



# Master Thesis

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## Title

# Study and Design of an Intelligent Speed Controller for an Electric Motor

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## ***Dedication***

*I dedicate this modest work to my dearest mother, whose endless sacrifices and unwavering support have been the foundation of everything I have achieved. From my earliest years, she has been my greatest source of strength and encouragement, and no words can ever truly express my gratitude.*

*I also dedicate this work to my dearest grandfather, who always believed in me and never stopped pushing me forward with his words of encouragement and faith in my potential.*

*To my two brothers, my little sister, and my dear aunt — this work is yours as much as it is mine. Your presence in my life has been a constant source of warmth and motivation.*

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*To my entire family, thank you for everything.*

***DERRADJ ABDELLAH***

## *Dedication*

*I dedicate this modest work to my dearest mother, who has always been a source of love, tenderness, patience, and strength.*

*Thank you for your sacrifices, your prayers, your moral support, and your encouragement throughout my journey. No words can fully express my deep gratitude, respect, and love for you.*

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*Thank you from the bottom of my heart*

*ABBAS Sofiane*

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## Abstract

This thesis presents the study and design of an intelligent speed controller for electric motors, focusing on the separately excited DC motor, the single-phase induction motor, and the three-phase induction motor. Each machine is mathematically modeled and simulated using MATLAB/Simulink under different operating conditions.

Classical PI/PID controllers are first applied for speed regulation, then ANFIS-based intelligent controllers are introduced to improve performance and robustness. The results show that the ANFIS controllers provide better speed tracking, reduced overshoot, shorter settling time, and improved robustness against load and parameter variations.

Keywords: variable speed drive, DC motor, induction motor, V/f control, vector control, ANFIS.

## المخلص

تتناول هذه المذكرة دراسة وتصميم متحكم ذكي في سرعة المحركات الكهربائية، مع التركيز على محرك التيار المستمر ذو الإثارة المستقلة، والمحرك الحثي أحادي الطور، والمحرك الحثي ثلاثي الطور. تم نمذجة كل آلة رياضياً ومحاكاتها باستعمال MATLAB/Simulink تحت ظروف تشغيل مختلفة.

تم أولاً استعمال المتحكمات الكلاسيكية PI/PID لتنظيم السرعة، ثم تم إدماج متحكمات ذكية تعتمد على ANFIS من أجل تحسين الأداء والمتانة. أظهرت النتائج أن متحكمات ANFIS تحقق تتبعاً أفضل للسرعة، وتقليلًا في التجاوز، وزمناً أقصر للاستقرار، ومتانة أعلى ضد تغيرات الحمل ومعاملات المحرك.

الكلمات المفتاحية: مغير السرعة، محرك التيار المستمر، المحرك الحثي، التحكم V/f، التحكم الشعاعي، ANFIS.

## Résumé

Ce mémoire présente l'étude et la conception d'un contrôleur intelligent de vitesse pour les moteurs électriques, en se concentrant sur le moteur à courant continu à excitation séparée, le moteur asynchrone monophasé et le moteur asynchrone triphasé. Chaque machine est modélisée mathématiquement puis simulée sous MATLAB/Simulink dans différentes conditions de fonctionnement.

Des contrôleurs classiques PI/PID sont d'abord utilisés pour la régulation de vitesse, puis des contrôleurs intelligents basés sur l'ANFIS sont introduits afin d'améliorer les performances et la robustesse du système. Les résultats montrent que les contrôleurs ANFIS assurent un meilleur suivi de vitesse, une réduction du dépassement, un temps de réponse plus court et une meilleure robustesse face aux variations de charge et de paramètres.

Mots-clés : variateur de vitesse, moteur à courant continu, moteur asynchrone, commande V/f, commande vectorielle, ANFIS.

