقق من فعالية الطريقة المقترحة من خلال المحاكاة في ظروف تشغيل مختلفة، بما في ذلك تغيرات المعاملات واضطرابات الحمل، حيث أظهرت النتائج أداءً ديناميكيًّا متفوقًا وانخفاضًا في الخطأ في الحالة المستقرة وزيادة في المتانة مقارنة بطرق SMC التقليدية المعتمدة على النماذج، مما يؤكد فعالية هذا النهج في تحقيق تحكم موثوق وعالي الأداء في المحركات ذات الممانعة المتزامنة.

كلمات مفتاحية :

المحرك ذو الممانعة المتزامنة (SynRM)، التحكم بالانزلاق (SMC)، الشبكات العصبية المتكررة (RNN)، التحكم بدون نموذج، مرشح كالمان الممتد (EKF)، التحكم المتين.

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In the name of Allah, the Most Gracious, the Most Merciful.

Praise be to Allah, the One, and peace and blessings be upon him who has no messenger after him, and upon his family, his companions, and all those who follow his path until the Day of Resurrection.

It is with pleasure that I dedicate these few lines as a token of gratitude and deep appreciation to all those who have helped me, directly or indirectly, in the completion of my final year project.

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I express my gratitude and thanks to the jury members for their interest in our work and the honor they bestow upon us by agreeing to judge it.

Finally, we pay tribute to those without whom this work would not have been possible.

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Dedication

We dedicate this work to...

Our Dearest Mothers

Kind, honorable, and loving, you represent for us the ultimate symbol of goodness, the source of tenderness, and the finest example of devotion.

You have never ceased to encourage us and pray for our success.

We dedicate this work to you as a token of our deep love and gratitude. May Almighty God preserve you, bless you with good health, long life, and happiness.

Our Dearest Fathers

No dedication can truly express the love, esteem, devotion, and respect that we have always held for you.

Nothing in this world can equal the sacrifices and efforts you have made day and night for our education and well-being.

This work is the fruit of the sacrifices you have made for our upbringing, education, and training.

Our Dear Sisters and Brothers

As a testimony of the attachment, love, and affection we have for you, we dedicate this work to you with our sincere wishes for happiness, good health, and success.

Our Dear Friends

As a token of the friendship that unites us and in memory of all the wonderful moments we have shared together, we dedicate this work to you and wish you a life full of health, happiness, and prosperity.

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Notations and Abbreviations

Notations

The conditions for modelling SynRM are as follows:

R_s	Stator resistor	(Ω)
L_s	Stator inductance	(H)
L_m	Magnetising inductivity	(H)
l_s	Self-inductance of a stator phase	(H)
M_0	Maximum value of the mutual inductance between the stator phase and the rotor phase	(H)
p	Number of pole pairs	($-$)
F	Coefficient due to viscous friction	(N.m.s/rd)
T_{em}	Electromagnetic torque	(N.m)
J	Total moment of inertia	(Kg.m ²)
K_p	Proportional component of the PI controller	($-$)
K_i	Integral component of the PI controller	($-$)

Axis reference points:

(d, q)	Park reference frame axes (rotating at synchronisation speed)	
(α , β)	Concordia/Clarke reference frame axes (Park reference frame fixed to the stator)	
Θ_{sr}	Angular position of the rotor in relation to the stator	(rad)
Θ_s	Angular position of the stator in relation to the axis (d)	(rad)
Θ_r	Angular position of the rotor in relation to the axis (d)	(rad)

Electrical values at the stator:

$v_{s\ a, b, c}$	Three-phase stator voltages,	
$v_{s\ d, q}$	Two-phase stator voltages in reference frame (d, q)	(V)
$v_{s\ \alpha, \beta}$	Two-phase stator voltages in reference frame (α , β)	(V)
v_s	Modulus of the stator voltage vector,	(V)
$i_{s\ a, b, c}$	Instantaneous three-phase stator currents	(A)
$i_{s\ d, q}$	Two-phase stator currents in reference frame (d, q)	(A)
$i_{s\ \alpha, \beta}$	Two-phase stator currents in reference frame (α , β)	(A)

$[V_s]_{3 \times 1} = [v_{sa}, v_{sb}, v_{sc}]^T$	Vector of stator phase voltages a, b and c	(V)
$[I_s]_{3 \times 1} = [i_{sa}, i_{sb}, i_{sc}]^T$	Vector of the instantaneous currents of stator phases a, b and c	(A)

Magnetic quantities in the stator:

$\varphi_{sa, b, c}$	Instantaneous magnetic flux at the stator	(Wb)
$\varphi_{s\alpha, \beta}$	Two-phase stator fluxes in reference frame (α, β)	(Wb)
$\varphi_{sd, q}$	Two-phase stator fluxes in the rotating reference frame (d, q)	(Wb)
φ_s	Modulus of the stator flux vector	(Wb)
$[\varphi_s]_{3 \times 1} = [\varphi_{sa}, \varphi_{sb}, \varphi_{sc}]^T$	Instantaneous flux vector for stator phases a, b and c,	(Wb)

Mechanical quantities:

ω_r	Electrical pulsation corresponding to the speed of rotation	(rad/s)
ω_s	Electrical pulsation of stator magnitudes (rotating field)	(rad/s)
f	Electrical frequency of stator magnitudes	(Hz)
N_s	Rotating field speed	(rpm)
N_r	Nominal mechanical speed of rotation	(rpm)

Transformations:

s	Laplace operator,	
$P(\theta)$	Park transformation: $X_{sa, b, c} \rightarrow X_{sd, q}$ et $X_{ra, b, c} \rightarrow X_{rd, q}$,	
$C32$	Concordia/Clarke transformation : $X_{sa, b, c} \rightarrow X_{s\alpha, \beta}$ et $X_{ra, b, c} \rightarrow X_{r\alpha, \beta}$,	

SynRM control quantities:

T_{em}^*	Electromagnetic torque setpoint	(N.m)
ω	Estimated Rotor Speed	(rad/s)
T_{em}	Estimated electromagnetic torque	(N.m)
K_p, K_i	Proportional and integral components of the PI corrector	(–)
T_s	Sampling period	(s)
f_p	Carrier frequency	(Hz)
f_r	Reference frequency	(Hz)
δ	Phase shift between the stator and rotor flux vectors	(deg)
ω_n	Natural pulsation	(rad/s)

ζ	Damping Factor	(–)
ε	Static error	(–)
S_a, S_b, S_c	Switching states of the inverter switches	(–)
T_r	Resistant Couple	(N.m)
Ω	SynRM Rotor Speed	(rad/s)
T_s	Sampling period	(μ s)
α, β	Constant values	(–)

Recurrent Neural Network quantities:

X_i	Plant state vector	(–)
$\hat{X}_i$	X_i Estimated state	(–)
w_i	Weights vector	(–)
ϖ_i	Constant value	(–)
η_i	Learning rate	(–)
ϑ_k	Input vector	(–)
$\psi_i, \phi_i, \delta_i, e_i$	Vector functions	(–)
P_i	Real error	(–)
H_i	Prediction error	(–)
Q_i, R_i	Gradient matrix	(–)
K_i	Covariance matrices	(–)
S	EKF gain matrix	(–)
L_i	Bound function	(–)
m	RNN weights number	(–)
n	Number of external inputs n State dimension	(–)
d_{ij}	State dimension	(–)
B_i	Nonnegative integers	(–)
ℓ_i	Number of connections	(–)
X_i	Variable values	(–)

Abbreviation

SynRM	Synchronous reluctance machine
IE1	Standard Efficiency.
IE2	High-Efficiency.
IE3	Premium Efficiency
IE4	Super Premium Efficiency
IE5	Ultra-Premium Efficiency
FOC	Field Oriented Control.
DC	Direct Current.
MPC	Model Predictive Control.
MRAS	Model Reference Adaptive System.
IGBT	Insulated Gate Bipolar Transistor.
IGCT	Integrated Gate Commutated Thyristor.
MOSFET	Metal Oxide Semiconductor Field Effect Transistor.
PMSG	Permanent Magnet Synchronous Generator.
PWM	Pulse Width Modulation.
SVM	Space Vector Modulation.
THD	Total Harmonic Distortion.
CARMA	Control Auto Regressive on Moving Average
NTVS	Nonlinear Time-Varying Systems

General Introduction

General Introduction

In recent years, alternating current (AC) machines have witnessed increasing adoption in high-performance industrial drive systems, particularly those requiring efficient torque control, wide speed range, and low maintenance. Among the various emerging technologies, the Synchronous Reluctance Machine (SynRM) has gained significant attention as a cost-effective and environmentally friendly alternative to conventional electric machines.

The SynRM operates on the principle of magnetic reluctance, generating torque through the natural tendency of the rotor to align with the path of least magnetic reluctance created by a rotating stator field. Unlike permanent magnet synchronous motors (PMSM), SynRMs contain no magnets on the rotor, thereby eliminating risks associated with demagnetization and reducing reliance on rare-earth materials. Additionally, unlike induction motors, the SynRM rotor has no copper windings or cage, which leads to lower rotor losses and simpler, more robust construction.

One of the key advantages of SynRM is its high efficiency over a wide operating range, combined with low manufacturing cost and excellent mechanical reliability. These features make it an attractive solution for applications such as electric vehicles, high-speed spindles, pumping systems, compressors, and ventilation systems. However, the SynRM also presents inherent challenges. Its power factor is relatively low compared to permanent magnet machines, and its torque density is moderate unless the rotor design is carefully optimized. Furthermore, the highly nonlinear behavior of the magnetic circuit, the strong coupling between the direct-axis (d) and quadrature-axis (q) currents, and the sensitivity to parameter variations demand the use of advanced control strategies to fully exploit the machine's potential.

To address these challenges, robust and high-performance control techniques are essential. Among them, Field-Oriented Control (FOC) has become a standard approach, enabling decoupled control of flux and torque similar to separately excited DC machines. More recently, Sliding Mode Control (SMC) has emerged as a powerful alternative or complement due to its inherent robustness against model uncertainties, external disturbances, and parameter variations. SMC forces the system's state trajectory onto a predefined sliding surface, ensuring stable and precise behavior even under adverse conditions.

This thesis is structured into three chapters that collectively explore the modeling, control, and performance optimization of the SynRM using advanced sliding mode control strategies:

Chapter 1 provides a comprehensive overview of the SynRM, including its historical development, different rotor structures (flux barriers, axially/radially laminated, permanent magnet-assisted), operating principle based on reluctance variation, and the influence of the saliency ratio (L_d/L_q) on torque and power factor. It also reviews classical control methods such as FOC, DTC, MPC, and introduces the motivation for using SMC.

Chapter 2 is dedicated to the detailed mathematical modeling of the SynRM. It presents the electrical equations in the three-phase frame (abc), followed by transformations to the stationary two-phase frame ($\alpha\beta$) using Clarke transformation, and finally to the rotating frame (dq) linked to the rotor using Park transformation. This chapter also covers the nonlinear state-space representation, the modeling of the two-level Voltage Source Inverter (VSI), and Pulse Width Modulation (PWM) techniques for machine supply.

Chapter 3 delves into the theory and application of Sliding Mode Control (SMC) for the SynRM drive system. It explains the fundamentals of variable structure systems, the design of the sliding surface using Slotine's approach, the existence condition derived from Lyapunov's stability criterion, and the formulation of the control law as a combination of equivalent control (U_{eq}) and discontinuous control (U_n). Furthermore, the use of artificial recurrent neural networks to further enhance the control robustness and performance of SynRM machines is presented. This includes a comparative analysis of classical and intelligent sliding mode control, focusing on error tracking, current ripple reduction, and robustness against parameter variations.

Finally, a general conclusion will summarize the key findings of this work, highlight the main contributions, and provide insights into future research directions, including chattering mitigation techniques and the integration of intelligent control methods such as fuzzy logic or neural networks.

Chapter I: Generalities of Synchronous Reluctance Machines

I.1 Introduction

In recent years, the growing demand for energy-efficient, cost-effective, and robust electric machines has driven renewed interest in alternative motor technologies. Among these, the Synchronous Reluctance Machine (SynRM) has emerged as a promising candidate due to its simple and magnet-free rotor structure, high efficiency potential, and low manufacturing cost. This chapter provides a comprehensive overview of the SynRM, covering its historical development, operating principles, rotor geometries, and the influence of the L_d / L_q saliency ratio on performance. The chapter also reviews the most widely used control strategies for SynRM drives, including Field Oriented Control (FOC), Direct Torque Control (DTC), Model Predictive Control (MPC), and Sliding Mode Control (SMC), highlighting their advantages and limitations. The aim is to establish a solid foundation for understanding the behaviour and control of the SynRM under both steady-state and transient conditions.

I.2 Synchronous reluctance Machine (SynRM)

Synchronous Reluctance Machines (SynRMs) are an alternative motor technology that operates based on the reluctance concept to create motion and torque. Similar to Switched Reluctance Machines (SRMs), SynRMs do not contain permanent magnets and are fed by sinusoidal currents to generate a rotating magnetic field. They have a robust structure with a stator resembling that of an induction motor and a rotor without magnets or copper windings. These machines have the potential to be a low-cost solution for various applications.[\[BOU23\]](#)

I.3 History of the development of SynRM

The development of synchronous reluctance motors (SynRM) can be traced back to the 1930s, when researchers began studying "unexcited salient pole synchronous motors" (variable reluctance synchronous motors) for applications requiring precise and constant speed along with autonomous startup. These early motors had limited use due to their low power factor and efficiency [\[SMW93-Lub03\]](#). In the 1960s, interest in these motors was renewed, particularly in England, leading to the emergence of segmented rotor, flux barrier, and axially laminated rotor designs, which aimed to increase the saliency ratio and improve performance. At the same time in France, the Vernier reluctance motor with a large number of teeth and high torque density became popular for low-speed direct drive applications. The term "Switched Reluctance Motor" was introduced in 1969, and in the 1990s, electronic-controlled double saliency reluctance machines gained traction in academia and industry, with

companies like Allenwest Ltd. in the UK and Sicme-Motori in Italy commercializing them. In recent years, SynRM has garnered interest due to its simple structure, high efficiency, low manufacturing cost, and robust operation. However, challenges remain in reducing noise and increasing efficiency, prompting the need for ongoing research and optimization in design and control methods. Current applications of SynRM include the textile industry, machine tools, high-speed applications, and more recently, electric vehicles, pumping, and ventilation systems. [\[TRU16-TP15\]](#)

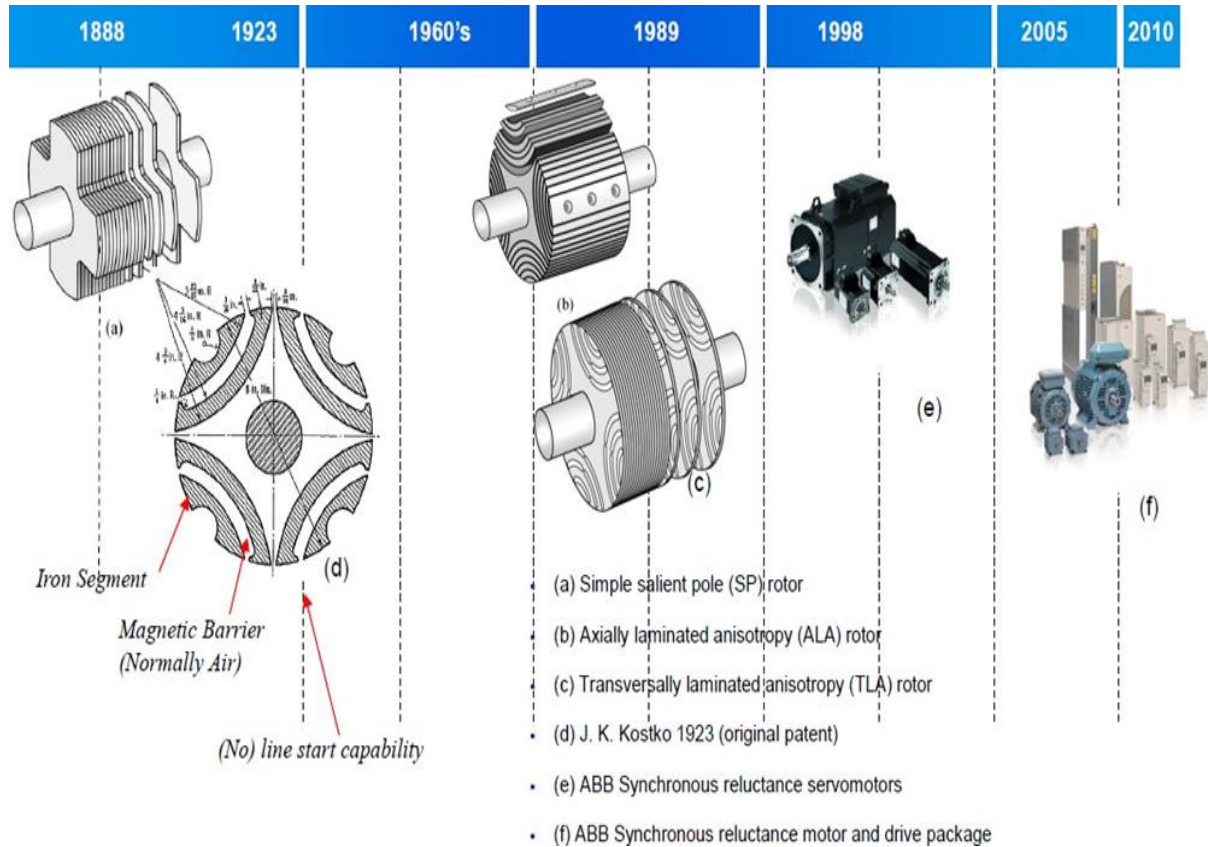


Figure I-1: History of SynRM development [\[TRU16\]](#).

I.4 Different structures of Syn-RM

The purpose of this section is to provide a concise summary of the various types of Syn-RM based on their rotor geometry and materials. This is significant because there is a wide range of structures available for Syn-RMs. Numerous researchers and designers have proposed optimized rotor concepts with saliency maximization as a main criterion [\[MOH24\]](#).

I.4.1 Flux barriers Syn-RM

In order to increase the saliency ratio of the motor, empty spaces of material have been added into the rotor. These gaps are called flux barriers. Their purpose is to restrict the flux in the q -axis direction without hindering it in the d -axis.

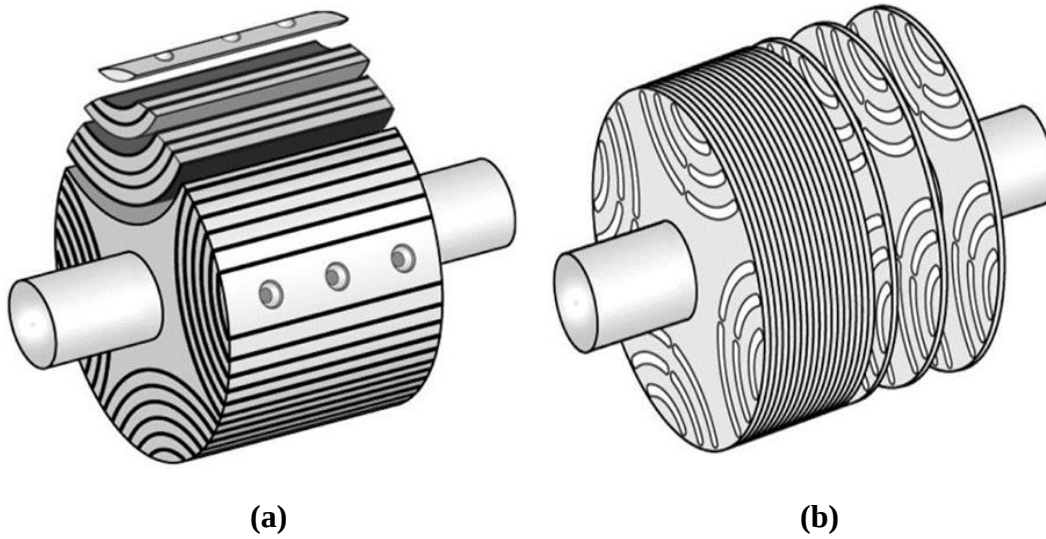


Figure I-2: Radially laminated (a) and axially laminated (b) flux barriers type Syn-RM rotors. Figure I-1 depicts two distinct types of motors that have flux barriers. The first one (Figure I-2(a)) is radially laminated and its manufacturing process is complicated. This type of rotor has limited mechanical strength, so it is only suitable for low-speed applications, and its power output is restricted to 2 kW. Nevertheless, it is noteworthy that the saliency ratio in a two-pole motor can reach values as high as 20 [BOI92] and 10 for a four-pole motor [TJE93].

The second rotor shown in Figure I-2 (b) is an axially laminated motor. It is easier to manufacture compared to the radially laminated rotor and has superior mechanical strength. However, the saliency ratio of this rotor type is not as high as that of the radially laminated motor, with a maximum value of 13 for a two-pole rotor according to [TSA06]. While the flux barriers are typically just air, it is also possible to use a non-magnetic material instead to boost the rotor's mechanical durability.