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## LIST OF ACRONYMS

|               |                                                                                                           |
|---------------|-----------------------------------------------------------------------------------------------------------|
| <b>AC</b>     | : Alternating Current                                                                                     |
| <b>ANFIS</b>  | : Adaptive Neuro-Fuzzy Inference System                                                                   |
| <b>AW</b>     | : Auxiliary Winding                                                                                       |
| <b>DC</b>     | : Direct Current                                                                                          |
| <b>DSP</b>    | : Digital Signal Processor                                                                                |
| <b>EMF</b>    | : Electromotive Force                                                                                     |
| <b>FIS</b>    | : Fuzzy Inference System                                                                                  |
| <b>FLC</b>    | : Fuzzy Logic Controller                                                                                  |
| <b>FOC</b>    | : Field Oriented Control                                                                                  |
| <b>GTO</b>    | : Gate Turn-Off Thyristor                                                                                 |
| <b>IGBT</b>   | : Insulated Gate Bipolar Transistor                                                                       |
| <b>MAMCP</b>  | : Moteur Asynchrone Monophasé à Condensateur Permanent (Permanent Capacitor Single-Phase Induction Motor) |
| <b>MATLAB</b> | : Matrix Laboratory                                                                                       |
| <b>MMF</b>    | : Magnetomotive Force                                                                                     |
| <b>MOSFET</b> | : Metal Oxide Semiconductor Field Effect Transistor                                                       |
| <b>MW</b>     | : Main Winding                                                                                            |
| <b>NNI</b>    | : Neural Network Identifier                                                                               |
| <b>PI</b>     | : Proportional-Integral                                                                                   |
| <b>PID</b>    | : Proportional-Integral-Derivative                                                                        |
| <b>PWM</b>    | : Pulse Width Modulation                                                                                  |
| <b>RPM</b>    | : Revolutions Per Minute                                                                                  |
| <b>SPWM</b>   | : Sinusoidal Pulse Width Modulation                                                                       |
| <b>VFD</b>    | : Variable Frequency Drive                                                                                |
| <b>VSD</b>    | : Variable Speed Drive                                                                                    |

## Symbols and Abbreviations

| Symbol                  | Meaning                                                   | Unit              |
|-------------------------|-----------------------------------------------------------|-------------------|
| <b>Latin Symbols</b>    |                                                           |                   |
| <b>a</b>                | Winding transformation coefficient (auxiliary/main ratio) | Dimensionless     |
| <b>A</b>                | State matrix of the system                                | —                 |
| <b>B</b>                | Input matrix of the system                                | —                 |
| <b>B</b>                | Magnetic flux density                                     | T                 |
| <b>B<sub>m</sub></b>    | Viscous friction coefficient                              | N·m·s/rad         |
| <b>C</b>                | Capacitance of the series capacitor (auxiliary winding)   | F                 |
| <b>C<sub>f</sub></b>    | Filter capacitance                                        | F                 |
| <b>C<sub>em</sub></b>   | Electromagnetic torque                                    | N·m               |
| <b>C<sub>r</sub></b>    | Load (resistive) torque                                   | N·m               |
| <b>C<sub>(s)</sub></b>  | Transfer function of the PI/PID controller                | —                 |
| <b>e</b>                | Back electromotive force (back-EMF)                       | V                 |
| <b>e<sub>b</sub></b>    | Back electromotive force of the armature                  | V                 |
| <b>E</b>                | Input DC voltage of the chopper                           | V                 |
| <b>E<sub>l</sub></b>    | Induced stator electromotive force                        | V                 |
| <b>e<sub>(t)</sub></b>  | Speed tracking error signal                               | rad/s             |
| <b>f</b>                | Supply frequency                                          | Hz                |
| <b>f</b>                | Viscous friction coefficient                              | N·m·s/rad         |
| <b>f<sub>c</sub></b>    | Cut-off frequency of the LC filter                        | Hz                |
| <b>F</b>                | Electromagnetic force acting on a conductor               | N                 |
| <b>F<sub>(s)</sub></b>  | LC filter transfer function                               | —                 |
| <b>G<sub>m(s)</sub></b> | Simplified mechanical transfer function of the motor      | —                 |
| <b>I<sub>a</sub></b>    | Armature current                                          | A                 |
| <b>I<sub>ch</sub></b>   | Load current at chopper output                            | A                 |
| <b>I<sub>d</sub></b>    | Direct-axis current (d-axis), FOC                         | A                 |
| <b>I<sub>ds</sub></b>   | Stator current, d-axis component                          | A                 |
| <b>I<sub>dr</sub></b>   | Rotor current, d-axis component                           | A                 |
| <b>I<sub>f</sub></b>    | Field excitation current                                  | A                 |
| <b>i<sub>L</sub></b>    | Filter inductor current                                   | A                 |
| <b>i<sub>load</sub></b> | Load current at inverter output                           | A                 |
| <b>I<sub>q</sub></b>    | Quadrature-axis current (q-axis), FOC                     | A                 |
| <b>I<sub>qs</sub></b>   | Stator current, q-axis component                          | A                 |
| <b>I<sub>qr</sub></b>   | Rotor current, q-axis component                           | A                 |
| <b>I<sub>ra</sub></b>   | Rotor current, α-axis (single-phase model)                | A                 |
| <b>I<sub>rβ</sub></b>   | Rotor current, β-axis (single-phase model)                | A                 |
| <b>I<sub>sa</sub></b>   | Stator current, main winding axis (α)                     | A                 |
| <b>I<sub>sβ</sub></b>   | Stator current, auxiliary winding axis (β)                | A                 |
| <b>J</b>                | Total moment of inertia of rotor and load                 | kg·m <sup>2</sup> |
| <b>K</b>                | Motor construction constant                               | Dimensionless     |
| <b>K<sub>e</sub></b>    | Back-EMF constant                                         | V·s/rad           |
| <b>K<sub>i</sub></b>    | Integral gain of the PI/PID controller                    | Dimensionless     |
| <b>K<sub>d</sub></b>    | Derivative gain of the PID controller                     | Dimensionless     |

|              |                                                                  |                 |
|--------------|------------------------------------------------------------------|-----------------|
| $K_p$        | Proportional gain of the PI/PID controller                       | Dimensionless   |
| $K_T$        | Motor torque constant                                            | $N \cdot m/A$   |
| $K_{vf}$     | V/f proportionality constant                                     | $V \cdot s/rad$ |
| $l$          | Active length of conductor inside the magnetic field             | m               |
| $L$          | Filter inductance                                                | H               |
| $L_a$        | Armature winding inductance                                      | H               |
| $L_{af}$     | Field-armature mutual inductance                                 | H               |
| $L_{ch}$     | Load inductance                                                  | H               |
| $L_f$        | Field winding inductance                                         | H               |
| $L_f$        | Filter inductance (LC filter)                                    | H               |
| $L_{lr}$     | Rotor leakage inductance                                         | H               |
| $L_{ls}$     | Stator leakage inductance                                        | H               |
| $L_r$        | Total rotor inductance                                           | H               |
| $L_{ra}$     | Rotor inductance, $\alpha$ -axis (single-phase model)            | H               |
| $L_{r\beta}$ | Rotor inductance, $\beta$ -axis (single-phase model)             | H               |
| $L_s$        | Total stator inductance                                          | H               |
| $L_{sa}$     | Stator self-inductance, main winding ( $\alpha$ )                | H               |
| $L_{s\beta}$ | Stator self-inductance, auxiliary winding ( $\beta$ )            | H               |
| $M$          | Mutual (magnetizing) inductance between stator and rotor         | H               |
| $M_\alpha$   | Magnetizing inductance, main winding axis                        | H               |
| $M_\beta$    | Magnetizing inductance, auxiliary winding axis                   | H               |
| $M_{fd}$     | Mutual inductance between field and armature circuits            | H               |
| $M_r$        | Rotor magnetizing inductance                                     | H               |
| $M_s$        | Stator magnetizing inductance                                    | H               |
| $M_{sr}$     | Maximum mutual inductance between one stator and one rotor phase | H               |
| $N$          | Rotational speed of the motor                                    | rpm             |
| $N_l$        | Number of stator turns per phase                                 | Dimensionless   |
| $N_a$        | Actual number of turns of the auxiliary winding                  | Dimensionless   |
| $N_p$        | Actual number of turns of the main winding                       | Dimensionless   |
| $N_r$        | Actual rotor speed                                               | rpm             |
| $N_s$        | Synchronous speed of the rotating magnetic field                 | rpm             |
| $P$          | Number of pole pairs                                             | Dimensionless   |
| $R$          | Load resistance                                                  | $\Omega$        |
| $R_a$        | Armature winding resistance                                      | $\Omega$        |
| $R_{ch}$     | Load resistance at chopper output                                | $\Omega$        |
| $R_f$        | Field winding resistance                                         | $\Omega$        |
| $r_L$        | Internal resistance of the filter inductor                       | $\Omega$        |
| $r_{ra}$     | Rotor resistance, $\alpha$ -axis (referred to stator)            | $\Omega$        |
| $r_{r\beta}$ | Rotor resistance, $\beta$ -axis (referred to stator)             | $\Omega$        |
| $R_s$        | Stator phase resistance                                          | $\Omega$        |
| $r_{sa}$     | Main winding stator resistance                                   | $\Omega$        |
| $r_{s\beta}$ | Auxiliary winding stator resistance                              | $\Omega$        |
| $s$          | Slip of the induction motor                                      | Dimensionless   |
| $S_{(t)}$    | Switching state function of the PWM inverter                     | Dimensionless   |