I.4.2 Permanent magnet assisted Syn-RM

Recently, there has been growing interest in considering Permanent Magnet Assisted Synchronous Reluctance Motors (PMA-SynRM) as a potential alternative motor drive in applications demanding high performance. This interest is particularly notable in the context of electric vehicle traction systems, where the electric motor needs to operate effectively over a wide range of speeds [JOO17-SAR19]. The Interior PMSM (IPMSM) employs a combination of magnets and reluctance torque outputs to achieve high efficient drives. Additionally, the utilization of flux-weakening control techniques enables consistent power output or operation within high-speed flux-weakening ranges. Nonetheless, certain drawbacks, such as the relatively substantial d-axis current at high speeds and the motor's uncontrollable transition to generator mode following unexpected inverter shutdowns, significantly constrain the motor's use in high-speed electric vehicles [MOR01-JAN99].

On the other hand, akin to the IPMSM, the PMa-SynRM, offers a promising solution to these challenges. Furthermore, the application of the flux-weakening control technique permits the establishment of extended operational zones with constant power output. Figure I-3 illustrates a configuration of a 4-pole PMa-SynRM featuring a ferrite magnet and reference d,q-axis. As depicted in the figure, the main difference between the conventional IPMSM and the showcased PMa-SynRM lies in the magnetization of the bar magnets along the negative q-axis [JAN99-BAE09].

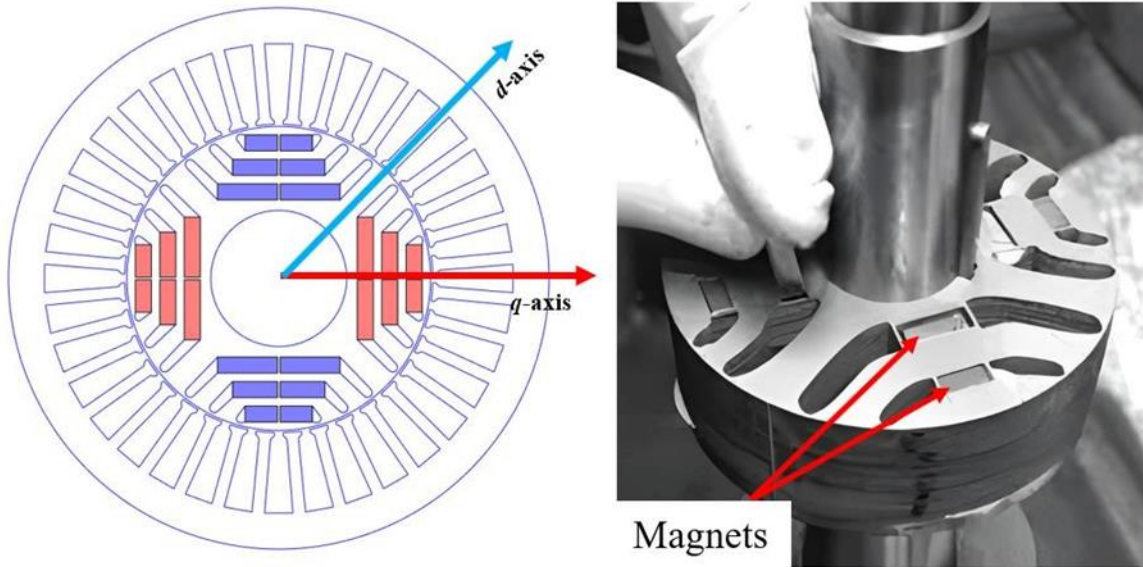


Figure I-3: PMa-SynRM rotor structure view [MOR01].

I.4.3 Switched reluctance machine

The origin of the switched-reluctance machine (SRM) can be traced back to the 19th century, with records of its application found in a locomotive propulsion system built by Davison in 1838 [MIL93]. The research on reluctance machines began even earlier in 1824 [MIL01]. However, the modern research on SRM started with the development of modern power electronics in the late of 1960s [LAW83].

SRM's are a promising alternative to AC drive systems that do not require permanent magnets, although their use has been limited compared to other types of machines. The SRM uses a step- by-step control system to generate torque and has shown to have high specific power and high- speed capabilities. Its rotor is made solely of steel, making it a sturdy option. However, the SRM's complex control system, high noise level, and unusual power electronics architecture have posed challenges in some application systems. As depicted in Figure I-4, the stator and rotor of a typical SRM have a salient geometry, and the winding on the stator pole is often single-toothed, making it difficult to analyze the torque performance due to its highly nonlinear electromagnetic behavior [LAW83].

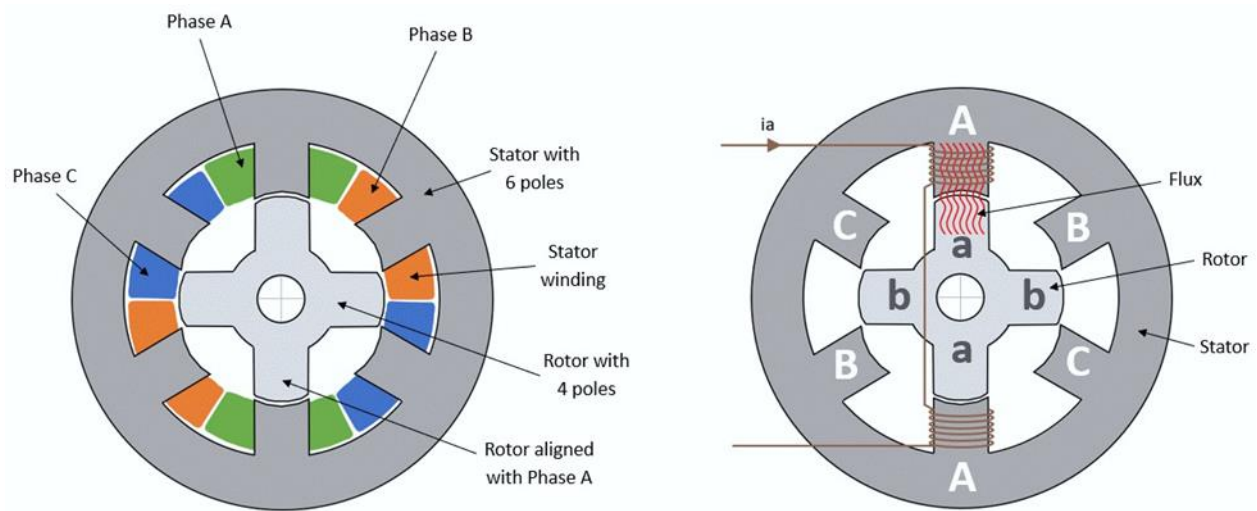


Figure I-4: 6/4 type switched reluctance machine (SRM) [\[LAW80\]](#).

I.5 Working Principle of SynRM

I.5.1 Reluctance Concept

The operating principle of SynRM is based on two main concepts:

- The reluctance effect.
- The application of a rotating sinusoidal magneto-motive force generated by the stator of a three-phase induction type machine.

The main idea of the variable reluctance concept is shown in figure (I.5). Objects

(a) and (b) are made by isotropic magnetic material and are exposed to a magnetic field Ψ . Object (a) does not generate a couple since it has the same geometric dimensions in all directions and hence the same reluctances in the d-axis and q-axis. This is called an isotropic object, whereas object (b) has different geometric dimensions in the d and q-axis which refer to different reluctances; this is called an anisotropic object. If the d-axis of the object has a misalignment angle of ε with the magnetic field, this will introduce a field distortion that is aligned with the q-axis and will increase the reluctance in the d-axis; see the field solution in Figure (I.5) (right). As a result, an electromagnetic potential energy is created that can develop an electromechanical torque to force the object to be aligned with the minimum reluctance at the field direction (d-axis) of the object (b) [\[RM11-Tag15\]](#).

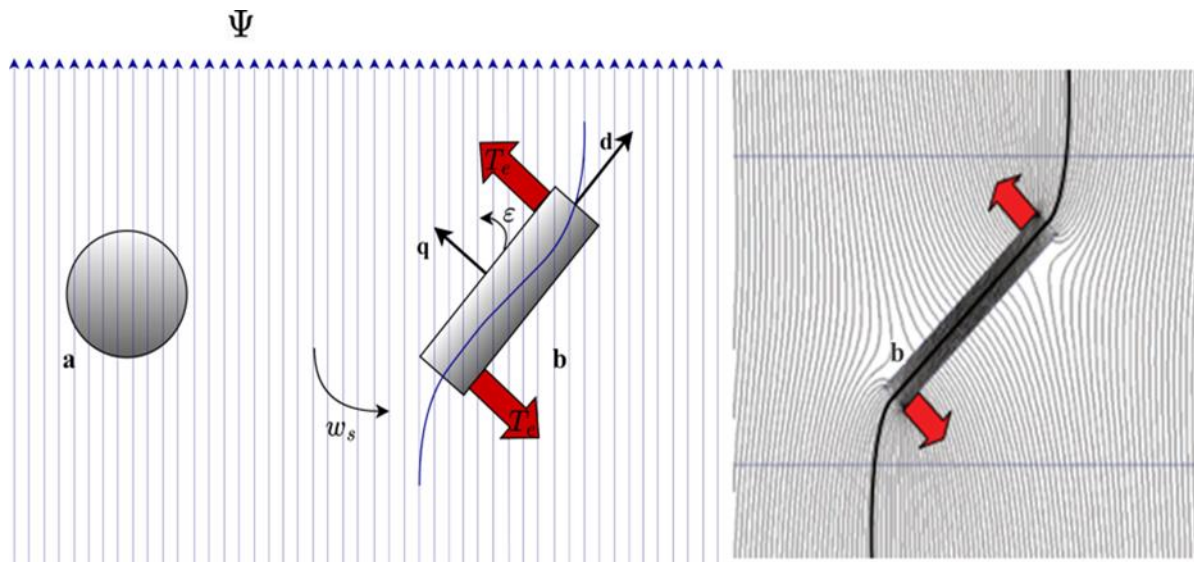


Figure I-5: Two objects in a magnetic field Ψ a) isotropic geometry b) anisotropic geometry [\[Tag15\]](#).

In the case of SynRM, this same principle of reluctance variation will create the couple. The rotor will try to align with the magnetic field created by the stator coils. As the stator has a rotating magneto-motive force, the rotor will also rotate generating the movement [\[Mar16\]](#).

I.5.2 Functioning

The SynRM (Synchronous Reluctance Motor) is a type of electric motor that operates on the principle of reluctance torque. It has a stator with windings that produce a rotating magnetic field, similar to that of an induction motor. It consists of a three-phase winding with p pairs of poles, which is fed by a balanced three-phase current system of pulsating currents w_s and distributed to create a rotating magneto-motive force (mmf) that is as sinusoidal as possible. However, unlike a conventional salient-pole synchronous motor, the SynRM does not have an excitation winding in the rotor. The rotor is constructed solely of salient poles using air gaps and steel segments and is designed with a special shape to induce torque through reluctance. This geometric anisotropy of the rotor causes it to continuously follow the rotating field in a steady state, resulting in a highly efficient motor that provides high torque and precise control at low speeds. SynRM technology is becoming increasingly popular in various industrial applications such as compressors, pumps, and fans due to its high efficiency, low maintenance, and reliable performance [\[Tag15-Ram06-Mar16\]](#).

I.6 The advantages and disadvantages of SynRM [\[EPO23-BEL-DJA24\]](#)

I.6.1 Advantages

- The machine maintains a constant speed under all operating conditions due to its magnetic locking principle. By producing synchronized torque, it effectively handles load variations whether underloaded or overloaded while keeping the speed stable.
- Unlike conventional synchronous motors, which are not self-starting and require a starting mechanism, this machine self-starts thanks to the reluctance principle. This contributes to its enhanced durability and simpler design, resulting in reduced maintenance requirements.
- Additionally, unlike switched reluctance motors, it does not suffer from torque ripple, leading to smoother operation.
- One of the key advantages of using synchronous motors is the ability to control the power factor. An over-excited synchronous motor, which operates with a leading power factor, can be used in parallel with induction motors and other lagging power factor loads to improve the overall system power factor.
- Synchronous motors can be designed with wider air gaps compared to induction motors, enhancing mechanical stability. Moreover, the electromagnetic power developed in a synchronous motor varies linearly with the applied voltage.
- Synchronous motors also typically achieve higher efficiency often exceeding 90% especially in low-speed and unity power factor applications.
- Notably, this machine does not require a DC supply for excitation, simplifying its construction and further reducing maintenance needs.
- For high-quality synchronous reluctance motors that offer reliable performance without the need for continuous maintenance, consider reaching out to Epoch Automation Pvt. Ltd

I.6.2 Disadvantages

- The power factor is lower in comparison to that of permanent magnet synchronous machines.
- It exhibits lower efficiency relative to high-end permanent magnet machines.
- Exhibits lower torque density relative to machines utilizing magnets or windings.
- Susceptible to higher levels of noise and vibration due to saliency-induced effects.

- Produces lower starting torque than induction motors unless specifically optimized.
- Dynamic performance is limited in the absence of control systems.
- Control strategies are required to maintain synchronization under varying load Condit.

I.7 Applications of Synchronous reluctance Machine

- It is popularly used in many low power applications such as fiber spinning mills because of inherent simplicity, robustness of construction and low. cost.
- Widely used for many constant speed applications such as recording instruments, timing devices, control apparatus and phonograph.
- Used as proportioning devices in pumps or conveyors.
- Applied in axillary time mechanism,
- Used in processing of continuous sheet or film material.
- Used in regulators and turntables.
- Applied in wrapping and folding machines.
- It can be used in synchronized conveyors.
- In metering pumps also, the synchronous reluctance motor is used.
- Used in synthetic fiber manufacturing equipment

I.8 Influence of the L_d/L_q factor on machine performance

The saliency rate, given by the following equation, is an important parameter for monitoring the performance of a SynRM.

$$\zeta = \frac{L_d}{L_q} \quad (I.1)$$

To achieve a SynRM with high performance, the main goal is to obtain a high L_d value and a low L_q value, which results in maximum d -axis flux and minimum q -axis flux.

The graphic vector diagram is an essential means of displaying and understanding the operating principles and characteristics of SynRM's performance.

The d -axis and q -axis rotor magnetic flux components are differentiated in the vector diagram, d - axis corresponding to the high reluctance direction and q -axis representing the low reluctance direction. Torque is produced by changing the reluctance between these axes.

The stator current vector in the vector diagram is broken down into its d -axis and q -axis components, which are vital in determining the torque output and the machine's overall performance [MIL25].

The electromagnetic steady-state torque produced by this machine is expressed as follows:

$$T_e = p \cdot (L_{sd} - L_{sq}) I_{sd} I_{sq} \quad (I-2)$$

To boost the torque without changing the value of the current, the inductance of the direct axis (L_d) must be increased in relation to the inductance of the quadrature axis (L_q).

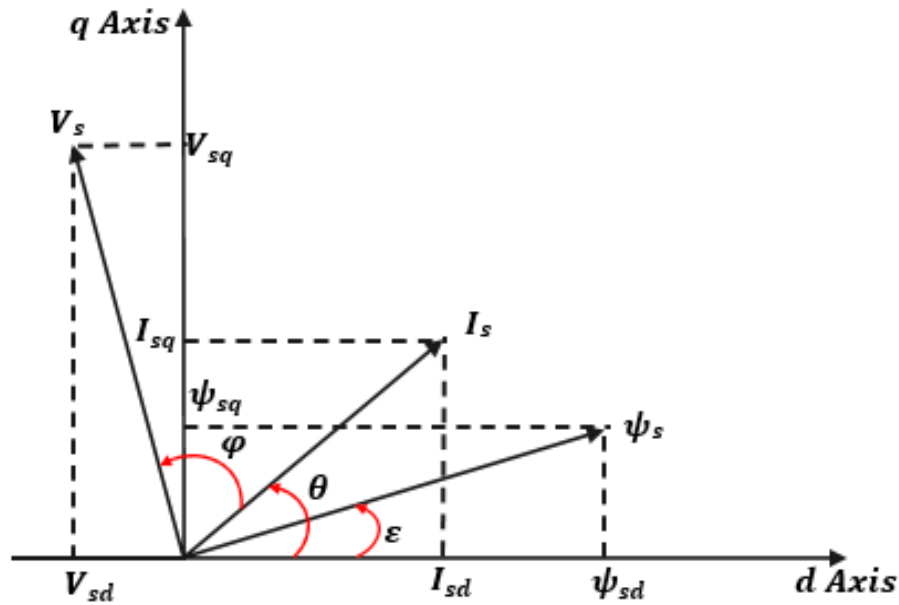


Figure I-6: SynRM steady-state vector diagram.

In other hands, to increment the torque, the saliency ratio (ζ) must be augmented [\[HAM09\]](#).

The angle θ , represents the position (I_s) of the stator current vector relative to the d-axis. The current vector in the stator is kept fixed in steady state.

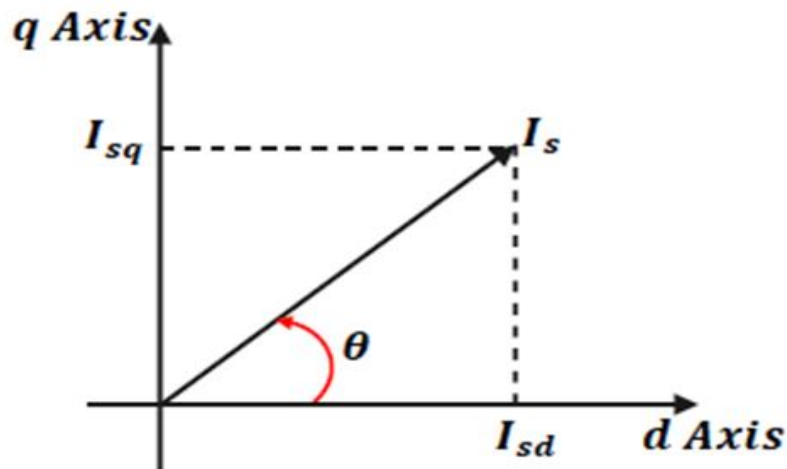


Figure I-7: Position of the stator current on the (d - q) axis.

The amplitude of the stator current is defined by the following equation:

$$I_s = \sqrt{I_{sd}^2 + I_{sq}^2} \quad (I.4)$$

Electromagnetic torque as a function of stator current I_s and θ :

$$T_e = p(L_d - L_q)I_s^2 \cdot \sin(2\theta) \quad (I.5)$$

The preceding relationships show that when the stator current is fixed at a specific value, the maximum torque is reached at $\theta = \pi/4$, which corresponds to $I_{sd} = I_{sq}$, a mode of operation associated with a particular control strategy. Fixing $\theta = \pi/4$ in equation (I.5) gives equation (I.6), which reveals the saliency ratio ζ .

$$T_e = pL_d I_s^2 \cdot \left(1 - \frac{1}{\zeta}\right) \quad (I.6)$$

The SynRM is an AC machine designed to be supplied with three-phase sinusoidal currents in steady state. The machine's power factor, expressed as the phase shift between the fundamental of the current in the line and the respective phase-to-neutral voltage, can be defined. It also corresponds to the ratio between the apparent power absorbed by the machine and the active power. It is essential that this ratio is as close as possible to 1 in order to limit the electrical power of the source feeding the machine. By ignoring losses in the machine model, we get a simple expression for the power factor:

$$\cos\varphi = \frac{(\zeta-1) \cdot \sin\theta}{\sqrt{(\zeta)^2 + \tan^2\theta}} \quad (I.7)$$

If a specific strategy of control is applied (requiring $\tan\theta = \sqrt{\zeta}$) the factor of power is maximized and only depends on the ratio L_d/L_q

The expression of the power factor is then given by the expression [\[BET93- LUB03\]](#).

$$(\cos\varphi)_{max} = \frac{(\zeta-1)}{(\zeta+1)} \quad (I.8)$$

In Figure (I.8) we can see how the power factor varies as a function of the saliency ratio L_d/L_q . The power factor becomes more important for saliency ratios greater than 8. Based on equations (I.6) and (I.8), to achieve optimum machine performance, the structure of the rotor has to be chosen to reach the maximum possible value of L_d and the highest possible L_d/L_q ratio [\[LUB03\]](#).

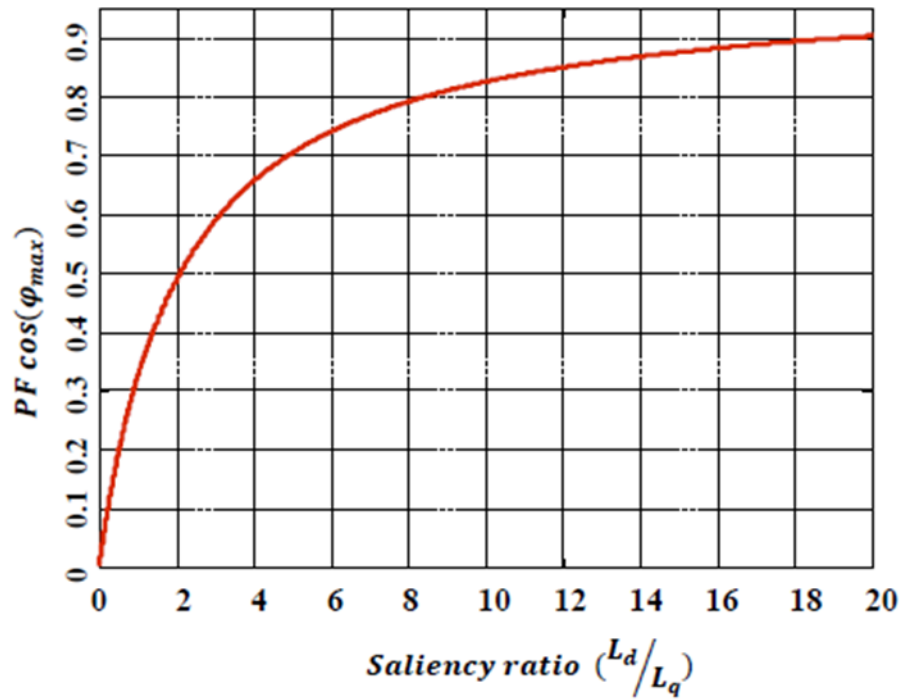


Figure I-8: Power Factor as a relation to the saliency ratio L_d/L_q [LUB03].

I.9 Literature review of the synRM control drives

The control of electrical machines has been the subject of many researches and works for a long time. In order to take advantage and exploit their best performances, different control techniques have been developed to control the Syn-RM. The main controls and their derivatives belong to the same category, so in this section we will focus on the most well-known techniques that have been the subject of numerous publications [MOH24].

I.9.1 Field oriented control

The vector control of AC machines is now well-known. Many industrial companies market speed drives for synchronous and asynchronous machines using this control method. The principle of vector control is identical to that of separately excited DC machine control. However, it is necessary to use a specific reference frame where the electromagnetic torque is expressed simply based on the components of currents along two axes (d-axis and q-axis). Typically, the d-axis component of the stator current acts as the excitation and allows for adjusting the flux value in the machine. The q-axis component serves as the armature current and enables torque control. With this type of control, high-performance drive systems are achieved, capable of delivering the rated torque at the rated speed even during startup.

The use of the vector control technique for synchronous motor drive systems is becoming more popular due to the rapid development of microprocessors and digital signal processors (DSP's). The vector control method, which was first introduced in the 1960's by Blaschke,

Hasse, and Leonhard in Germany, has gained popularity due to advancements in technology [GAR06- WAN09].

The underlying principle of FOC is to separate the torque and magnetic field components of the motor and control them independently. This is achieved by transforming the stator currents from a stationary reference frame (abc) to a rotating reference frame (d,q) using Park's transformation. The d -axis is aligned with the rotor flux, while the q -axis is perpendicular to it. By controlling the current in the d -axis, which produces the torque, and the current in the q -axis, which produces the magnetic field, the torque and magnetic field components can be decoupled. The Park's transformation concept was introduced by Park in 1929, and FOC controllers are based on the reference frame theory, maintaining a 90-degree angle (electrical) between the stator and rotor field components [CAS02]. It can control the current individually along d - q axes in the machine. A typical FOC structure for Syn-RM is shown in Figure I-9.

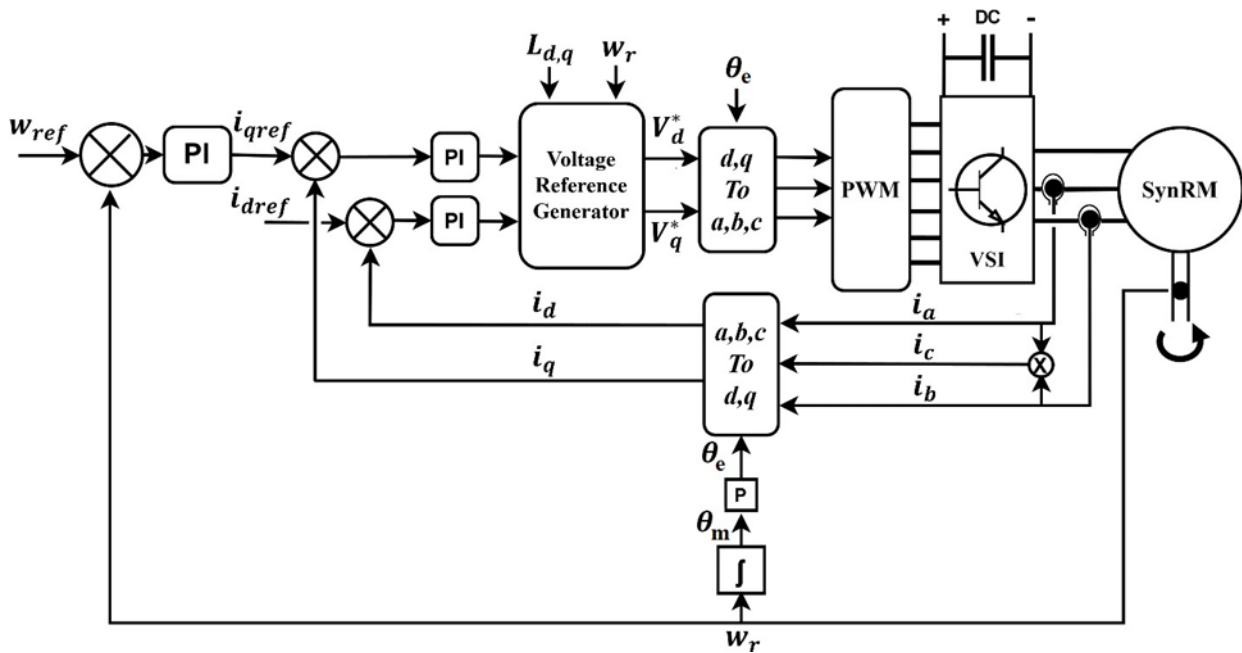


Figure 1-9: Schematic block of Field Oriented Control (FOC).

I.9.2 Direct torque control

As depicted in Figure I-10, the direct torque control (DTC) principle involves selecting voltage vectors directly based on the difference between the desired and actual values of torque and flux. This is done by comparing torque and flux errors using hysteresis comparators (Figure I-11). These comparators help choose a voltage vector from a predefined set without needing pulse width modulation. Instead, one of six voltage vectors is used consistently for a specific sampling period [BOL11]. DTC offers benefits like indirect control of stator current and voltage, achieving nearly sinusoidal stator flux and currents, and fast

dynamic response from the machine. However, the DTC has some drawbacks such as high torque ripple and variable switching frequency caused by the variable errors generated by hysteresis comparator. Thus, the settings of these comparators (bandwidths) have a direct effect on the switching frequency and the steady state error of torque and flux [ARU18].

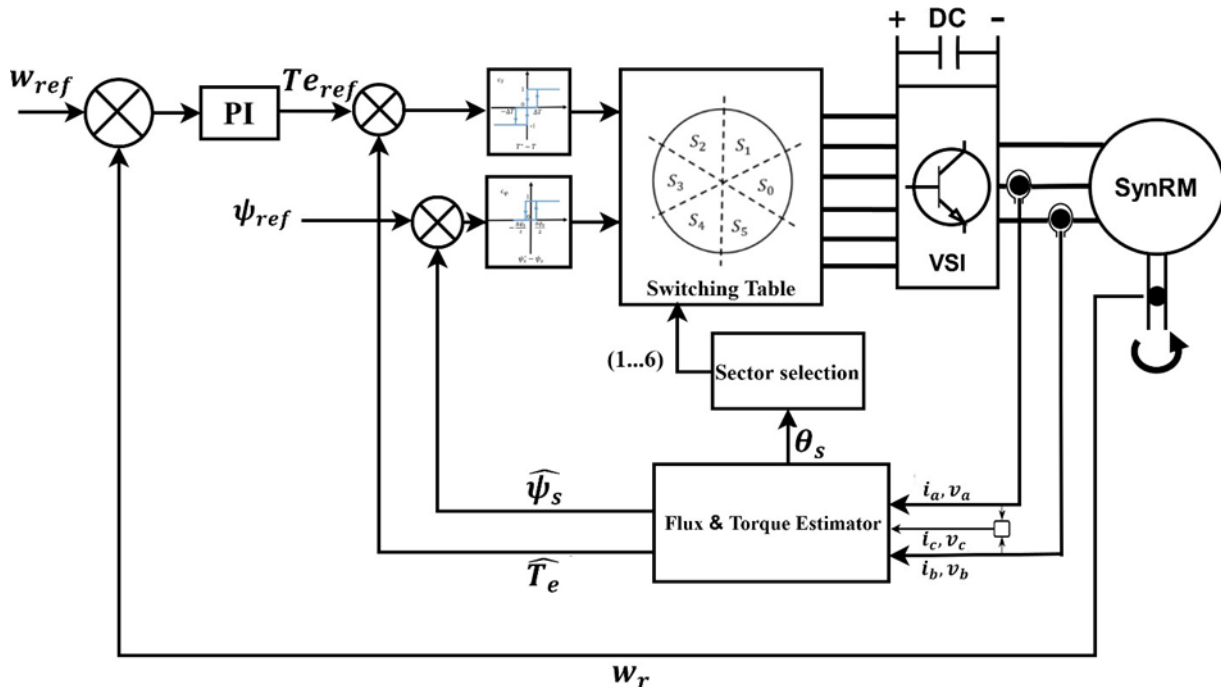


Figure I-10: Schematic block of Direct Torque Control (DTC).

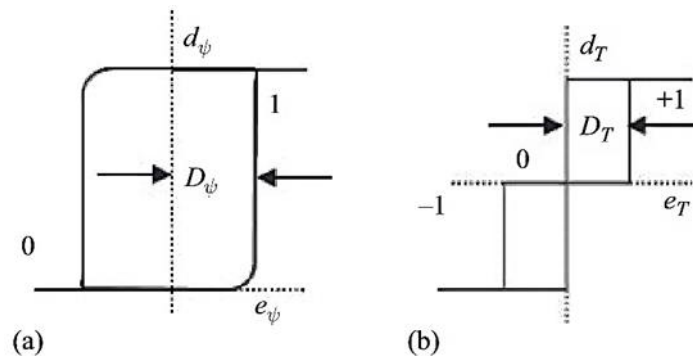


Figure I-11: Flux and torque comparators.

I.9.3 Space vector modulation based direct torque control

Direct Torque Control - Space Vector Modulation (DTC-SVM) is a control technique used in electric motor drives to achieve high-performance torque and speed control. It combines the principles of Direct Torque Control (DTC) and Space Vector Modulation (SVM) to achieve superior performance [AMM17]. As shown in Figure I-12. The objective of SVM is to choose suitable switching vectors to minimize the distortion caused by current harmonics. By

carefully selecting the voltage vectors within each sampling period, a constant switching frequency is maintained. As a result, the DTC-SVM technique reduces torque fluctuations and effectively manages the stator flux and torque. Furthermore, it offers several advantages over other control techniques such as fast dynamic response, reduced torque ripple, and simplified control structure. It is widely used in industrial applications such as electric vehicles, wind turbines, and robotics [AMM16].

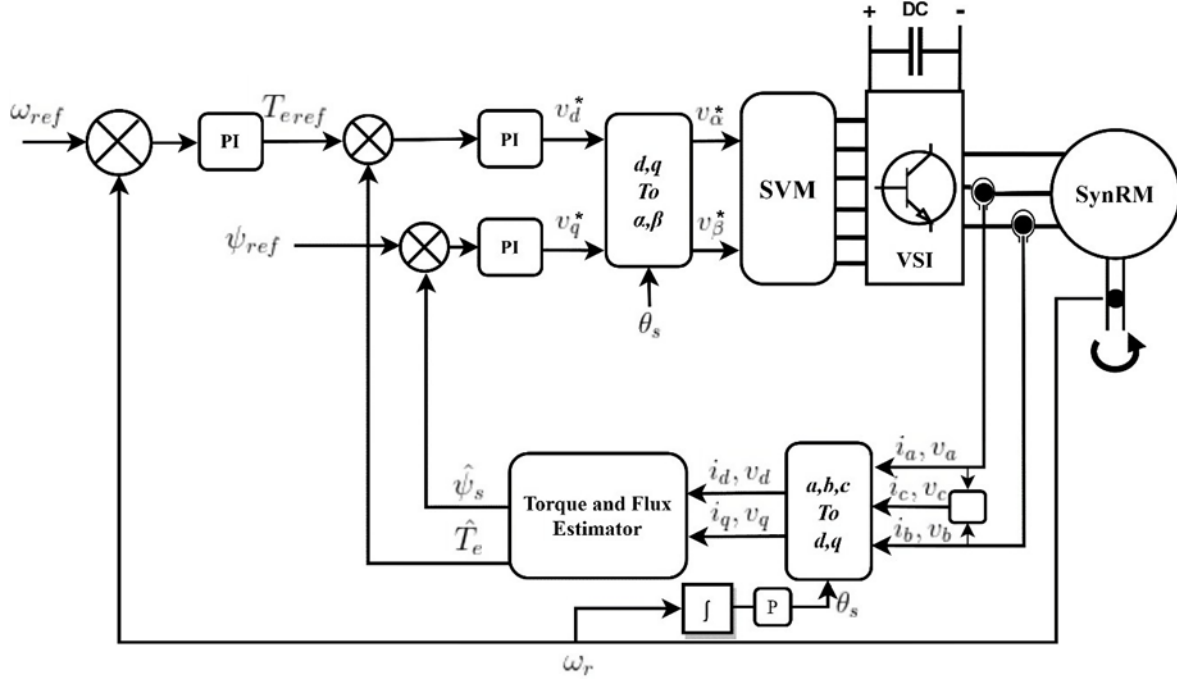


Figure I-12: Schematic block of SVM based Direct Torque Control (DTC-SVM).

I.9.4 Model predictive control

Model Predictive Control (MPC) is a type of advanced control technique that predicts the future behavior of a system and determines the optimal control action to achieve a desired outcome. In the context of Syn-RMs, MPC is a control strategy that can be used to optimize the machine's performance by adjusting the stator currents to achieve the desired torque and speed. As depicted in Figure I-13 MPC of Syn-RMs involves creating a current model of the machine and its associated power electronics which is why this approach is referred to as Predictive Current Control (SMC), and then using this model to predict the future behavior of the system. The control algorithm then computes an optimal control action that minimizes a cost function based on the predicted future quantities [LIU17]. This cost function can be designed to optimize a range of performance criteria, such as minimizing torque ripple, maximizing efficiency, or achieving a desired speed trajectory [BEN22].

MPC of Syn-RMs has been shown to provide superior performance compared to traditional control techniques such as field-oriented control (FOC) and direct torque control (DTC). This

is because MPC can take into account the non-linear behavior of Syn-RMs, which can result in better accuracy and stability of the control system. However, MPC requires a more complex control algorithm and higher computational power, which may limit its applicability in some real-time control applications [ARO17]. The standard schematic diagram of Syn-RM is depicted in Figure I-13, where Maximum Torque Per Ampere (MTPA) is employed to optimize the i_d and i_q current references.

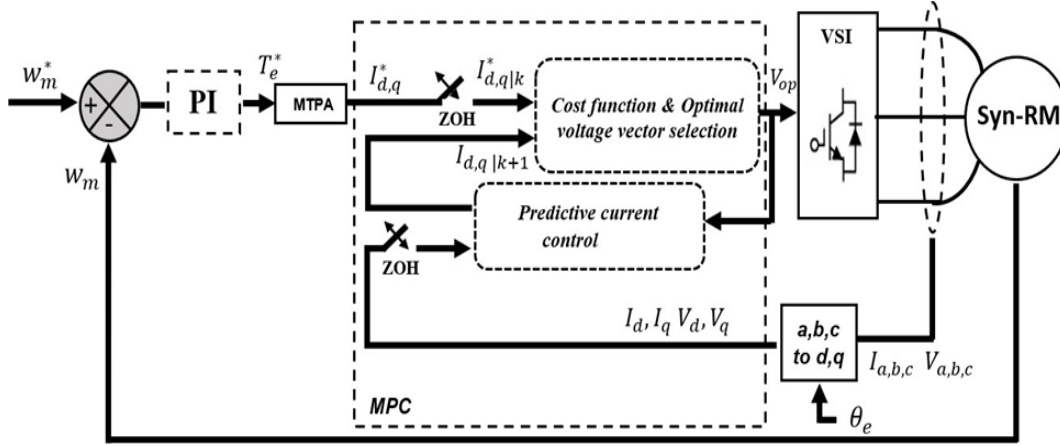


Figure I-13: Schematic block of Model Predictive Control (MPC).

I.10 Sliding mode control in the literature

Variable structure control systems (VSCS) evolved from the pioneering work in Russia of Emel'yanov and Barbashin in the early 1960s. The ideas did not appear outside of Russia until the mid-1970s when a book by Itkis (1976) and a survey paper by Utkin (1977) were published in English. VSCS concepts have subsequently been utilised in the design of robust regulators, model-reference systems, adaptive schemes, tracking systems, state observers and fault detection schemes. The ideas have successfully been applied to problems as diverse as automatic flight control, control of electric motors, chemical processes, helicopter stability augmentation systems, space systems and robots [EDW98].

In the scientific literature, Sliding Mode Control (SMC) has been extensively studied and applied as a robust control strategy for electric machines and drives, addressing challenges such as parameter uncertainties, external disturbances, and nonlinear dynamics. Research has demonstrated the effectiveness of SMC in speed regulation and torque control of induction and synchronous machines, especially in applications like electric vehicles and industrial drives. For example, studies on SMC-based speed control for permanent magnet synchronous motors (PMSMs) in electric vehicle drives have shown fast dynamic response and enhanced robustness, often using multivariable estimation to improve performance under load variations

[SEN25]. Similarly, advanced SMC techniques with exponential reaching laws have been proposed for brushless DC (BLDC) motor control, showing significant improvements in disturbance rejection and reduced chattering compared to classical controllers [SEY26]. Comparative research also highlights SMC's superior performance over conventional vector control in electrical machines, with better tracking and stability in variable-speed systems [MOU18]. Additional work has focused on integrating SMC with adaptive observers for sensor less induction motor control, enhancing robustness when full state measurements are unavailable [HAM23]. Overall, the literature reflects a growing trend toward applying modified SMC algorithms such as high-order or adaptive versions to improve control accuracy and performance in electric machine drives across a range of modern applications [SAB11].

I.10.1 Sliding mode control of the SynRM

Several recent studies have focused on developing advanced sliding mode control strategies for electric motor drive systems to enhance dynamic performance and robustness. In reference [BOU25], a PID-SMC controller was proposed, integrating proportional, integral, and discontinuous derivative actions within the sliding mode framework, resulting in improved transient response, reduced chattering, and enhanced robustness against model uncertainties and external disturbances. In reference [UAS23], an advanced reaching law for sliding mode control was introduced for the speed control of a synchronous reluctance motor (SynRM), along with the development of a nonlinear observer to effectively eliminate the influence of uncertainties on the output. In reference [LIN21], a novel control design based on ANFTSMC combined with a nonlinear disturbance observer (NDO) was proposed for the SynRM drive system. A non-singular sliding mode control law with an adaptive gain reaching law was developed to adjust lumped uncertainties online and generate current loop commands. Additionally, the time-varying parameters of the SynRM and the controller parameters were estimated online using a proposed analytical magnetic model. In reference [BEN13], a new robust control approach for the induction motor was presented, combining sliding mode control with fuzzy logic control (FLC) to handle parameter uncertainties and external disturbances. Finally, in reference [SHU25], a novel sensor less method was developed and evaluated for accurately estimating the SynRM angular speed and position over a wide speed range using a model-based observer, where both the SynRM mathematical model and the entire observer scheme were formulated in the estimated $\gamma\delta$ reference frame.

I.11 Conclusion

This chapter has presented the fundamental concepts related to the Synchronous Reluctance Motor (SynRM), including its structure, operating principles, and key characteristics. Different rotor configurations and their impact on machine performance have been discussed, along with the advantages that make SynRM an attractive solution for modern industrial applications, such as its simple construction, robustness, and absence of permanent magnets. In addition, an overview of the main modeling approaches and conventional control strategies has been provided, highlighting their effectiveness as well as their limitations, particularly under parameter variations and nonlinear operating conditions. This general analysis establishes a solid foundation for understanding the behavior of SynRM drives and justifies the need for advanced control techniques. The following chapter will focus on the detailed modeling and the development of an advanced control strategy for the SynRM, aiming to improve performance, robustness, and adaptability.