|              |                                                           |               |
|--------------|-----------------------------------------------------------|---------------|
| $T$          | Switching period of the PWM signal                        | s             |
| $T_m$        | Mechanical time constant of the motor                     | s             |
| $T_n$        | Integral time constant of the PI controller               | s             |
| $t_{on}$     | ON-time of the power switch                               | s             |
| $t_{off}$    | OFF-time of the power switch                              | s             |
| $u_{(t)}$    | Control signal (PI/PID controller output)                 | Dimensionless |
| $U_a$        | Armature supply voltage                                   | V             |
| $U_c$        | Voltage across the series capacitor (auxiliary winding)   | V             |
| $U_f$        | Field excitation voltage                                  | V             |
| $U_{red}$    | Rectifier output DC voltage                               | V             |
| $U_{sa}$     | Stator supply voltage, main winding axis ( $\alpha$ )     | V             |
| $U_{s\beta}$ | Stator supply voltage, auxiliary winding axis ( $\beta$ ) | V             |
| $u_{ds}$     | Stator voltage, d-axis component                          | V             |
| $u_{qs}$     | Stator voltage, q-axis component                          | V             |
| $V_0$        | Average output DC voltage of the converter                | V             |
| $V_l$        | Applied stator phase voltage                              | V             |
| $V_a$        | Average armature voltage applied to the motor             | V             |
| $V_{avg}$    | Average output voltage of the PWM converter               | V             |
| $V_c$        | Capacitor output voltage (LC filter)                      | V             |
| $V_{dc}$     | Fixed DC supply voltage                                   | V             |
| $V_{in}$     | Input DC voltage of the converter                         | V             |
| $V_{inv}$    | Inverter output voltage                                   | V             |
| $V_{Lm}$     | Peak value of the line-to-line AC voltage                 | V             |
| $V_s$        | Stator phase voltage (RMS)                                | V             |
| $V_s^*$      | Stator voltage reference                                  | V             |
| $X$          | State vector of the system                                | —             |

## Greek Symbols

|                 |                                                          |                   |
|-----------------|----------------------------------------------------------|-------------------|
| $\alpha$        | Duty cycle of the PWM signal / firing angle of thyristor | Dimensionless / ° |
| $\theta$        | Angular position of the dq reference frame               | rad               |
| $\theta_0$      | Initial angle of the dq reference frame                  | rad               |
| $\Phi$          | Magnetic flux per pole                                   | Wb                |
| $\Phi_m$        | Maximum magnetic flux per pole                           | Wb                |
| $\tau$          | Time constant (general notation)                         | s                 |
| $\tau_a$        | Armature electrical time constant                        | s                 |
| $\tau_{CL}$     | Desired closed-loop time constant                        | s                 |
| $\tau_{mec}$    | Mechanical time constant of the motor                    | s                 |
| $\Psi_{ds}$     | Stator flux linkage, d-axis                              | Wb                |
| $\Psi_{dr}$     | Rotor flux linkage, d-axis                               | Wb                |
| $\Psi_{qs}$     | Stator flux linkage, q-axis                              | Wb                |
| $\Psi_{qr}$     | Rotor flux linkage, q-axis                               | Wb                |
| $\Psi_{ra}$     | Rotor flux linkage, $\alpha$ -axis (single-phase model)  | Wb                |
| $\Psi_{r\beta}$ | Rotor flux linkage, $\beta$ -axis (single-phase model)   | Wb                |
| $\Psi_s$        | Stator flux linkage (general)                            | Wb                |
| $\Psi_{sa}$     | Stator flux linkage, $\alpha$ -axis (single-phase model) | Wb                |
| $\Psi_{s\beta}$ | Stator flux linkage, $\beta$ -axis (single-phase model)  | Wb                |

|                |                                      |       |
|----------------|--------------------------------------|-------|
| $\omega_s$     | Stator angular supply frequency      | rad/s |
| $\Omega_g$     | Slip angular frequency               | rad/s |
| $\Omega_r$     | Mechanical rotor angular speed       | rad/s |
| $\Omega_{ref}$ | Reference angular speed              | rad/s |
| $\Omega_s$     | Synchronous electrical angular speed | rad/s |