Chapter II: Modeling of Synchronous Reluctance Machine

II.1 Introduction

The modeling of any system is essential for the implementation of an appropriate control strategy. In this chapter, we focus on the modeling of the different components of the studied system, which consists of the SynRM, the power electronic converter, and the DC bus. In fact, the modeling of electrical machines generally involves complex mathematical equations. For the SynRM, this complexity arises mainly from the rotor's anisotropic structure, which creates different magnetic reluctances along the direct (d) and quadrature (q) axes. This magnetic saliency is the key factor in torque production but also makes the analytical model more challenging to establish. However, by adopting certain simplifying assumptions, such as sinusoidal distribution of windings, linear magnetic behavior, and neglecting saturation and losses, it becomes possible to derive a simplified and usable mathematical model.

II.2 Simplifying hypotheses

To develop the equivalent electrical model of the machine, certain assumptions need to be made. In the case of SynRM, the modelling relies on the following simplifying assumptions:

- The hysteresis in the magnetic parts as well as the iron losses are neglected.
- Assuming the magnetic circuit is unsaturated.
- Not considering notch and gap harmonics.
- Assuming the spatial distribution of the magneto-motive forces in the air gap is sinusoidal.
- Neglecting the effect of temperature on the resistance values [\[BOU23\]](#).

II.3 Electrical equations of the machine in the abc reference frame

The electrical equations governing the operation of a synchronous machine in a three-phase fixed reference frame (a,b,c) linked to the stator are written in the following form [\[MAT95\]](#).

$$[V_{abc}] = [R_s] [I_{abc}] + \frac{d}{dt} [\Psi_{abc}] \quad (\text{II.1})$$

$$[I_{abc}] = \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}; \quad [V_{abc}] = \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}; \quad [\Psi_{abc}] = \begin{bmatrix} \psi_a \\ \psi_b \\ \psi_c \end{bmatrix}; \quad [R_s] = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix} \quad (\text{II.2})$$

where:

- ψ_a , ψ_b , and ψ_c are the magnetic fluxes that link each stator winding.
- v_a , v_b , and v_c are the individual phase voltages across the stator windings.
- i_a , i_b , and i_c are the currents flowing in the stator windings.
- R_s is the equivalent resistance of each stator winding.

The total fluxes of the stator phases can be written in the stator reference frame in the following

matrix form:

$$[\Psi_{abc}] = [L][I_{abc}] \quad (\text{II.3})$$

where $[L]$ is the inductance matrix which depends on the angle θ defined in Figure 2.1:

$$[L] = \begin{bmatrix} L_a(\theta) & M_{ab}(\theta) & M_{ac}(\theta) \\ M_{ba}(\theta) & L_b(\theta) & M_{bc}(\theta) \\ M_{ca}(\theta) & M_{cb}(\theta) & L_c(\theta) \end{bmatrix} \quad (\text{II.4})$$

Assuming the first space harmonic, the inductance shown in Figure II.1 varies with the position of the rotor can be expressed as follows [\[DOP07\]](#):

$$\left| \begin{array}{l} L_a(\theta) = L_l + L' + L'' \cos(2\theta) \\ L_b(\theta) = L_l + L' + L'' \cos\left(2\theta + \frac{2\pi}{3}\right) \\ L_c(\theta) = L_l + L' + L'' \cos\left(2\theta - \frac{2\pi}{3}\right) \end{array} \right| \quad \left| \begin{array}{l} M_{ab}(\theta) = M_{ba} + M' + M'' \cos\left(2\theta - \frac{2\pi}{3}\right) \\ M_{ac}(\theta) = M_{ca} + M' + M'' \cos\left(2\theta + \frac{2\pi}{3}\right) \\ M_{bc}(\theta) = M_{cb} + M' + M'' \cos(2\theta) \end{array} \right|$$

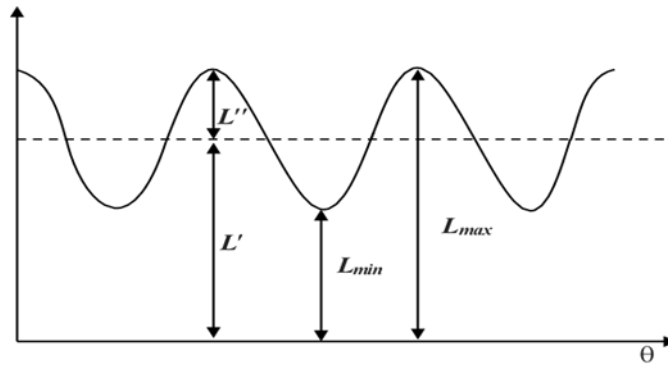


Figure II-1: Variation in inductance as a function of rotor position.

Where

L_l : leakage inductance

L' : Constant self-inductance term for one phase

L'' : Amplitude of the first harmonic of the natural inductance of a phase

$$M' = -\frac{1}{2} L' \quad , \quad M'' = L'' \quad [\text{CHA83}].$$

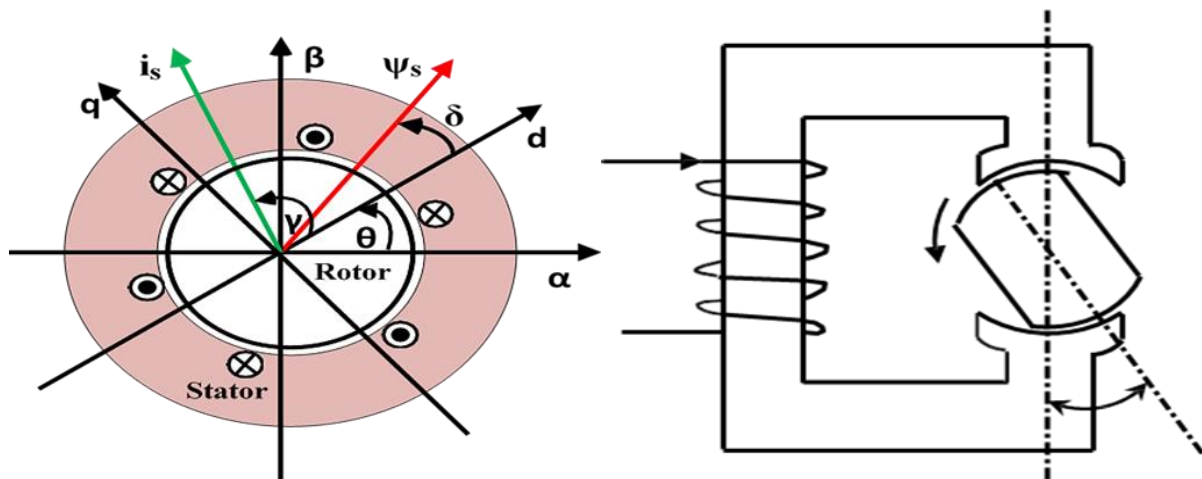


Figure II-2: (a) Representation of the synchronous machine in the d-q and α - β reference frames (b) Schematic diagram of a reluctance machine

II.4 Electrical equations of the machine in frame $\alpha\beta$

With a two-phase current system, produced by two coils offset by $\pi/2$ in space, we can create a rotating field identical to that created by any three-phase system. There are two transformations: the Clarke transformation and the Concordia transformation. The Clarke transformation preserves the amplitude of the quantities, but not the power or the torque (they must be multiplied by a coefficient $3/2$). On the other hand, the one of Concordia, which is normalized, keeps the power, but not the amplitudes. It is preferred to use the Clarke

transformation, because it facilitates the analysis of temporal signals and the adjustment of the limitations. One can obtain, in the general case [\[HAM10\]](#).

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = [c_{32}] \begin{bmatrix} x_\alpha \\ x_\beta \\ x_h \end{bmatrix} ; \begin{bmatrix} x_\alpha \\ x_\beta \\ x_h \end{bmatrix} = [c_{23}] \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (\text{II.5})$$

where:

$$[c_{32}] = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix}; \quad [c_{32}] = [c_{23}]^{-1} = \begin{bmatrix} 1 & -\frac{1}{2} & 1 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} & 1 \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} & 1 \end{bmatrix}$$

The system of equations (II.1) becomes:

$$[V_{\alpha\beta h}] = [R_s] [I_{\alpha\beta h}] + \frac{d}{dt} [\Psi_{\alpha\beta h}] \quad (\text{II.6})$$

where:

$$[V_{\alpha\beta h}] = \begin{bmatrix} v_\alpha \\ v_\beta \\ v_h \end{bmatrix}; [I_{\alpha\beta h}] = \begin{bmatrix} i_\alpha \\ i_\beta \\ i_h \end{bmatrix}; [\Psi_{\alpha\beta h}] = \begin{bmatrix} \psi_\alpha \\ \psi_\beta \\ \psi_h \end{bmatrix}$$

The flux matrix is given by:

$$[\Psi_{\alpha\beta h}] = [L_{\alpha\beta h}] [I_{\alpha\beta h}] \quad (\text{II.7})$$

with:

$$[L_{\alpha\beta h}] = \begin{bmatrix} L_l + \frac{2}{3}(L' + L'' \cos(2\theta)) & \frac{2}{3}L'' \sin(2\theta) & 0 \\ \frac{2}{3}L'' \sin(2\theta) & L_l + \frac{2}{3}(L' + L'' \cos(2\theta)) & 0 \\ 0 & 0 & L_l \end{bmatrix}$$

With the neutral of the machine isolated, we have $i_h = 0$ and we can write:

$$\begin{bmatrix} V_\alpha \\ V_\beta \end{bmatrix} = R_s \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \Psi_\alpha \\ \Psi_\beta \end{bmatrix} \quad (\text{II.8})$$

II.5 Electrical equations of the machine in frame dq

The park matrix is :

$$[P] = \frac{2}{3} \begin{bmatrix} \cos(\theta) & \cos\left(\theta - \frac{2\pi}{3}\right) & \cos\left(\theta + \frac{2\pi}{3}\right) \\ -\sin(\theta) & -\sin\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \quad (\text{II.9})$$

Calculating the inverse of $[P]$ is straightforward:

$$[P]^{-1} = \begin{bmatrix} \cos(\theta) & -\sin(\theta) & 1 \\ \cos\left(\theta - \frac{2\pi}{3}\right) & -\sin\left(\theta - \frac{2\pi}{3}\right) & 1 \\ \cos\left(\theta + \frac{2\pi}{3}\right) & -\sin\left(\theta + \frac{2\pi}{3}\right) & 1 \end{bmatrix} \quad (\text{II.10})$$

If we project all the quantities into the d-q reference frame linked to the rotor Figure II.2 using Park's transformation, we write, in the general case:

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = [P]^{-1} \begin{bmatrix} x_d \\ x_q \\ x_h \end{bmatrix} \quad (\text{II.11})$$

Equation (2.1) then becomes:

$$[P]^{-1} [V_{dqh}] = [R_s] [P]^{-1} [I_{dqh}] + [P]^{-1} \frac{d[\Psi_{dqh}]}{dt} + \frac{d[P]^{-1}}{dt} [\Psi_{dqh}] \quad (\text{II.12})$$

Where $[X_{dqh}]$ denotes any vector of quantities expressed in the reference frame linked to the rotor. Multiplying the two members of (1.12) by $[P]$ gives:

$$[V_{dqh}] = [R_s] [I_{dqh}] + \frac{d[\Psi_{dqh}]}{dt} + p\Omega [P] \frac{d[P]^{-1}}{d\theta} [\Psi_{dqh}] \quad (\text{II.13})$$

With:

$$[P] \frac{d[P]^{-1}}{d\theta} = \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}$$

We finally arrive at the following equations:

$$[V_{dqh}] = [R_s] [I_{dqh}] + \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_h \end{bmatrix} \frac{d[I_{dqh}]}{dt} + p\Omega \begin{bmatrix} 0 & -L_q & 0 \\ L_d & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} [I_{dqh}] \quad (\text{II.14})$$

$$\text{With : } \begin{cases} L_d = L_l + \frac{3}{2}(L' + L'') \\ L_q = L_l + \frac{3}{2}(L' - L'') \\ L_h = L_l \end{cases}$$

Since the neutral of the machine is isolated, which naturally implies $i_h = 0$, we can write:

$$\begin{bmatrix} V_d \\ V_q \end{bmatrix} = R_s \begin{bmatrix} R_s & -p\Omega L_q \\ p\Omega L_d & R_s \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (\text{II.15})$$

p: Number of pole pairs

II.6 Vector equivalent circuit and main performance characteristics:

In the case where it is necessary to take into account the presence and influence of rotor currents. In his thesis [PAR07] and in his publications [PAR08], J.D.Park uses the model illustrated in figure II.3 to model a solid rotor SynRM. This model is similar to that of a MAS where the slip is always equal to unity.

Due to the sinusoidal distribution of the stator winding in a SynRM, the flux harmonics in the air gap add only an extra term to the stator leakage inductance. As a result, the SynRM's behaviour can be described using the standard equations for a wound-field synchronous machine, specifically the Park equations. However, unlike conventional machines, the SynRM lacks an excitation (field) winding, and the rotor does not have a cage. Additionally, the SynRM can be started synchronously from a standstill using suitable inverter control [LIO94].

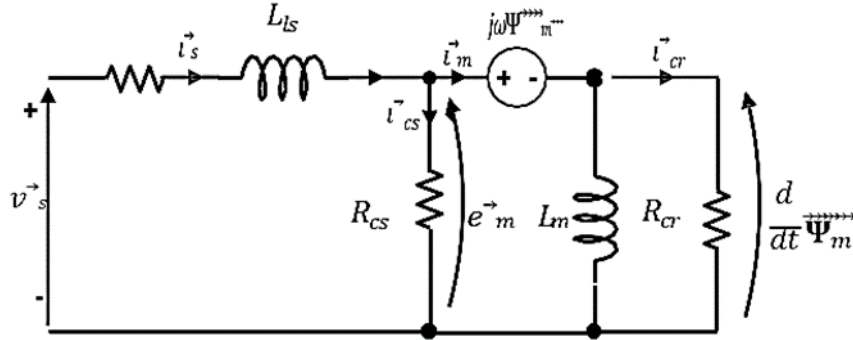


Figure II-3: SynRM equivalent vector circuit including rotor and stator iron losses.

Where:

L_{ls} : Total winding leakage inductance.

Ψ_m : Air gap linkage flux.

ω : Reference frame electrical angular speed.

R_s : Winding resistance.

R_{cs} : Stator iron losses resistance

R_{cr} : Rotor iron losses resistance

i_m : Magnetization current

e_m : Air gap electromotive voltage (internal voltage of stator winding).

II.7 Electromagnetic Torque

In synchronous reluctance machine the electromagnetic torque is produced by the interaction between air gap flux and its corresponding magnetizing current.

$$T_e = \frac{P_e}{\Omega} = P(\Psi_d i_q - \Psi_q i_d) = P(L_d - L_q) i_d i_q \quad (\text{II.16})$$

The mechanical speed of rotation can be deduced from Newton's fundamental law of motion:

$$J \frac{d\Omega}{dt} + F\Omega = T_e - T_L \quad (\text{II.17})$$

Where :

$$\frac{d\theta_e}{dt} = \omega, \quad \omega = P\Omega$$

II.8 Non-linear state model

In general, to present a state model, we need to define the state vector x , the input vector u and the output vector y . The input vector consists of the stator voltages. The state vector is made up of electrical quantities (currents) and mechanical quantities (speed and/or position).

[\[AGH23\]](#).

II.8.1 State model in rotating frame (d - q)

In the case of torque or angular speed control, the non-linear state model state model in the rotating reference frame d-q is described by the system below:

$$\begin{bmatrix} \dot{i}_d \\ \dot{i}_q \\ \dot{x}_c \end{bmatrix} = \begin{bmatrix} -\frac{R}{L_d} + \frac{pL_q}{L_d} i_q \Omega \\ -\frac{R}{L_q} + \frac{pL_d}{L_q} i_d \Omega \\ \frac{3p(L_d-L_q)}{2J} i_d i_q - \frac{F}{J} \Omega \end{bmatrix} + \begin{bmatrix} \frac{1}{L_d} & 0 & 0 \\ 0 & \frac{1}{L_q} & 0 \\ 0 & 0 & -\frac{1}{J} \end{bmatrix} \begin{bmatrix} v_d \\ v_q \\ T_L \end{bmatrix} \quad (\text{II.18})$$

In the case of controlling the rotor position θ , this will be considered as a new state variable and the new state model is therefore written:

$$\begin{bmatrix} \dot{i}_d \\ \dot{i}_q \\ \dot{\Omega} \\ \dot{\theta} \end{bmatrix} = \begin{bmatrix} -\frac{R}{L_d} + \frac{pL_q}{L_d} i_q \Omega \\ -\frac{R}{L_q} + \frac{pL_d}{L_q} i_d \Omega \\ \frac{3p(L_d-L_q)}{2J} i_d i_q - \frac{F}{J} \Omega \\ \Omega \end{bmatrix} + \begin{bmatrix} \frac{1}{L_d} & 0 & 0 \\ 0 & \frac{1}{L_q} & 0 \\ 0 & 0 & -\frac{1}{J} \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} u_d \\ u_q \\ T_r \end{bmatrix} \quad (\text{II.19})$$

Figure2-4 presents the SynRM model in synchronous reference frame with inputs stator voltages, and load torque. The outputs are the rotor angular velocity.

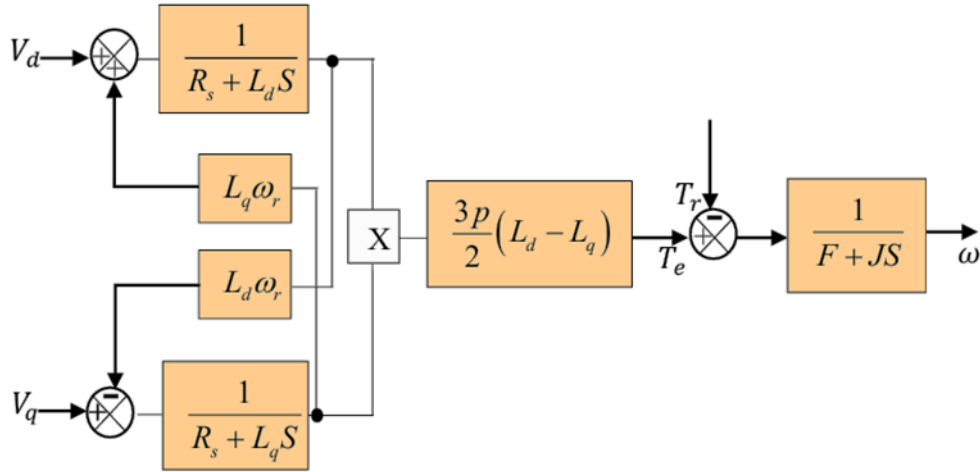


Figure II-4: Block diagram of the SynRM system.

II.9 Modeling the synchronous machine power supply

Direct current to alternating current converters, more commonly known as inverters, depending on the type of power source and the topology of the power circuit, are classified as voltage source inverters (VSI) and current source inverters (CSI). The main purpose of these topologies is to provide a three-phase voltage source, where the amplitude, phase and frequency of the voltages can be controlled. The two-level voltage inverter is made up of three arms formed by electronic switches chosen essentially according to the power and frequency of the system to be controlled, each arm comprising two IGBT power electronic components equipped with diodes mounted in anti-parallel. Figure II-5 shows a power supply diagram for a SynRM with a voltage inverter.

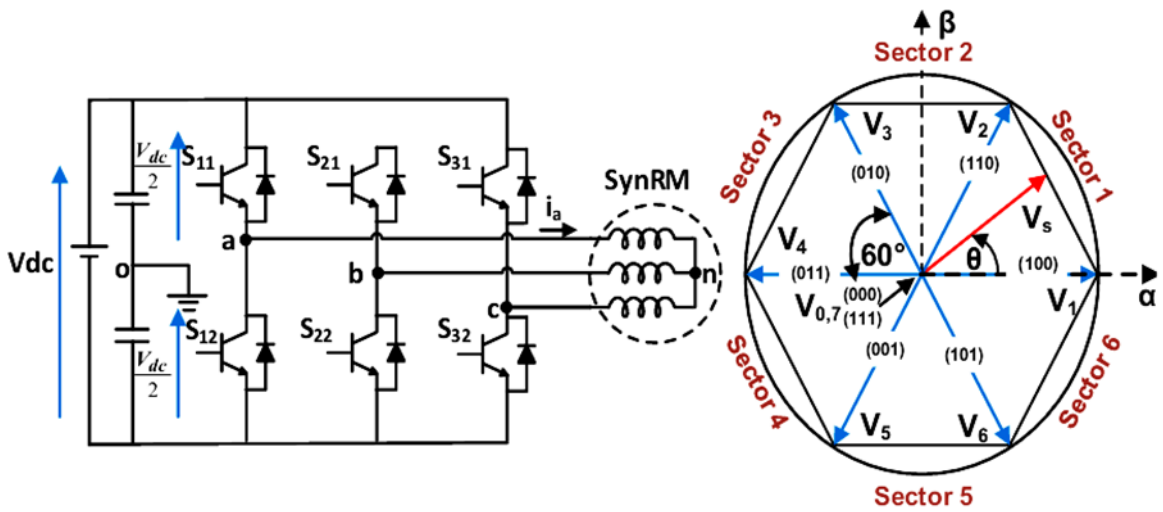


Figure II-5: Two level voltage source inverter.

S_{11}	S_{21}	S_{31}	Vector Voltage
0	0	0	$V_0=0$
1	0	0	$V = \frac{2}{3} V_{dc}$
1	1	0	$V = \frac{1}{3} V_{dc} + j \frac{\sqrt{3}}{3} V_{dc}$
0	1	0	$V = -\frac{1}{3} V_{dc} + j \frac{\sqrt{3}}{3} V_{dc}$
0	1	1	$V = -\frac{2}{3} V_{dc}$
0	0	1	$V = -\frac{1}{3} V_{dc} - j \frac{\sqrt{3}}{3} V_{dc}$
1	0	1	$V = \frac{1}{3} V_{dc} - j \frac{\sqrt{3}}{3} V_{dc}$
1	1	1	$V_7=0$

Table II-1: Switching States and Voltage Vector.

The inverter has eight switching states as shown in Table II-1. The two switches in the same branch cannot be activated at the same time, as this would short-circuit the input voltage. Therefore, the nature of the two switches in the same leg is complementary

$$\begin{aligned}
 S_{11} + S_{12} &= 1 \\
 S_{21} + S_{22} &= 1 \\
 S_{31} + S_{32} &= 1
 \end{aligned} \tag{II.20}$$

The modulation signals are thus selected so meet some specifications, like harmonic elimination, higher fundamental component and so on. The phase voltages can be obtained from the line voltages as

$$\begin{bmatrix} V_{ab} \\ V_{bc} \\ V_{ca} \end{bmatrix} = \begin{bmatrix} V_{an} - V_{bn} \\ V_{bn} - V_{cn} \\ V_{cn} - V_{an} \end{bmatrix} \tag{II.21}$$

As we are dealing with balanced voltages $V_{an} + V_{bn} + V_{cn} = 0$, we therefore have:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} V_{ab} - V_{cn} \\ V_{bc} - V_{an} \\ V_{ca} - V_{bn} \end{bmatrix} \tag{II.22}$$

The voltages between the phases can also be written in the following form:

$$\begin{bmatrix} V_{ab} \\ V_{bc} \\ V_{ca} \end{bmatrix} = \begin{bmatrix} V_{ao} - V_{bo} \\ V_{bo} - V_{co} \\ V_{co} - V_{ao} \end{bmatrix} \tag{II.23}$$

Substituting II.22 into II.23 we get:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} V_{ao} \\ V_{bo} \\ V_{co} \end{bmatrix} \quad (\text{II.24})$$

And note that:

$$\begin{bmatrix} V_{ao} \\ V_{bo} \\ V_{co} \end{bmatrix} = V_{dc} \begin{bmatrix} S_{11} - \frac{1}{2} \\ S_{21} - \frac{1}{2} \\ S_{31} - \frac{1}{2} \end{bmatrix} \quad (\text{II.25})$$

Therefore:

$$\begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_{11} \\ S_{21} \\ S_{31} \end{bmatrix} \quad (\text{II.26})$$

The three-phase tensions above can be controlled based on the rotation of the spatial vector, as illustrated in Figure II-6. One can deduce from this figure that the three phase tensions are alternately represented by a rotating vector V_s . To do this, we use a fixed 2-D coordinate system, in which the horizontal α -axis is aligned in the same direction as phase A, and β is the vertical axis as usual [DOM14]. V_s is now equivalent to the voltage vectors V_α and V_β

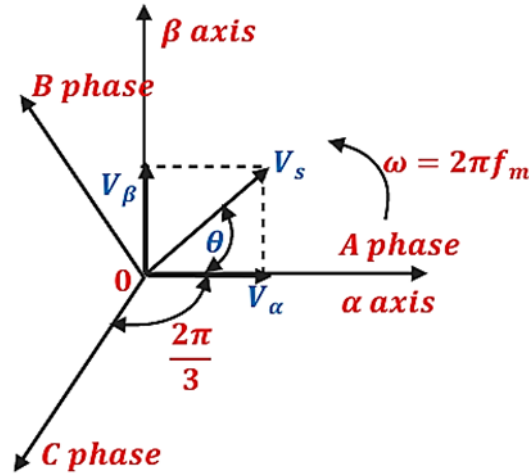


Figure II-6: Relationship between three-phase voltage and coordinates $\alpha \beta$.

It is obvious that the three-phase voltage is completely controlled by V_α and V_β , these voltage vectors are described by a set of formulae as below

$$\begin{bmatrix} V_\alpha \\ V_\beta \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} V_{an} \\ V_{bn} \\ V_{cn} \end{bmatrix} \quad (\text{II.27})$$

The matrix in the above equation is also expressed as:

$$\begin{aligned} V_\alpha &= V_{an} \\ V_\beta &= \frac{(2V_{bn} + V_{an})}{\sqrt{3}} \end{aligned} \quad (\text{II.28})$$

$$V_s = \sqrt{V_\alpha^2 + V_\beta^2} \quad (\text{II.29})$$

$$\theta = \arctan\left(\frac{V_\beta}{V_\alpha}\right) \quad (\text{II.30})$$

$$\omega = 2\pi f_m \quad (\text{II.31})$$

II.10 Simulation results:

To complete the theoretical study presented above, a numerical simulation is essential. The algorithm is tested in the MATLAB environment. The SynRM parameters are shown in Table II.2

Variable	Symbol	Value
d-axis inductance	L_d	0.1524 [H]
q-axis inductance	L_q	0.0345 [H]
Number of poles	p	2
Friction coefficient	f	0.00015 [N.m ⁻¹]
Moment of inertia	J	0.00044 [Kg.m ²]
Stator resistance	R_s	8.1 [Ω]
d-time constant	$T_d = L_d/R_s$	0.01881 [s]
q-time constant	$T_q = L_q/R_s$	0.00426 [s]

Table II.2: Parameters of the machine used [MIL25].

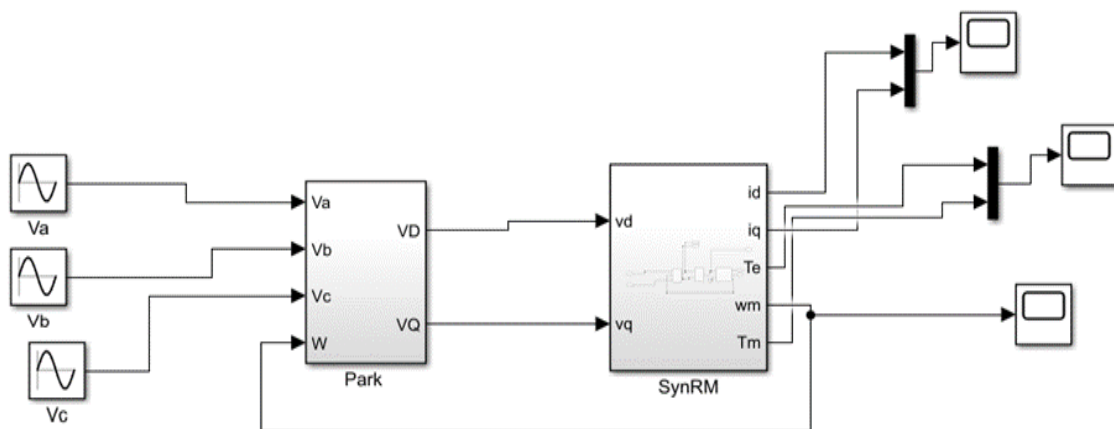


Figure II.7: SynRM simulation diagram

II.10.1 SynRM is fed directly from the public network.

SynRM is supplied by purely sinusoidal and balanced sources, expressed as follows

$$\begin{cases} V_a = V_m \sin(\omega t) \\ V_b = V_m \sin(\omega t - \frac{2\pi}{3}) \\ V_c = V_m \sin(\omega t + \frac{2\pi}{3}) \end{cases}$$

The following figures show the evolution of SynRM's characteristics fed directly from a balanced purely sinusoidal source, followed by the application of a mechanical load during the time interval [1.5s,3s].

At start-up, at no load, Figure II.8 illustrates the evolution of the SynRM's rotor speed, with ripples damping off at 0.25 s heralding the onset of steady state and reaching a synchronous value of 314.16 rad/s.

Figure 1.9 shows the evolution of the electromagnetic torque. Initially, the torque of the SynRM reaches the value of -8.6 N.m, then +5.5 N.m and exhibits oscillations which disappear after 0.1s when it reaches 3.46 N.m, then it decreases in an almost linear fashion and stabilises at its minimum value.

Figure II.10 shows the evolution of the stator currents along the direct and quadrature axes in a manner more or less analogous to the evolution of the speed, while slight oscillations can be seen, which disappear after 0.1s.

By applying a load torque equal to $T_L = 2\text{N.m}$ at time $t = 1.5\text{s}$, it can be seen that after small oscillations the speed returns to the synchronous speed of 314.16 rad/s , the current I_q decreases slightly from 7.92A to 7.83A , while the current I_d increases and stabilises at 3.84A.

The torque reaches a value of 3.5N.m and then stabilises at 2. 02N.m (slightly higher than the load torque), which explains the coupling between the direct axis and the quadratic axis.

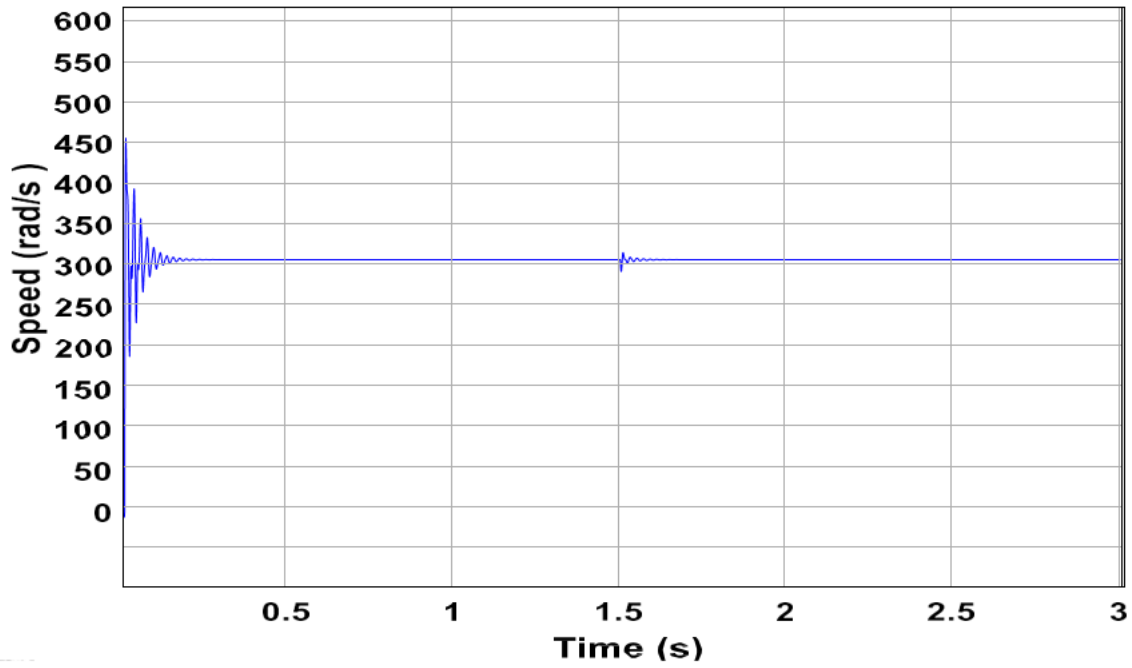


Figure II.8: Rotational speed with no load and then under load at $t = 1.5\text{s}$, the machine supplied by the electrical network.

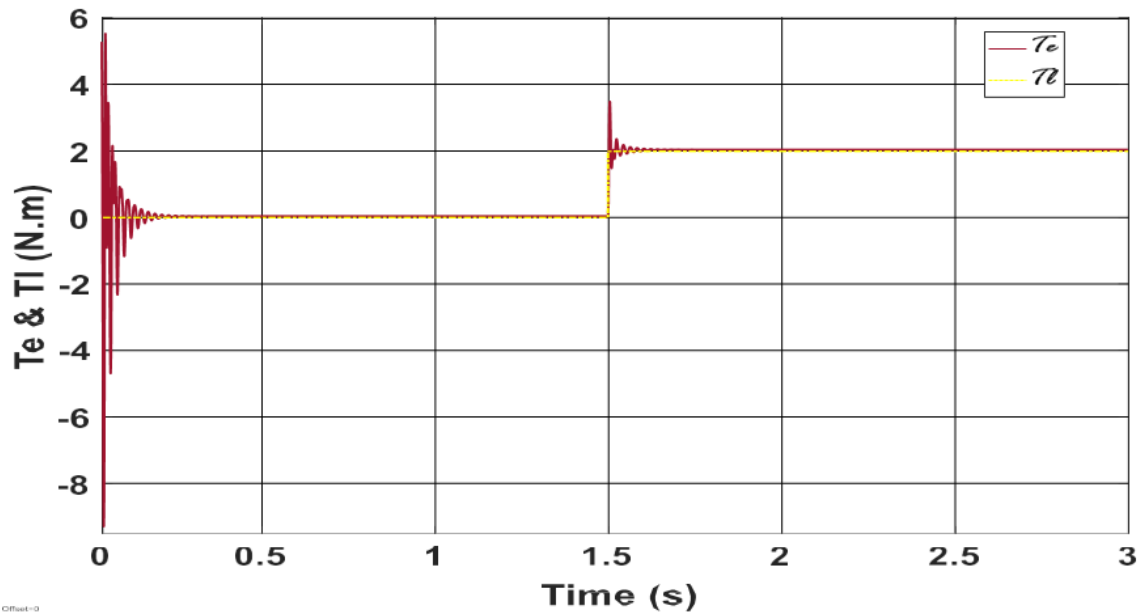


Figure II.9: Electromagnetic torque response of the machine supplied by the electrical network.

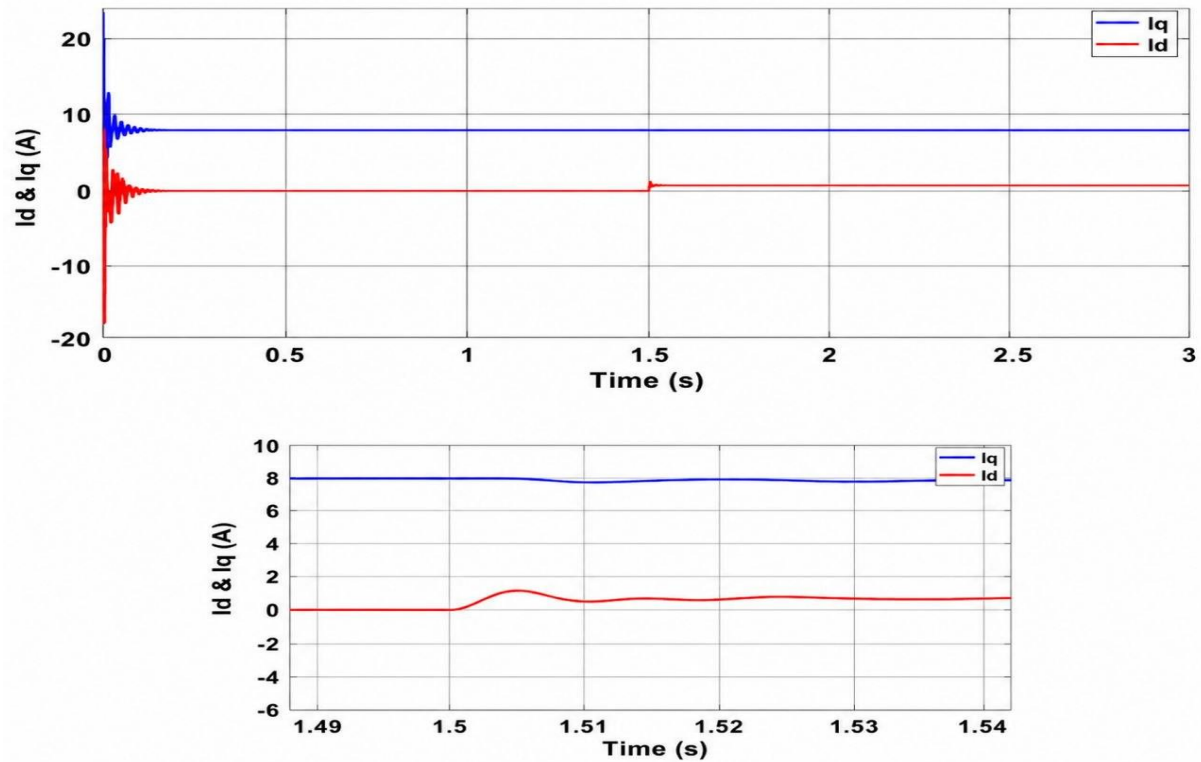


Figure II.10: Components of the stator currents along the d and q axes of the machine supplied by the electrical network.

II.10.2 SynRM is supplied by an inverter

Figure II.11, II.12, II.13, illustrate the progression of SynRM characteristics under the influence of PWM in the form of an S-Function. The study includes the introduction of a load $T_l=2\text{N.m}$ within the time interval $[4.5\text{s}, 9\text{s}]$. The outcomes in this scenario closely resemble those achieved with a purely sinusoidal power source. Nevertheless, a closer examination of the figures, particularly those representing electromagnetic torque, stator currents, and currents in the two direct axes and in quadrature, reveals that this power supply approach induces an elevation in ripples. This increase is primarily attributed to the harmonics introduced by the inverters and PWM, exerting a notable impact on the electromagnetic torque

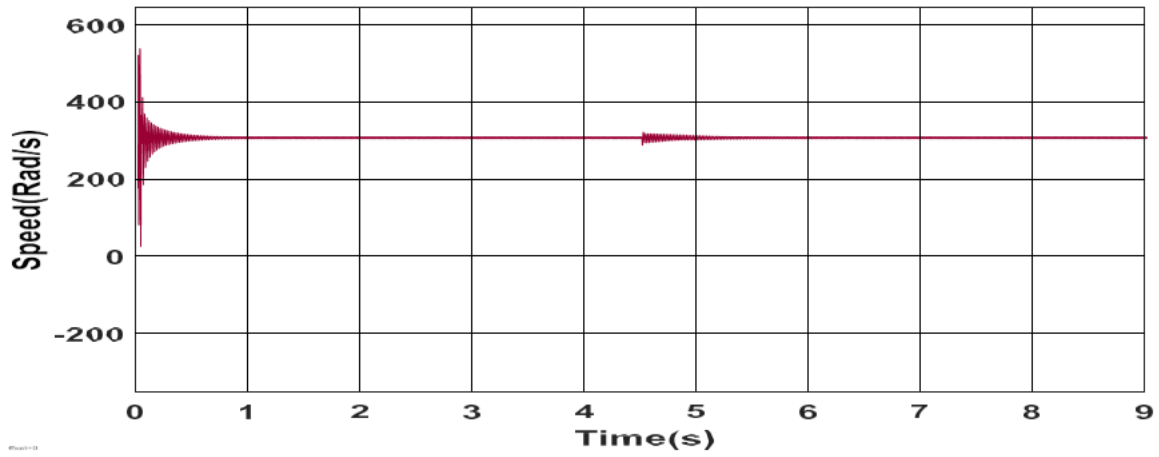


Figure II.11: Rotation speed at no load and under load at $t=4.5$, the machine powered by the inverter.

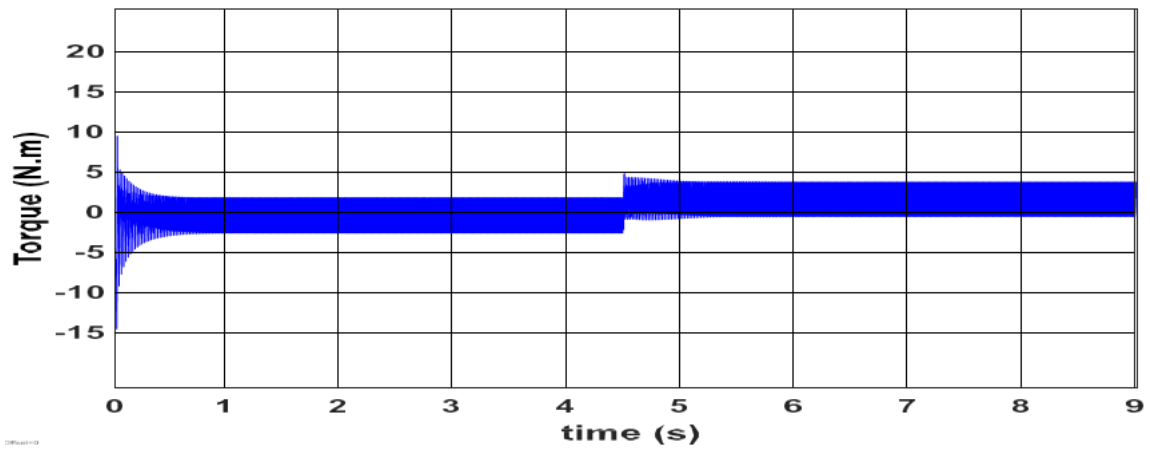


Figure II.12: Electromagnetic torque response of the machine powered by the inverter.