### Acronyms and Abbreviations

|                 |                                                               |      |
|-----------------|---------------------------------------------------------------|------|
| <b>AC</b>       | Alternating Current                                           | —    |
| <b>ANFIS</b>    | Adaptive Neuro-Fuzzy Inference System                         | —    |
| <b>AW</b>       | Auxiliary Winding                                             | —    |
| <b>DC</b>       | Direct Current                                                | —    |
| <b>DSP</b>      | Digital Signal Processor                                      | —    |
| <b>EMF</b>      | Electromotive Force                                           | —    |
| <b>FIS</b>      | Fuzzy Inference System                                        | —    |
| <b>FLC</b>      | Fuzzy Logic Controller                                        | —    |
| <b>FOC</b>      | Field-Oriented Control (Vector Control)                       | —    |
| <b>GTO</b>      | Gate Turn-Off Thyristor                                       | —    |
| <b>IAE</b>      | Integral of Absolute Error                                    | —    |
| <b>IGBT</b>     | Insulated Gate Bipolar Transistor                             | —    |
| <b>MAMCP</b>    | Permanent Capacitor Single-Phase Induction Motor              | —    |
| <b>MAS</b>      | Single-Phase Asynchronous Motor (Moteur Asynchrone Monophasé) | —    |
| <b>MATLAB</b>   | Matrix Laboratory (simulation software)                       | —    |
| <b>MCC</b>      | DC Motor (Moteur à Courant Continu)                           | —    |
| <b>MMF</b>      | Magnetomotive Force                                           | —    |
| <b>MOSFET</b>   | Metal-Oxide-Semiconductor Field-Effect Transistor             | —    |
| <b>MW</b>       | Main Winding                                                  | —    |
| <b>PI</b>       | Proportional-Integral (controller)                            | —    |
| <b>PID</b>      | Proportional-Integral-Derivative (controller)                 | —    |
| <b>PWM</b>      | Pulse Width Modulation                                        | —    |
| <b>RMS</b>      | Root Mean Square                                              | —    |
| <b>Simulink</b> | Simulation and Model-Based Design tool (MathWorks)            | —    |
| <b>VFD</b>      | Variable Frequency Drive                                      | —    |
| <b>VSD</b>      | Variable Speed Drive                                          | —    |
| <b>V/f</b>      | Voltage-to-Frequency ratio (scalar control law)               | V/Hz |

# General Introduction

Electrical machines are a major component in present industrial systems and many daily life applications. They are very common because they can change electrical energy to mechanical motion efficiently, quickly and with high controllability. Electric motors are used in many fields, from heavy industrial applications such as conveyor belts and rolling mills to domestic and consumer equipment such as washing machines and printers. Their importance has increased further in recent years with the development of electric vehicles, locomotives and other systems related to sustainable energy [1].

In electric drive systems, the motor is not employed alone. Typically, a complete drive consists of a power source, power electronic converters, an electric motor, a mechanical load, sensors and a controller. The controller's role is to control the electrical quantities supplied to the motor to get the desired mechanical behaviour, mainly speed and torque. Thus, speed control is one of the most important tasks in electric motor drives, particularly when the motor has to operate under varying load conditions or over a wide speed range [2].

The speed of an electric motor is controlled by varying the electrical power supplied to the armature and if necessary the field excitation. This can be done in DC motor drives by DC-DC converters and PWM control. In AC motor drives like induction motors and permanent magnet motors, inverters are utilized to generate variable frequency and variable amplitude voltages. These converters provide the control of the motor in the constant torque region and also in the high speed region where flux weakening techniques may be required [3].

Designing control systems for motors using classical control methods such as PI and IP controllers is common because of their simple structure and practical implementation. However, their performance is highly sensitive to the motor parameters, operating conditions and controller tuning. Classical control can lead to overshoot, slow response, and reduced robustness in some cases when the load or the machine parameters change. Therefore, there is a great interest in the development of more advanced control strategies that can improve the dynamic response and make the drive system more reliable under different operating conditions [4].

Like the wind turbines, the electric motor also needs for its control a clear mathematical model of the motor and its converter. The model is generally formulated from electrical equations, electromechanical conversion equations, and mechanical equations. Clarke and Park transformations are useful for AC machines because they transform three-phase AC quantities into equivalent quantities in a rotating dq reference frame. This reduces the complexity of the GENERAL INTRODUCTION 2 control problem and enables one to treat AC motors in the same manner as DC motors where torque and flux can be controlled independently [2].

The purpose of this thesis, “Study and Design of an Intelligent Speed Controller for an Electric Motor,” is to investigate the principles of electric motor speed control and to design a control system capable of improving the speed response of the motor. The work is based on the general structure of the electric drive systems, mathematical modeling of the electric machines and development of appropriate control strategies. Special focus is put on the relation between the motor model and the controller structure, because a good model is needed to achieve good control performance.