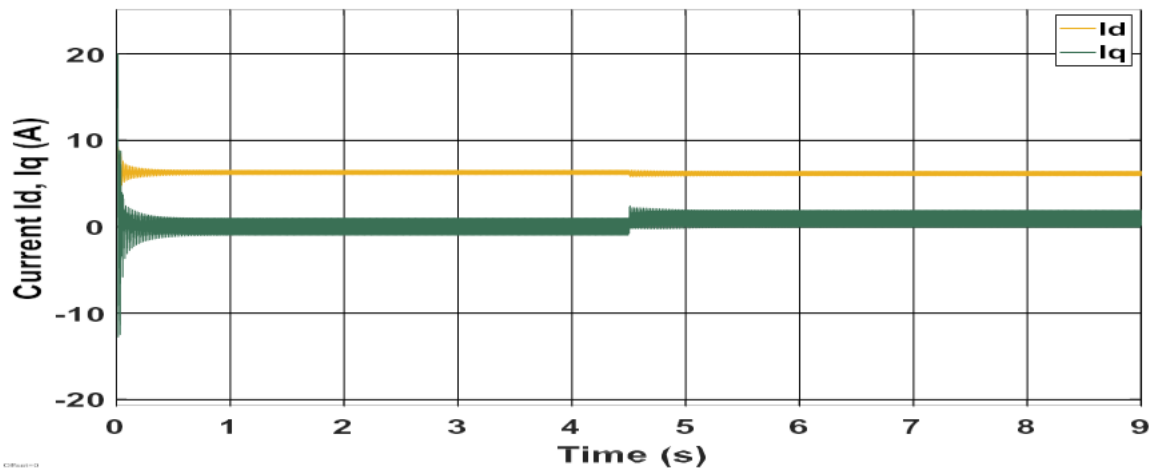


Figure II.13: Composants du courants statoriques selon les axes et la machine alimentée par l'onduleur.

II.11 Conclusion

This chapter delves into the characteristics and functionality of the SynRM, highlighting the significance of the L_d/L_q parameter in its design and optimization. A mathematical representation of the SynRM in the stationary reference frame (α - β) has been developed, while its behaviour in the synchronous reference frame (d-q) has also been analysed. This modelling is essential for applying advanced control strategies such as vector control. These representations provide a deeper insight into the electrical and magnetic properties of the machine and support the implementation of rotor flux orientation and speed regulation techniques. Furthermore, the inverter, a key element for supplying power to the SynRM, has been modelled.

***Chapter III: Model-free
Sliding Mode Control of The
SynRM Machine***

III.1 Introduction

Artificial intelligence techniques have become increasingly important in modern industrial applications due to their ability to address complex control and identification problems. These techniques provide efficient solutions for systems affected by parameter uncertainties, nonlinearities, and external disturbances. Among the most widely used artificial intelligence methods are artificial neural networks, which have demonstrated significant capabilities in modeling, estimation, and control of electrical drive systems. In parallel, Sliding Mode Control (SMC) has emerged as one of the most effective nonlinear control strategies because of its robustness against parameter variations and external perturbations. This control approach ensures high dynamic performance and stability, making it particularly suitable for electric machine drive applications.

This chapter presents the fundamental concepts of Sliding Mode Control, including its operating principles, convergence conditions, design methodology, advantages, and limitations. Subsequently, the application of SMC to the control of the Synchronous Reluctance Motor (SynRM) is developed through the derivation of the complete control law and stability analysis. Furthermore, artificial neural network techniques, particularly Recurrent Neural Networks (RNNs), are introduced and integrated within a model-free control framework for the SynRM. Finally, simulation results are presented to evaluate the effectiveness of the proposed control strategy and to demonstrate its robustness and performance under parameter variations and different operating conditions.

III.2 Theory of Sliding Mode Control:

Variable structure sliding mode control has been around since the early 1960s. It was studied and developed exclusively in the Soviet Union in the 1960s. Subsequently, numerous research studies have been conducted elsewhere, either to complement the theoretical study or to apply it to physical systems. The CSV, which by its nature is a nonlinear control. It is based on the switching of state variable functions, used to create a sliding manifold or hypersurface, whose purpose is to force the system's dynamics to match those defined by the hypersurface equation. When the state is maintained on this hypersurface, the system is in sliding mode. Its dynamics are then insensitive to external and parametric disturbances as long as the conditions of the sliding regime are met. The trajectory in the phase plane consists of three distinct parts, Figure III-1. [\[RSA10-FBE03\]](#).

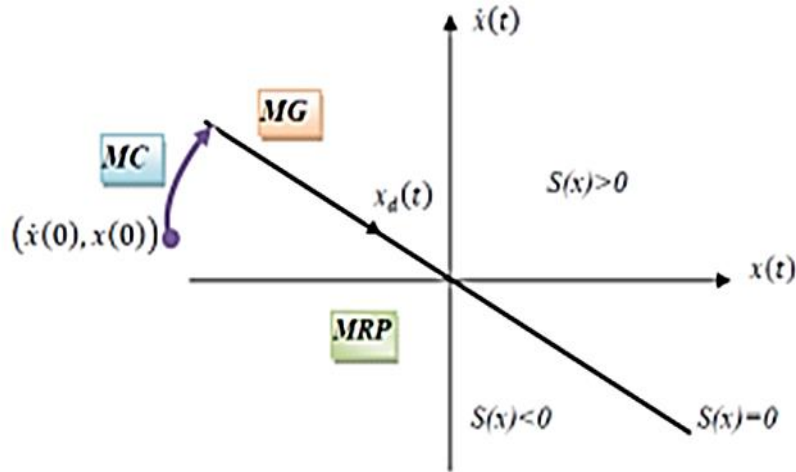
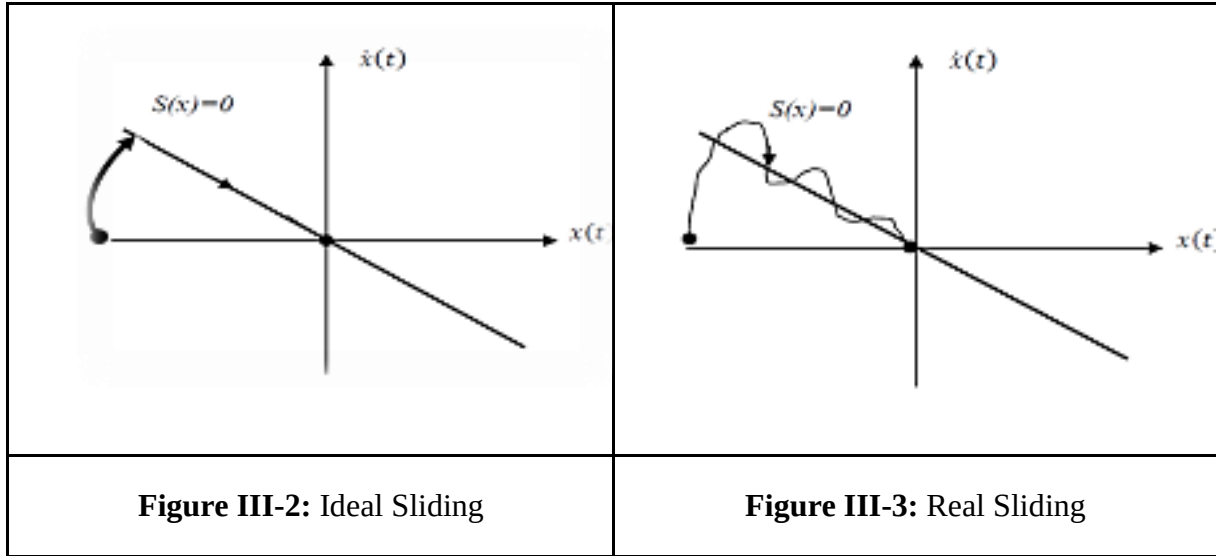


Figure III-1: Convergence of the sliding system.

- ❖ **Reaching Mode (RM)** : In this initial phase, the controlled variable (e.g., motor speed) moves from any initial state toward the switching surface defined by $S(x) = 0$, within a finite time. The design of the control law and the application of a convergence criterion (such as Lyapunov-based stability conditions) ensure this transition. For SynRM systems, this mode is essential for quick transient response during start-up or command changes [\[CTL18\]](#).
- ❖ **The sliding mode (SM)**: Once the system state reaches the sliding surface, it enters the sliding mode. Here, the system's behavior is governed only by the dynamics on the sliding surface, which is immune to matched disturbances and parameter variations. For the SynRM, this allows for stable and robust speed control, even in the presence of disturbances such as mechanical load fluctuations or magnetic saturation [\[CTL18\]](#).
- ❖ **The steady-state mode (SSM)**: This mode examines the system's behavior around the equilibrium point, typically at the origin of the phase plane. It characterizes the system's steady-state accuracy and performance, which is critical in nonlinear systems. For SynRM drives, a well-behaved steady-state mode ensures precise and stable operation, which is vital for high-performance industrial applications requiring speed and torque consistency [\[CTL18\]](#).

Ideal Sliding Regime: In theory, the switching element is supposed to be insensitive to noise, and the trajectory in the sliding regime perfectly describes the equation $S(x)=0$. The ideal sliding mode corresponds to an oscillation of infinite frequency and zero amplitude, the representative point of the system's evolution perfectly slides on the switching hyper surface Figure III-2.

Real Sliding Mode: In practice, the switching element is made from relays that exhibit imperfections such as switching delays. In this case, the phase trajectory of the sliding mode remains close to the switching surface, giving rise to undesirable oscillations that eliminate the system's precision and nonetheless its stability Figure III-3.



III.3 Sliding mode existence condition

The second important aspect, after the design of the switching surface, is to ensure the existence of the sliding mode, hence the condition for obtaining the sliding regime:

$$S(x) \cdot \dot{S}(x) < 0 \quad (\text{III.1})$$

The existence of a sliding mode requires the stability of the state trajectory at the surface of slip $S(x,t)=0$.

The movement along the line, which no longer represents the trajectory of both structures, screams slip mode. The switching phenomenon along this slip line can be demonstrated using of the following Figure III-4[FEN13]:

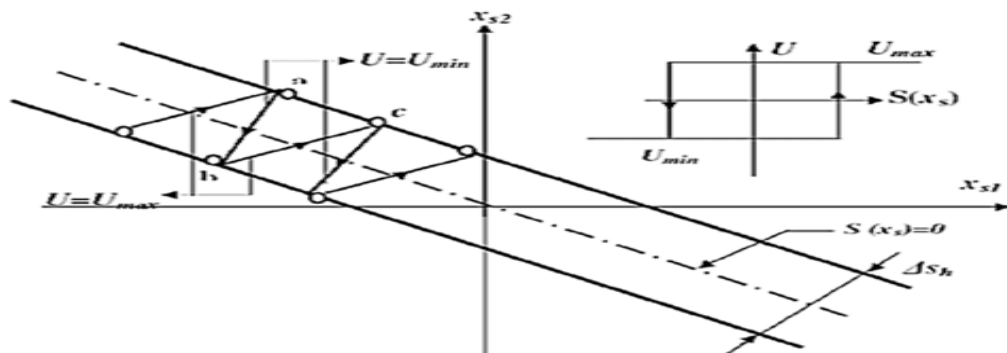


Figure III-4: Slide mode proof.

III.4 Design of the sliding mode control algorithm

The design of regulators using sliding modes systematically addresses stability issues and desired performance. The implementation of this control method mainly requires three steps:

- The choice of the surface,
- The establishment of the conditions for the existence of convergence,
- The determination of the control law.

III.4.1 choice of the sliding surface

The choice of the sliding surface concerns the necessary number of its surfaces and also their shape, depending on the application and the intended objective. For a system defined by:

With:

$$\dot{\mathbf{x}}(t) = \mathbf{f}(\mathbf{x}, t) + \mathbf{b}(\mathbf{x}, t)\mathbf{U}(t) \quad (\text{III.2})$$

$\mathbf{x}(t)$ State vector, $\mathbf{U}(t)$ input vector, and $\mathbf{f}(\mathbf{x}, t)$ and $\mathbf{b}(\mathbf{x}, t)$ nonlinear functions. A number of m sliding surfaces for a vector $\mathbf{U}$ of dimension m . With regard to the shape of the surface, two possibilities arise, either in the state space or in the phase plane [\[LNE99\]](#).

In the case of phase-space processing, the switching function is a scalar function, such that the variable to be controlled slides along this surface to reach the origin of the phase plane. Thus, the surface $S(\mathbf{x})$ represents the desired dynamic behavior of the system.

J. J. Slotine proposes a general form of equation for determining the slip surface that ensures the convergence of a variable to its setpoint value [\[LNE99- AEK06- AAO11\]](#) The equation is given by:

$$S(t) = \left(\frac{\partial}{\partial x} + \lambda_x \right)^{r-1} e(x) \quad (\text{III.3})$$

Where:

(x) : The error of the variable to regulate $e(x) = x_{ref} - x$.

λ_x : A positive constant that interprets the bandwidth of the desired control.

r : Relative degree, representing the number of times the surface must be differentiated to reveal the command.

For: $r = 1$, $S(x) = e(x)$

$r = 2$, $S(x) = \lambda_x e(x) + \dot{e}(x)$

$r = 3$, $S(x) = \lambda_x^2 e(x) + 2\lambda_x \dot{e}(x) + \ddot{e}(x)$.

$S(x) = 0$ is a linear differential equation whose unique solution is $e(x) = 0$. It tends to zero for a correct choice of the gain λ_x .

The trajectory tracking problem, whose objective is to keep $S(x) = 0$, is equivalent to an exact linearization of the error while respecting the convergence condition.

III.4.2 Conditions for the existence of convergence

The convergence conditions are situated within the framework of allowing the dynamics of the system to converge toward sliding surfaces regardless of disturbance. There are two conditions [BBE02-LEV93].

III.4.2.1 Direct switching function

It is studied by Emilyanov and Utkin. It aims to endow the surface with a convergent dynamic toward zero. It is given by:

$$S(x) \cdot \dot{S}(x) < 0 \quad (\text{III.4})$$

III.4.2.2 Lyapunov function

This approach is used to estimate the performance of the control, the study of the robustness and stability of nonlinear systems. It is a matter of choosing a positive scalar function $V(x) > 0$ for the state variables of the system and choosing a control law that will cause this function to decay $\dot{V}(x) < 0$.

For a Lyapunov candidate function of the form:

$$V(x) = \frac{1}{2} S^2(x) \quad (\text{III.5})$$

The drift of this function is:

$$\dot{V}(x) = S(x) \cdot \dot{S}(x) \quad (\text{III.6})$$

For the Lyapunov function to decrease, it suffices to ensure that its derivative is negative. This is verified if:

$$\dot{V}(x) = S(x) \cdot \dot{S}(x) < 0 \quad (\text{III.7})$$

This equation shows that the square of the distance to the surface, measured by $S^2(x)$, decreases all the time, forcing the system trajectory to move toward the surface of the two-sided figure III-5. This condition assumes an ideal sliding regime.

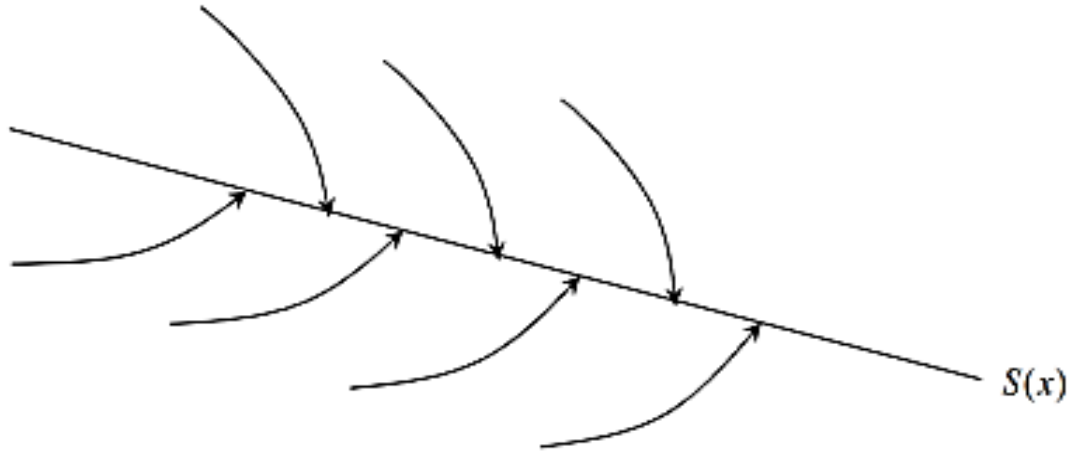


Figure III-5: Trajectory of the state with respect to the surface.

III.4.3 Determination of the control law:

If the sliding surface is chosen, as well as the convergence criterion, then it remains to determine the necessary control to bring the variable to be controlled back to the surface and then toward its equilibrium point while maintaining the existence condition of sliding modes.

The structure of a sliding mode controller consists of two parts, one concerning the exact linearization U_{eq} and the other, stability U_n . [FBE05]. The latter is very important in sliding mode control, as it is used to eliminate the effects of model inaccuracy and to reject external disturbances. $U(t) = U_{eq} + U_n$

U_{eq} Corresponds to the equivalent command proposed by Filippov and Utkin; it serves to maintain the variable to be controlled on the sliding surface (x). The equivalent is deduced by considering that the derivative of the surface is zero. $\dot{S}(x) = 0$

The derivative of the surface is:

$$\dot{S}(x) = \frac{dS}{dt} = \frac{\partial S}{\partial x} \cdot \frac{\partial x}{\partial t} = \frac{\partial S}{\partial x} [f(x, t) + B(x, t)U_{eq}(t)] + \frac{\partial S}{\partial x} [B(x, t)U_n] \quad (III.8)$$

During the sliding mode and the steady state, the surface is zero and consequently, its derivative and the discontinuous part are also zero. Hence, we deduce the expression of the equivalent command.

$$U_{eq} = \left[\frac{\partial S}{\partial x} B(x, t) \right]^{r-1} \left[\frac{\partial S}{\partial x} f(x, t) \right] \quad U_n = 0 \quad (III.9)$$

With the existence condition:

$$\left[\frac{\partial S}{\partial x} B(x, t) \right]^{r-1} \neq 0 \quad (III.10)$$

U_n : Is a term introduced to satisfy the convergence condition $S(x) \cdot \dot{S}(x) < 0$ During the convergence mode and replacing the equivalent command by its expression in (3.7) we obtain the new expression for the derivative of the surface:

$$\dot{S}(x) = \frac{\partial S}{\partial x} [B(x, t) U_n] \quad (\text{III.11})$$

The problem comes down to finding

$$S(x) \cdot \dot{S}(x) = S(x) \frac{\partial S}{\partial x} [B(x, t) U_n] < 0 \quad (\text{III.12})$$

The simplest solution is to choose U_n in the form of a relay Figure 3-6 In this case, the command is written as follows:

$$U_n = K \cdot \text{sign}(S(x)) = K \cdot \frac{S(x)}{|S(x)|} \quad (\text{III.13})$$

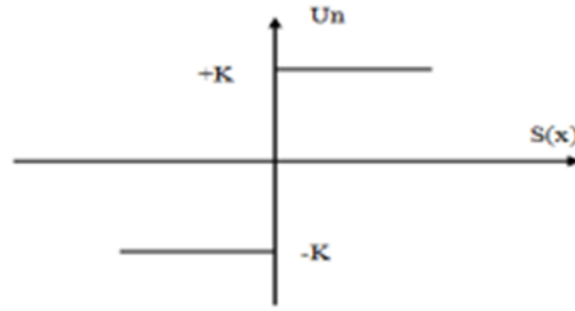


Figure III-6: Sign function.

By replacing the expression (3.13) in (3.12), we obtain:

$$S(x) \cdot \dot{S}(x) = S(x) \frac{\partial S}{\partial x} B(x, t) K \cdot \frac{S(x)}{|S(x)|} < 0 \quad (\text{III.14})$$

Where the factor $\frac{\partial S}{\partial x} B(x, t)$ is always negative and the gain K is chosen positive to satisfy the condition (III.14).

III.5 Application of Sliding Mode Control Modeling for a SynRM Machine

III.5.1 SynRM dynamic model

Considering the SynRM model of the d-q frame presented in the chapter 2 by expressions (II.18) and (II.19), we find that:

$$\frac{di_d}{dt} = \frac{1}{L_d} (v_d - R_s i_d - \omega_e L_q i_q) \quad (\text{III.15})$$

$$\frac{di_q}{dt} = \frac{1}{L_q} (v_q - R_s i_q - \omega_e L_d i_d) \quad (\text{III.16})$$

$$T_e = \frac{3}{2} P (L_d - L_q) i_d i_q \quad (\text{III.17})$$

$$J \frac{d\omega_m}{dt} + F\Omega = T_e - T_L - B\omega_m \quad (\text{III.18})$$

$$\omega_e = p\omega_m \quad (\text{III.19})$$

Define:

$$K_T = \frac{3}{2} P (L_d - L_q) \quad (\text{III.20})$$

Then:

$$T_e = K_T i_d i_q \quad (\text{III.21})$$

The mechanical equation becomes:

$$\frac{d\omega_m}{dt} = \frac{1}{J} (K_T i_d i_q - T_L - B\omega_m) \quad (\text{III.22})$$

III.5.2 Control objective

The aim of the SMC is to force the SynRM to track:

$$i_d \rightarrow i_d^*$$

$$i_q \rightarrow i_q^*$$

$$\omega_m \rightarrow \omega_m^*$$

where:

i_d^*, i_q^* are the reference currents, and: ω_m^* is the reference mechanical speed. Usually, the control structure is cascade control:

$$\text{Speed SMC} \rightarrow i_q^*$$

$$\text{Current SMC} \rightarrow v_d, v_q$$

The d-axis current reference i_d^* is usually chosen as a constant value or computed using an optimization strategy such as MTPA.

III.5.3 Current sliding mode control

➤ d-axis current error

Define the d-axis current tracking error:

$$e_d = i_d^* - i_d \quad (\text{III.23})$$

Choose the sliding surface:

$$s_d = e_d \quad (III.24)$$

Therefore:

$$s_d = i_d^* - i_d \quad (III.25)$$

The derivative is:

$$\dot{s}_d = \dot{i}_d^* - \dot{i}_d \quad (III.26)$$

Using the SynRM d-axis model:

$$i_d = \frac{1}{L_d} (v_d - R_s i_d - \omega_e L_q i_q) \quad (III.27)$$

So:

$$\dot{s}_d = \dot{i}_d^* - \frac{1}{L_d} (v_d - R_s i_d - \omega_e L_q i_q) \quad (III.28)$$

The reaching law is chosen as:

$$\dot{s}_d = -c_d s_d - k_d \text{sat}\left(\frac{s_d}{\phi_d}\right) \quad (III.29)$$

where:

$$c_d > 0$$

$$k_d > 0$$

$$\phi_d > 0$$

The saturation function is used instead of the sign function to reduce chattering. Thus:

$$\dot{i}_d^* - \frac{1}{L_d} (v_d - R_s i_d + \omega_e L_q i_q) = -c_d s_d - k_d \text{sat}\left(\frac{s_d}{\phi_d}\right) \quad (III.30)$$

Solving for v_d :

$$v_d = L_d \left[\dot{i}_d^* + c_d s_d + k_d \text{sat}\left(\frac{s_d}{\phi_d}\right) \right] + R_s i_d - \omega_e L_q i_q \quad (III.31)$$

This is the SMC d-axis voltage control law.

➤ **q-axis current error**

Define the q-axis current tracking error:

$$e_q = i_q^* - i_q \quad (III.32)$$

Choose:

$$s_q = e_q \quad (III.33)$$

So:

$$s_q = i_q^* - i_q \quad (III.34)$$

The derivative is:

$$\dot{s}_q = \dot{i}_q^* - \dot{i}_q \quad (III.35)$$

Using the q-axis current equation:

$$i_q = \frac{1}{L_q} (v_q - R_s i_q - \omega_e L_d i_d) \quad (III.36)$$

Therefore:

$$\dot{s}_q = \dot{i}_q^* - \frac{1}{L_q} (v_q - R_s i_q - \omega_e L_d i_d) \quad (III.37)$$

Choose the reaching law:

$$\dot{s}_q = -c_q s_q - k_q \text{sat} \left(\frac{s_q}{\phi_q} \right) \quad (III.37)$$

Then:

$$\dot{i}_q^* - \frac{1}{L_q} (v_q - R_s i_q - \omega_e L_d i_d) = -c_q s_q - k_q \text{sat} \left(\frac{s_q}{\phi_q} \right) \quad (III.39)$$

Solving for v_q :

$$v_q = L_q \left[\dot{i}_q^* + c_q s_q + k_q \text{sat} \left(\frac{s_q}{\phi_q} \right) \right] + R_s i_q - \omega_e L_d i_d \quad (III.40)$$

This is the SMC q-axis voltage control law.

III.5.4 Speed sliding mode control

The speed loop generates the q-axis current reference i_q^* .

The mechanical equation is:

$$\dot{\omega}_m = \frac{1}{J} (K_T i_d i_q - T_L - B \omega_m) \quad (III.41)$$

The speed error is:

$$e_\omega = \omega_m^* - \omega_m \quad (III.42)$$

A common sliding surface for the speed loop is:

$$s_\omega = e_\omega + \lambda_\omega \int e_\omega dt \quad (III.43)$$

where:

$$\lambda_{\omega} > 0$$

The derivative is:

$$\dot{s}_{\omega} = \dot{e}_{\omega} + \lambda_{\omega} e_{\omega} \quad (\text{III.44})$$

Since:

$$\dot{e}_{\omega} = \dot{\omega}_m^* - \dot{\omega}_m \quad (\text{III.45})$$

then:

$$\dot{s}_{\omega} = \dot{\omega}_m^* - \dot{\omega}_m + \lambda_{\omega} e_{\omega} \quad (\text{III.46})$$

Substitute the mechanical model:

$$\dot{s}_{\omega} = \dot{\omega}_m^* - \frac{1}{J} (K_T i_d i_q - T_L - B \omega_m) + \lambda_{\omega} e_{\omega} \quad (\text{III.47})$$

Choose the speed reaching law:

$$\dot{s}_{\omega} = -c_{\omega} s_{\omega} - k_{\omega} \text{sat}\left(\frac{s_{\omega}}{\phi_{\omega}}\right) \quad (\text{III.48})$$

Therefore:

$$\dot{\omega}_m^* - \frac{1}{J} (K_T i_d i_q - T_L - B \omega_m) + \lambda_{\omega} e_{\omega} = -c_{\omega} s_{\omega} - k_{\omega} \text{sat}\left(\frac{s_{\omega}}{\phi_{\omega}}\right) \quad (\text{III.49})$$

Solving for i_q , the required q-axis current reference is:

$$i_q^* = \frac{J \left[\dot{\omega}_m^* + \lambda_{\omega} e_{\omega} + c_{\omega} s_{\omega} + k_{\omega} \text{sat}\left(\frac{s_{\omega}}{\phi_{\omega}}\right) \right] + \hat{T}_L + B \omega_m}{K_T i_q^*} \quad (\text{III.50})$$

where: $\hat{T}_L$ is the estimated load torque. Since:

$$K_T = \frac{3}{2} p (L_d - L_q) \quad (\text{III.51})$$

the expression becomes:

$$i_q^* = \frac{J \left[\dot{\omega}_m^* + \lambda_{\omega} e_{\omega} + c_{\omega} s_{\omega} + k_{\omega} \text{sat}\left(\frac{s_{\omega}}{\phi_{\omega}}\right) \right] + \hat{T}_L + B \omega_m}{\frac{3}{2} p (L_d - L_q) i_q^*} \quad (\text{III.52})$$

Important condition:

$$i_q^* \neq 0$$

because the SynRM torque is:

$$T_e = K_T i_d i_q \quad (\text{III.53})$$

If $i_q^* = 0$, the torque becomes zero.

III.5.5 Complete SMC controller of SynRM Speed loop

$$e_\omega = \omega_m^* - \omega_m \quad (\text{III.54})$$

$$s_\omega = e_\omega + \lambda_\omega \int e_\omega dt \quad (\text{III.55})$$

$$i_q^* = \frac{J \left[\dot{\omega}_m^* + \lambda_\omega e_\omega + c_\omega s_\omega + k_\omega \text{sat} \left(\frac{s_\omega}{\phi_\omega} \right) \right] + \hat{T}_L + B \omega_m}{\frac{3}{2} p (L_d - L_q) i_q^*}$$

➤ d-axis current loop

$$e_d = i_d^* - i_d \quad (\text{III.56})$$

$$s_d = e_d \quad (\text{III.57})$$

$$v_d = L_d \left[\dot{i}_d^* + c_d s_d + k_d \text{sat} \left(\frac{s_d}{\phi_d} \right) \right] + R_s i_d + \omega_e L_q i_q \quad (\text{III.58})$$

➤ q-axis current loop

$$e_q = i_q^* - i_q \quad (\text{III.59})$$

$$s_q = e_q \quad (\text{III.60})$$

$$v_q = L_q \left[\dot{i}_q^* + c_q s_q + k_q \text{sat} \left(\frac{s_q}{\phi_q} \right) \right] + R_s i_q + \omega_e L_d i_d \quad (\text{III.61})$$

III.5.6. Equivalent control and switching control form

The SMC voltage control can be written as:

$$v_d = v_{deq} + v_{dn} \quad (\text{III.62})$$

$$v_q = v_{qeq} + v_{qn} \quad (\text{III.63})$$

For the d-axis:

$$v_{deq} = L_d \left(\dot{i}_d^* + c_d s_d \right) + R_s i_d + \omega_e L_q i_q \quad (\text{III.64})$$

$$v_{dn} = L_d k_d \text{sat} \left(\frac{s_d}{\phi_d} \right) \quad (\text{III.65})$$

For the q-axis:

$$v_{qeq} = L_q \left(\dot{i}_q^* + c_q s_q \right) + R_s i_q + \omega_e L_d i_d \quad (\text{III.66})$$

$$v_{qn} = L_q k_q \text{sat} \left(\frac{s_q}{\phi_q} \right) \quad (\text{III.67})$$

The terms $v_{d,eq}$ and $v_{q,eq}$ represent the continuous equivalent control. The terms $v_{d,sw}$ and $v_{q,sw}$ represent the robust switching control.

III.5.7. Stability proof using Lyapunov function

Choose the Lyapunov function:

$$V = \frac{1}{2} s^2 \quad (\text{III.68})$$

For the d-axis:

$$V_d = \frac{1}{2} s_d^2 \quad (\text{III.69})$$

$$\dot{V}_d = s_d \dot{s}_d \quad (\text{III.70})$$

Using:

$$\dot{s}_d = -c_d s_d - k_d \text{sat} \left(\frac{s_d}{\phi_d} \right) \quad (\text{III.71})$$

we get:

$$\dot{V}_d = -c_d s_d^2 - k_d s_d \text{sat} \left(\frac{s_d}{\phi_d} \right) \quad (\text{III.72})$$

Since:

$$c_d > 0, \quad k_d > 0$$

then:

$$\dot{V}_d \leq 0$$

Therefore, the d-axis current error converges to zero. The same reasoning applies to the q-axis:

$$\dot{V}_q \leq 0$$

and to the speed loop:

$$\dot{V}_\omega \leq 0$$

Thus, the complete SMC-controlled SynRM system is stable.

III.5.8. Saturation function

To reduce chattering, use:

$$\text{sat}\left(\frac{s}{\phi}\right) = \begin{cases} 1, & \frac{s}{\phi} > 1 \\ \frac{s}{\phi}, & \left|\frac{s}{\phi}\right| \leq 1 \\ -1, & \frac{s}{\phi} < -1 \end{cases} \quad (\text{III.73})$$

If a pure sliding mode controller is required, replace saturation by sign:

$$\text{sat}\left(\frac{s}{\phi}\right) \rightarrow \text{sign}(s)$$

Then:

$$\dot{s} = -cs - k \text{sign}(s) \quad (\text{III.74})$$

But this causes more chattering.

III.5.9. Final complete control law

$$v_d = L_d \left[\dot{i}_d^* + c_d(i_d^* - i_d) + k_d \text{sat}\left(\frac{i_d^* - i_d}{\phi_d}\right) \right] + R_s i_d + p\omega_m L_q i_q \quad (\text{III.75})$$

$$v_q = L_q \left[\dot{i}_q^* + c_d(i_q^* - i_q) + k_q \text{sat}\left(\frac{i_q^* - i_q}{\phi_d}\right) \right] + R_s i_q + p\omega_m L_d i_d \quad (\text{III.76})$$

$$i_q^* = \frac{J[\dot{\omega}_m^* + \lambda_\omega(\omega_m^* - \omega_m) + c_\omega s_\omega + k_\omega \text{sat}\left(\frac{s_\omega}{\phi_\omega}\right)] + \hat{T}_L + B\omega_m}{\frac{3}{2}p(L_d - L_q)i_d^*} \quad (\text{III.77})$$

with:

$$s_\omega = (\omega_m^* - \omega_m) + \lambda_\omega \int (\omega_m^* - \omega_m) dt \quad (\text{III.78})$$

III.6 Model-Free Sliding Mode Control

III.6.1 Recurrent neural network with online training:

The recurrent neural network model is one of the most widespread NN used models for systems state identification [RRU19]. It is more practical for discrete-time systems that demand adaptability to different operating conditions. In this study, the RNN is proposed to identify the SynRM three-phase currents without considering the system parameters.

III.6.1.1 RNN identifier for unknown plan

A discrete-time nonlinear system can be defined as in [SGE04]:

$$\mathbf{x}_{k+1} = F(\mathbf{x}_k, \boldsymbol{\vartheta}_k) \quad (\text{III.79})$$

To identify the variables states of the considered system in (3.78) taking into account the dynamic approximation issue, a series-parallel RNN configuration is proposed as in [LDJ19]:

$$\mathbf{X}_{i,k+1} = \mathbf{w}_i^T \boldsymbol{\Phi}(\mathbf{x}_k, \boldsymbol{\vartheta}_k) + \boldsymbol{\varpi}_i^T \boldsymbol{\psi}(\mathbf{x}_k, \boldsymbol{\vartheta}_k), \quad i = 1, \dots, n \quad (\text{III.80})$$

where $\mathbf{x}_k = [x_{1,k}, x_{2,k}, \dots, x_{n,k}]$ is the plant state vector, X_i is the estimated state of the i th neuron, which corresponds the i th component of state vector $\mathbf{x}_k$ in (3.79), $\mathbf{w}_i \in \mathfrak{R}^{L_i}$ is the vector of adjustable weights, $\boldsymbol{\varpi}_i$ are matrices with constant values, $\boldsymbol{\vartheta} \in \mathfrak{R}^m$ is the input vector, $\boldsymbol{\psi}_i$ denotes a linear function, and $\boldsymbol{\phi}_i \in \mathfrak{R}^{L_i}$ is the corresponding vector function, which can be defined as:

$$\boldsymbol{\Phi}_i(\mathbf{x}_k, \boldsymbol{\vartheta}_k) = \begin{bmatrix} \boldsymbol{\Phi}_{i_1} \\ \boldsymbol{\Phi}_{i_2} \\ \vdots \\ \boldsymbol{\Phi}_{i_{B_i}} \end{bmatrix} = \begin{bmatrix} \prod_{j \in I_1} \delta_{i_j}^{d_{ij}(1)} \\ \prod_{j \in I_2} \delta_{i_j}^{d_{ij}(2)} \\ \vdots \\ \prod_{j \in I_{B_i}} \delta_{i_j}^{d_{ij}(B_i)} \end{bmatrix} \quad (\text{III.81})$$

where $d_{ij,k}$ are nonnegative integers, B_i is the number of connections, $\{I_1, I_2, \dots, I_{B_i}\}$ is a collection of non-ordered subsets of $\{1, 2, \dots, n + m\}$, n is the state dimension, m is the number of external inputs, and δ_i is presented as:

$$\boldsymbol{\delta}_i = \begin{bmatrix} \delta_{i_1} \\ \vdots \\ \delta_{i_n} \\ \delta_{i_{n+1}} \\ \vdots \\ \delta_{i_{n+m}} \end{bmatrix} = \begin{bmatrix} S(\mathbf{x}_1) \\ \vdots \\ S(\mathbf{x}_n) \\ \boldsymbol{\vartheta}_1 \\ \vdots \\ \boldsymbol{\vartheta}_m \end{bmatrix} \quad (\text{III.82})$$

where $\boldsymbol{\vartheta}_i = [\vartheta_1, \vartheta_2, \dots, \vartheta_m]^T$ are the RNN external inputs and $S(\bullet)$ can be described as follows:

$$S(\ell_i) = \alpha_i \tanh(\beta_i \ell_i) \quad \text{with } \alpha \text{ and } \beta > 0 \quad (\text{III.83})$$

where ℓ is the variable state.

III.6.1.2 EKF training algorithm

The RNN representing the system model is trained online through the EKF training algorithm. The EKF-based training is the most well-known online approach, being the fastest one in the learning process, by converging in less iterations as compared with other methods [ESA08]. The main objective of the training procedure is to reduce the prediction error between the real

and estimated states (x_i and X_i), by generating the adaptive weights corresponding to the respective state for the i_{th} neuron. The EKF training algorithm is defined by the following equations [RRU19],[ESA08]:

$$\mathbf{e}_{i,k} = \mathbf{x}_{i,k} - \mathbf{X}_{i,k} \quad (III.84)$$

$$\mathbf{K}_{i,k} = \mathbf{P}_{i,k} \mathbf{H}_{i,k} \mathbf{M}_{i,k}^{-1} \quad (III.85)$$

$$\mathbf{w}_{i,k+1} = \mathbf{w}_{i,k} + \eta_i \mathbf{K}_{i,k} \mathbf{e}_{i,k} \quad (III.86)$$

$$\mathbf{P}_{i,k+1} = \mathbf{P}_{i,k} - \mathbf{K}_{i,k} \mathbf{H}_{i,k}^T \mathbf{P}_{i,k} + \mathbf{Q}_{i,k} \quad (III.87)$$

$$\mathbf{M}_{i,k} = \mathbf{R}_{i,k} + \mathbf{H}_{i,k}^T \mathbf{P}_{i,k} \mathbf{H}_{i,k} \quad (III.88)$$

where $\mathbf{e}_{i,k} \in \Re$ is the real error, $\mathbf{P}_{i,k+1} \in \Re^{Li \times Li}$ is the prediction error, Li is the RNN weights number, η_i is the learning rate, $\mathbf{w}_i \in \Re^{Li}$ denotes the weights matrix which is considered as the system state for the EKF, $\mathbf{K}_{i,k} \in \Re^{Li \times m}$ is the EKF gain matrix, $\mathbf{Q}_{i,k} \in \Re^{Li \times Li}$ and $\mathbf{R}_{i,k} \in \Re^{m \times m}$ are the covariance matrices, $\mathbf{H}_{i,k} \in \Re^{Li \times m}$ is a gradient matrix, in which each entry ($H_{ij,k}$) is the derivative of one RNN state X_i relative to one RNN weight $w_{ij,k}$, and calculated as follows:

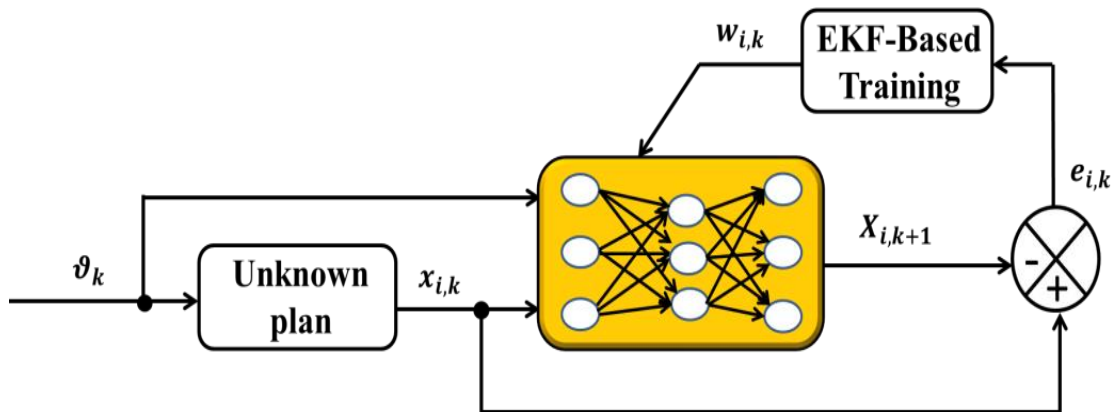


Figure 3-7: Training algorithm scheme.

$$H_{ij,k} = \left[\frac{\partial X_{i,k}}{\partial w_{ij,k}} \right]_{w_{i,k}=w_{i,k+1}}^T \quad (i=1,2,\dots,n) \text{ and } (j=1,2,\dots,Li) \quad (III.89)$$

Generally, the RNN states are initialized as zero and the covariance matrices $\mathbf{P}_{i,k}$, $\mathbf{Q}_{i,k}$, and $\mathbf{R}_{i,k}$ are diagonals. In fact, there are available several studies that explain in detail the training as well as the convergence analysis of the RNN identifier based on the EKF technique [ESA08], [AYA07]. Figure III-7 shows the training algorithm scheme considered in this study.