The methodology followed in this work consists first of presenting the basic principles of electric motors and their control systems. Then, mathematical models of representative electric machines are studied including DC motors and AC motors. Power electronic converters like choppers and inverters are considered as they act as the interface between the power source and the motor. Then the control structure is designed using principles of feedback control, PWM generation, current control, speed control and flux weakening when high-speed operation is required. The performance of the proposed control system is evaluated by means of numerical simulation using models that can be implemented in MATLAB/Simulink.

The thesis is organized in several chapters. The first chapter covers the general theory of electric machines and basic principles of motor drives. Chapter 2 describes the modeling of the main components of the drive: motor, converter and mechanical load. The following chapter is dedicated to the design of the speed control system and to the control strategy to be used for torque and flux regulation. The results obtained under different operating conditions such as normal speed operation and high speed operation with flux weakening are shown and discussed in the simulation chapter. Finally, thesis ends with a general conclusion summarizing the main results and suggesting possible future improvements for advanced motor control .

**CHAPTER I :**  
**FUNDAMENTALS OF ELECTRICAL**  
**MACHINES AND THEIR CONTROL**  
**METHODS**

## I. INTRODUCTION

Electric motors are an essential component of modern industrial systems, converting electrical energy into the mechanical motion required for many processes. Their application ranges from simple domestic appliances to complex industrial drives where speed, torque, reliability and energy efficiency are important performance requirements.

Electric motors operate on a simple principle. When electricity flows through a conductor in a magnetic field, there is an interaction between the magnetic field and the electrical current. This interaction creates a mechanical force on the conductor which results in rotation. This principle is valid for both direct current and alternating current machines.

Control methods are required in motor drive systems for speed control, enhanced dynamic response, and stable operation under different load conditions. The PI and PID controllers are among the most established control approaches and are still widely adopted due to their straightforward design and satisfactory performance in most linear operating conditions. But nonlinear or uncertain operating conditions are increasingly addressed by intelligent control methods that enhance robustness and adaptability.

### I.1 Overview of Electric Motors

Electric motors are electromechanical devices whose primary function is the conversion of electrical energy into mechanical energy, specifically in the form of rotational motion. It is the process of converting electrical energy to mechanical energy. This is done by using a magnetic field to interact with conductors that have electrical current flowing through them. This interaction produces a mechanical force.

Electric motors are essential for industrial and household systems due to the high energy efficiency, operational reliability, and ease of speed regulation. These are key components in systems like pumps, fans and manufacturing machinery.

Electric motors are generally classified into two main groups depending on the type of power supply: direct current motors (DC) and alternating current motors (AC). They differ in technical characteristics and have their own scope of industrial or domestic applications.

#### I.1.1 Classification of Electric Motors

There are two main types of electric motors . DC motors and AC motors.

- DC motors use direct current power . There are a few types of DC motors , including brushed and brushless . The major application of these are in drive systems requiring accurate and easy adjustable speed regulation.
- AC motors run on alternating current. They are mainly classified into Single phase and Three phase motors:
  - Single phase motors are widely used in low power domestic appliances while three-phase motors are preferred in industrial drives due to their higher efficiency and better output power.

- Among AC motors , the most popular are induction motors , particularly squirrel cage motors , owing to their ruggedness and low maintenance . Synchronous motors are used for constant speed applications [2].

### **I.1.2 Industrial Applications**

This combination of efficiency, reliability and continuous duty has made electric motors a staple in a diverse range of industrial applications.

Typical applications include pumps, ventilation fans, air compressors and material handling conveyors. In manufacturing, electric motors are the heart of automated production lines, making sure the processes keep running smoothly and around the clock.

Heavy-duty applications in the oil and gas, mining and transportation industries also depend heavily on electric motors.

There are several factors to consider when selecting a motor such as power requirements, speed control, efficiency and load conditions.

## **I.2 DC Motor Fundamentals**

A DC motor is an electromechanical device that converts DC electrical power into mechanical rotational output. This type of motor is widely used in applications requiring accurate speed regulation and good torque response. The choice of a particular DC motor configuration depends on the means of excitation and the performance required. These configurations are separately excited, shunt, series and compound D.C. motors. Each type has specific characteristics suitable for specific industrial applications [3].

### **I.2.1 Structure of DC Motor**

The DC motor is mainly made up of stationary magnetic part called stator and a rotating part called rotor. The main magnetic field is produced by the stator, and the rotor contains the armature winding where the electromechanical energy conversion takes place.

The commutator and brushes enable the direction of the current in the armature conductors to be reversed as the armature rotates. This ensures the developed electromagnetic torque is in the same direction which allows the motor to rotate continuously [6].