In the final step, the RNN identifier model can be derived from (III.79), (III.80) and (3.81) as follow:

$$\mathbf{X}_{i,k+1} = \mathbf{w}_{i1}\mathbf{S}(\mathbf{x}_{i,k}) + \mathbf{w}_{i2}\mathbf{S}(\mathbf{x}_{j,k}) + \mathbf{w}_{i3}\mathbf{S}(\mathbf{x}_{j+1,k}) + \dots + \mathbf{w}_{iz}\mathbf{S}(\mathbf{x}_{j+(z-1),k}) + \mathbf{\varpi}_i\mathbf{\vartheta}_{i,k} \quad (\text{III.90})$$

where $\mathbf{X}_i$ denotes the estimated state, z is the number of state, $\mathbf{x}_i$ is the respective real state, $\mathbf{x}_{j+(z-1)}$ denotes the remaining states with ($j \neq i$), and $\mathbf{\vartheta}_{i,k}$ is the respective input voltage.

III.6.2 Proposed RNN-SMC for the SynRM machine:

In conventional MB-SMC, prior knowledge of the system model, as well as accurate parameter values are mandatory. However, the parameters of the SynRM change with the working conditions, which may negatively affect its control performance. Therefore, to achieve high control performance, the proposed model-free RNN-SMC comes to be an obvious choice. Only the measured data from the drive are required, avoiding therefore the machine parameters knowledge. Using the proposed control, the parameter-based predictive currents expressions in conventional MB-SMC will be compensated by the ones derived by the proposed RNN identifier.

According to (3.90), the proposed predictive control laws at the are defined as:

$$\begin{cases} \mathbf{v}_{d,eq} = \mathbf{w}_{d1}\mathbf{S}(\mathbf{i}_d) + \mathbf{w}_{d2}\omega_e\mathbf{S}(\mathbf{i}_q) + \mathbf{\varpi}_d\mathbf{s}_d \\ \mathbf{v}_{q,eq} = \mathbf{w}_{q1}\mathbf{S}(\mathbf{i}_q) + \mathbf{w}_{q2}\omega_e\mathbf{S}(\mathbf{i}_d) + \mathbf{\varpi}_q\mathbf{s}_q \end{cases} \quad (\text{III.91})$$

where $\mathbf{w}_{d,i}$ and $\mathbf{w}_{q,i}$ are updated in real time by the EKF using the error between measured and estimated currents. Based on the RNN-SMC definition the proposed RNN identifier does not require pre-training before implementation, wherein the initial weights are chosen randomly and the $\mathbf{\varpi}_d$ and $\mathbf{\varpi}_q$ in (3.91) are considered equal to $\mathbf{\varpi}_d=5$, and $\mathbf{\varpi}_q=5$, respectively. On the other hand, the covariance matrixes in (3.86) and (3.87) are set to $\mathbf{Q}_i=1000$ and $\mathbf{R}_i=100$. It must be noticed that the selection of $\mathbf{\varpi}_d$ and $\mathbf{\varpi}_q$ is very important since they have a strong impact on the control performance, whereas the values of $\mathbf{R}_i$ and $\mathbf{Q}_i$ have a direct relation with the dynamic of the EKF. The definition of these values is accomplished using trial and error procedures, by analyzing the control performance for different operating conditions, taking into account the dynamic model accuracy as well as the low current ripple. Furthermore, it is important to emphasize that these values are chosen constant in order to give the best possible performance as well as to simplify the controller design.

In contrast to the conventional MB-SMC using (III.64) - (III.66), the RNN-SMC in (III.91) does not use any machine parameters. The diagram scheme of the proposed RNN-SMC algorithm is shown in Figure III-8.

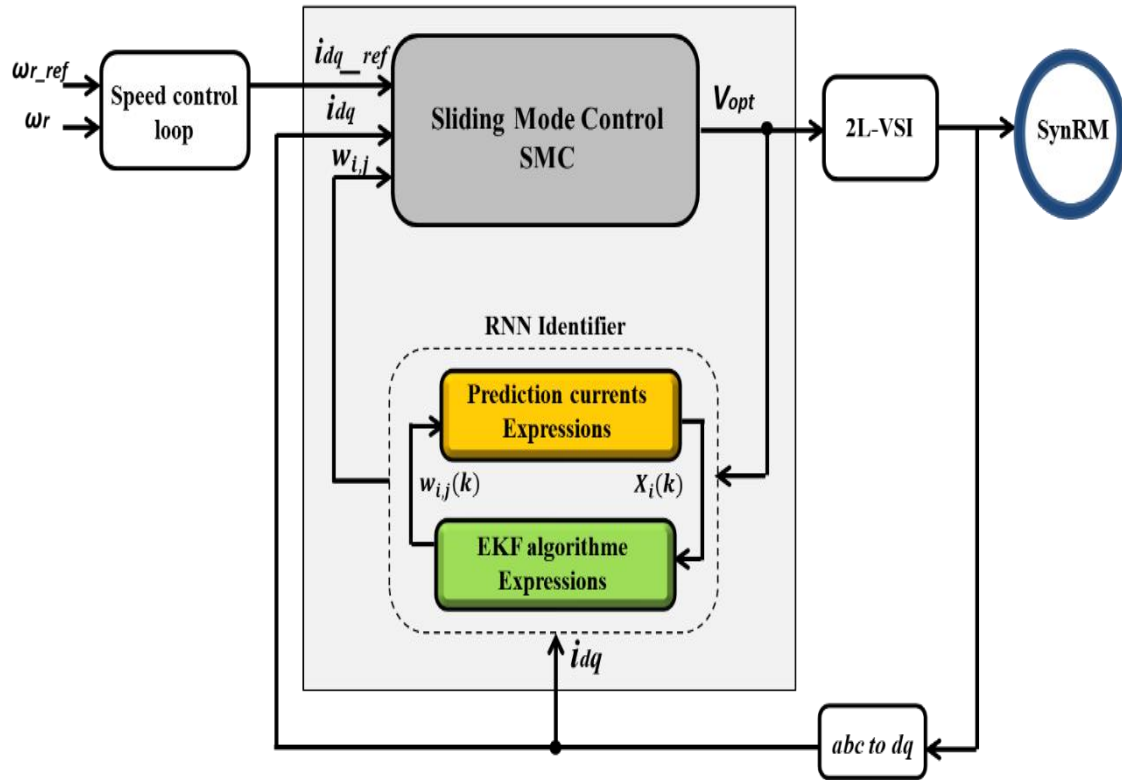


Figure III-8: Block diagram of the RNN-SMC proposed for SynRM machines.

Variable	Symbol	Value
d-axis inductance	L_d	$150e^{-3}[\text{H}]$
q-axis inductance	L_q	$25e^{-3}[\text{H}]$
Number of poles	P	2
Friction coefficient	B	$0.00036[\text{N.m}^{-1}]$
Moment of inertia	J	$0.0137[\text{Kg.m}^2]$
Stator resistance	R_s	$1.71[\Omega]$

Table III-1: Parameters of the SynRM machine [MIL25]

III.7 Simulation Results

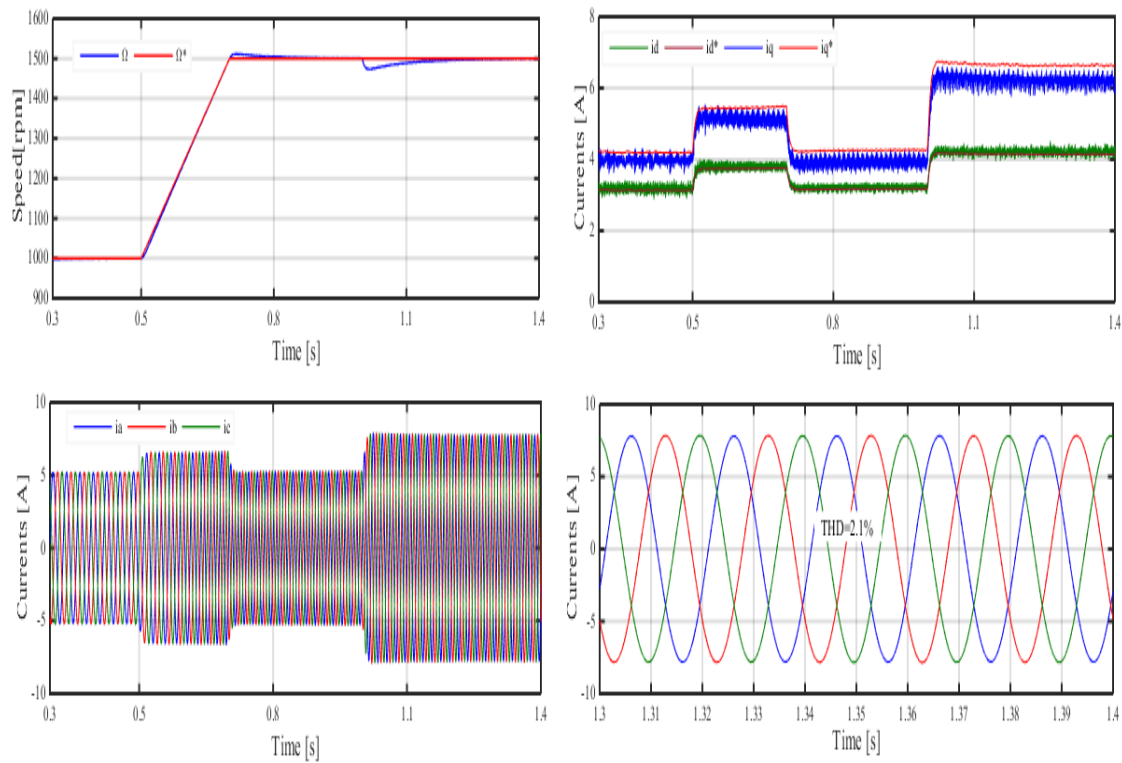


Figure III-9: Simulation results under speed and load variations of conventional MB-SMC with nominal parameters.

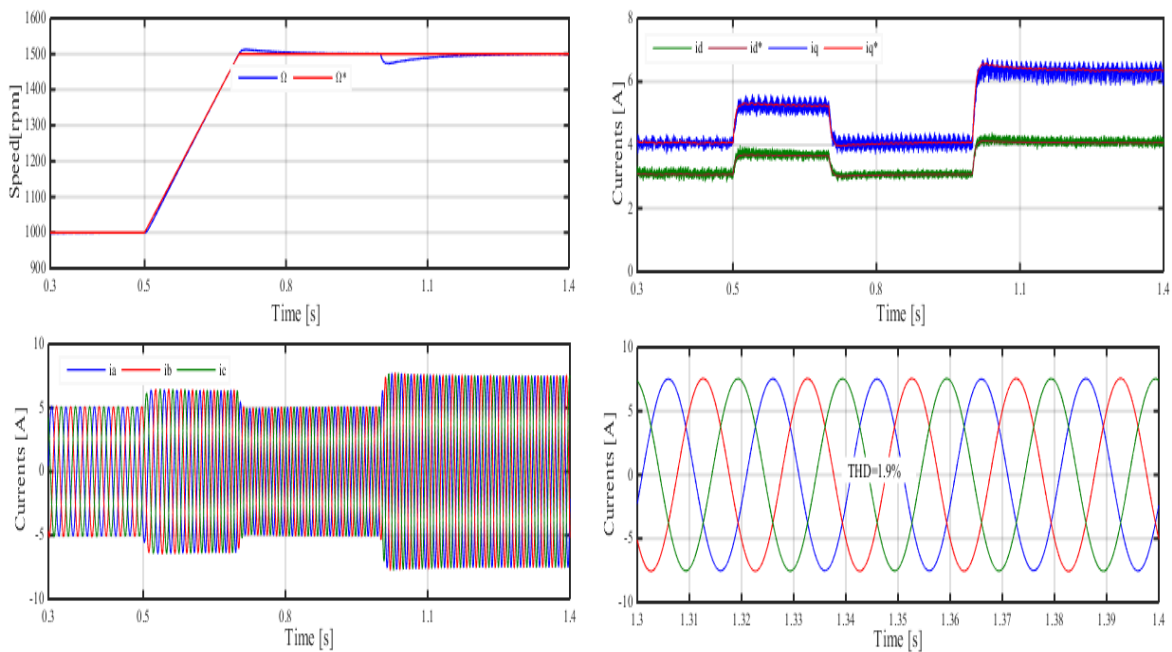


Figure III-10: Simulation results under speed and load variations of the proposed RNN-SMC with nominal parameters.

Figures III-9 and III-10 present the simulation results of the conventional MB-SMC and the proposed model-free RNN-SMC considering the same operating conditions. Firstly, the reference speed is set to 1000 rpm under a load torque of 7 N.m. Then, at the instant $t = 0.5$ s, the reference speed increases to its rated value, whereas at the instant $t = 1$ s the load torque steps to its rated value. The performance of the MB-SMC with nominal parameters is illustrated in Figure III III-9, where the bottom figures show the actual and zoomed SynRM currents, and the above figures illustrates the rotor speed together with its reference value and the dq-axes currents with their references, respectively. Analyzing the results, it is clearly observed that even with the nominal parameters of the MB-SMC, an obvious deviation is aroused in the q-axis current compared to its reference, which increases as the speed and load torque increase. However, in Figure III-10, the proposed RNN-SMC present best performance,

▪ Robustness tests :

In order to test the robustness of the proposed RNN-SMC, a parameters variation test is carried out, where the d-q inductance values are changed to -50%.

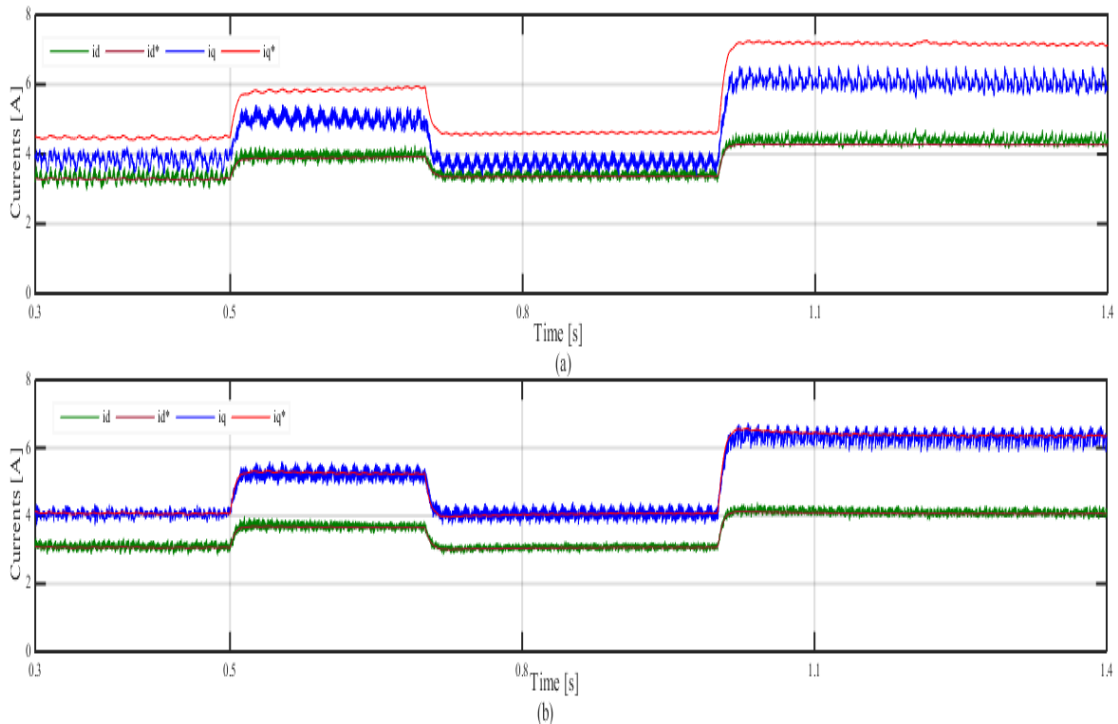


Figure III-11: Simulation results under speed and load variations of: (a) conventional MB-SMC under the influence of parameter variations ($0.5 \times L_d$ and $0.5 \times L_q$); (b) proposed RNN-SMC.

Because the MB-SMC is sensitive to parameter variations, the tracking q-axis current error increases even more under inductances variations, where the dq-axes inductances values are set lower than their nominal values (see Figure III-10(a)). As a result, high current ripples and harmonic distortions are obtained, especially under inductances variations. It is important to emphasize that the inductances variation is performed by changing their values in the MB-SMC algorithm. On the contrary, since the proposed RNN-SMC does not require the knowledge of the motor parameters, the tracking error of q-axis current is eliminated (see Figure III-11 (b)), showing an inherent immunity against the SynRM parameter variations.

As conclusion, both control schemes can achieve a fast dynamic response and provide a similar rotor speed performance, the considered RNN-SMC performs better in harmonic distortion. From the considered control strategies, it can be concluded that the MB-SMC with inductance variations has the lowest performance by presenting the highest THD values, while the proposed RNN-SMC shows the best performance under all the considered operating conditions.

III.8 Conclusion:

In conclusion, this chapter presented the design and application of Sliding Mode Control (SMC) for the Synchronous Reluctance Motor (SynRM), highlighting its robustness against parameter variations and external disturbances. A model-free control approach based on Recurrent Neural Networks (RNNs) and online EKF training was also introduced to eliminate the dependence on accurate machine parameters. The obtained results demonstrate that the proposed RNN-SMC strategy enhances control performance, improves tracking accuracy, and provides reliable operation under different operating conditions, making it a promising solution for advanced SynRM drive systems.

General Conclusion

General Conclusion

This thesis has focused on the modeling, control, and performance enhancement of the Synchronous Reluctance Motor (SynRM), with the primary goal of improving speed regulation and dynamic behavior using advanced control techniques, particularly Sliding Mode Control (SMC). This study investigated detailed mathematical modeling, performance analysis in different reference frames, and the design of a robust controller capable of handling nonlinearities, load disturbances, and parameter variations.

The key outcomes of the study can be summarized as follows:

Chapter I provided a comprehensive overview of Synchronous Reluctance Motors (SynRM), focusing on their historical development, different rotor structures (flux barriers, axially/radially laminated, permanent magnet-assisted). The unique advantages of these motors, such as high efficiency, low manufacturing cost, and reduced maintenance, were highlighted, along with their growing applications in sectors like electric vehicles and high-speed industrial machinery.

Chapter II focused on the precise mathematical modeling of the SynRM in the rotating (dq) reference frame using Park's transformation, after applying Clarke's transformation. A nonlinear state-space model accurately describing currents and speed was developed. This chapter also covered the modeling of the two-level Voltage Source Inverter (VSI) and Pulse Width Modulation (PWM) technique, and the influence of supply on torque ripple was analyzed.

Chapter III presented a comprehensive study of Sliding Mode Control (SMC) and its application to the Synchronous Reluctance Motor (SynRM). The theoretical foundations of SMC, including the design of sliding surfaces, convergence conditions, control law synthesis, and Lyapunov-based stability analysis, were thoroughly discussed. The robustness of SMC against parameter uncertainties and external disturbances makes it a suitable control strategy for high-performance SynRM drives.

A complete SMC-based control structure was developed for regulating the speed and current components of the SynRM in the d-q reference frame. To address the limitations associated with model dependency and parameter variations, a model-free control approach based on Recurrent Neural Networks (RNNs) was also investigated. The proposed RNN identifier, trained online using the Extended Kalman Filter (EKF), enables accurate prediction of system states without requiring prior knowledge of machine parameters.

Building on the outcomes of this thesis, several potential directions for future research can be proposed to further advance the control and performance of Synchronous Reluctance Motors:

- Integration of chattering reduction techniques: such as using smooth sign functions (e.g., Sat function instead of Sign), or adaptive SMC or higher-order SMC (HOSMC) to mitigate high-frequency oscillations.
- Extension to multilevel inverter topologies: such as neutral-point clamped (NPC) or cascaded H-bridge inverters, which could offer enhanced output waveform quality with lower harmonic distortion and higher efficiency.
- Integration of hybrid control strategies: combining sliding mode control with fuzzy logic (Fuzzy-SMC) or artificial neural networks (ANN-SMC) or adaptive algorithms, to further improve dynamic performance and robustness under variable load and speed conditions.
- Sensorless control: developing observers such as sliding mode observers or Lyapunov-based observers to accurately estimate speed and position, thereby reducing cost and increasing reliability.
- Experimental validation and hardware implementation: building a prototype of the SynRM and the proposed SMC controller, and testing the results on a real-time experimental platform (such as dSPACE or FPGA) to confirm effectiveness in real-time applications.

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