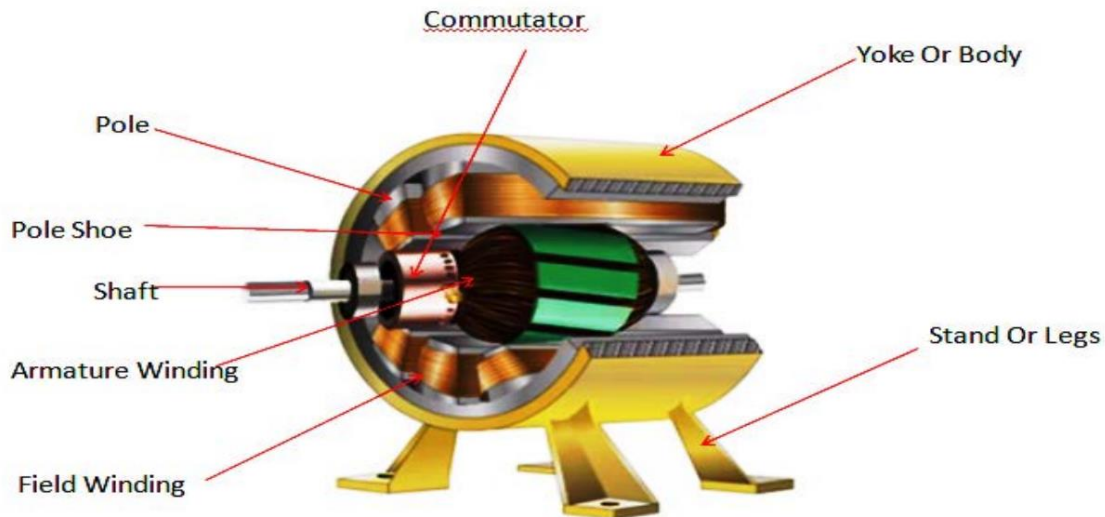


**Fig. I.1:** Structure of a DC Motor [6].

### I.2.2 Operating Principle

A DC motor works on the principle that a mechanical force is exerted on a current carrying conductor placed in a magnetic field. By using Fleming's left hand rule, the direction of this force can be found. This rule states that the field, the current and the force are at right angles to each other.

As the existing passes through the armature conductors, electromagnetic forces are generated on the conductors underneath the magnetic poles. These forces create an electromagnetic torque which turns the rotor. The commutator switches the current direction at the right moment, keeping the torque continuous in the same rotational direction.

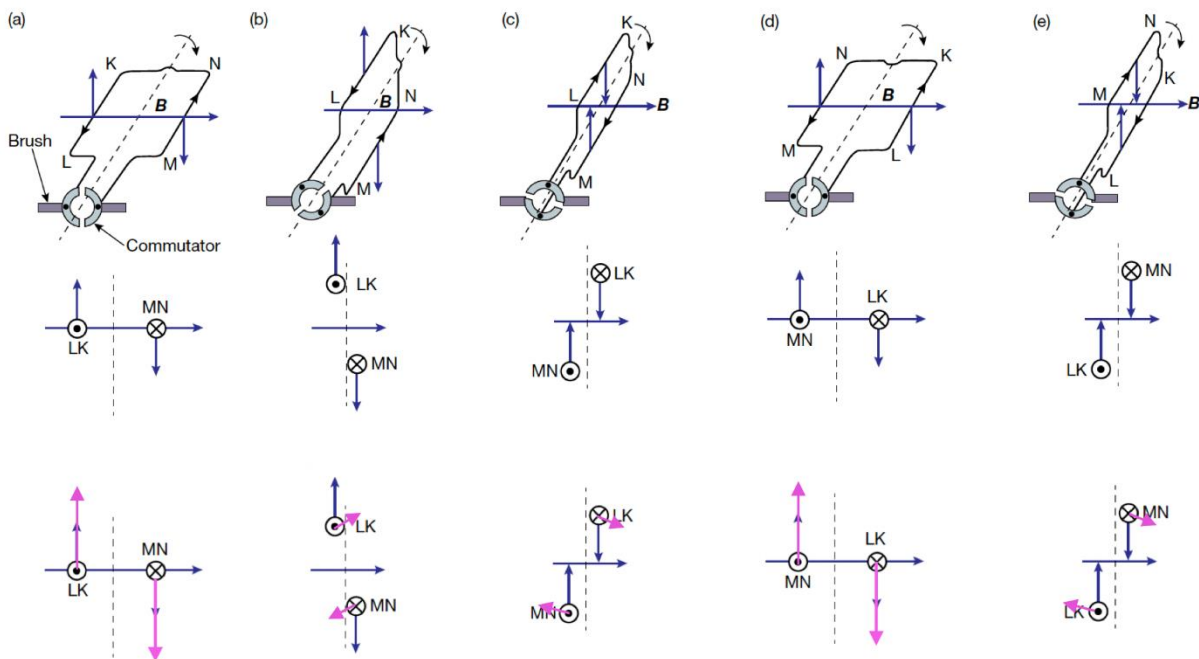


**Fig. I.2:** Operating principle of a DC motor showing the torque production mechanism [14].

### I.2.3 Advantages and Limitations

The major advantage that the dc motors have over the other motors is that they can be controlled over a wide range of speeds, simply by varying the armature voltage. They can also produce a high starting torque, so they are well suited to drives that need rapid acceleration or precise speed control.

These advantages are not without several disadvantages of DC motors. The brush-commutator assembly is a source of mechanical wear, with periodic servicing, which limits the reliability of DC motors in harsh, dusty or hazardous atmospheres. Therefore, AC motors are often used in modern industrial applications that require robustness and low maintenance.

### I.3 Mathematical Modeling of DC Motor

The mathematical modeling of the DC motor gives a formal description of the electrical dynamics and mechanical behavior of the motor in a form of a system of governing equations. This model is a necessary tool for performance analysis and forms the basis for designing suitable control strategies.

This model is derived from the Kirchhoff voltage law for the electrical circuits and the Newton's second law of motion for the mechanical part. The back electromotive force and the electromagnetic torque are included explicitly, since they are the coupling between electrical and mechanical subsystems [7].

#### I.3.1 Modeling of the Separately Excited DC Motor

The separately excited DC motor is fed by two independent voltage source . The field circuit is fed by the excitation voltage and the armature circuit is fed by the armature voltage . This arrangement permits the magnetic excitation and the armature current to be controlled independently of each other.

In this work, the mathematical model of the separately excited dc motor is obtained from the field circuit equation, armature circuit equation, electromagnetic torque equation and the mechanical equation of motion. The model is then implemented in MATLAB/Simulink to analyze the response of motor speed and electromagnetic torque.

##### I.3.1.1 Electrical equations

For the field circuit, the voltage equation is written as:

$$U_f = R_f I_f + L_f \frac{dI_f}{dt} \quad (I. 1)$$

For the armature circuit, the applied voltage is equal to the resistive voltage drop, the inductive voltage drop, and the back electromotive force. Therefore:

$$U_a = R_a I_a + L_a \frac{dI_a}{dt} + e \quad (I. 2)$$

The back electromotive force is proportional to the excitation current and the rotor speed:

$$e = M_{fd} I_f \Omega_r \quad (I.3)$$

where ( $U_f$ ) is the field voltage, ( $U_a$ ) is the armature voltage, ( $I_f$ ) is the field current, ( $I_a$ ) is the armature current, ( $R_f$ ) and ( $R_a$ ) are the field and armature resistances, ( $L_f$ ) and ( $L_a$ ) are the field and armature inductances, ( $M_{fd}$ ) is the mutual inductance between the field and armature circuits, ( $e$ ) is the back electromotive force, and ( $\Omega_r$ ) is the mechanical angular speed.

### I.3.1.2 Electromagnetic torque

The electromagnetic torque developed by the separately excited DC motor is directly proportional to the field current and the armature current. It is of the form:

$$C_{em} = M_{fd} I_f I_a \quad (I.4)$$

where ( $C_{em}$ ) is the electromagnetic torque.

### I.3.1.3 Mechanical equation

The formula for the mechanical behavior of the separately excited DC motor is derived from the Newton's law of rotational motion. The difference between electromagnetic torque and load torque is used to accelerate the rotor and overcome the viscous friction.

The mechanical equation is written as:

$$J \frac{d\Omega_r}{dt} = C_{em} - C_r - f\Omega_r \quad (I.5)$$

where ( $J$ ) is the moment of inertia, ( $\Omega_r$ ) is the mechanical angular speed, ( $C_{em}$ ) is the electromagnetic torque, ( $C_r$ ) is the load torque, and ( $f$ ) is the viscous friction coefficient.

By dividing both sides by ( $J$ ), the angular acceleration is obtained as:

$$\frac{d\Omega_r}{dt} = \frac{1}{J} (C_{em} - C_r - f\Omega_r) \quad (I.6)$$

In some parts of the simulation model, the electrical angular speed is denoted by  $\Omega_s$ . The relation between the electrical angular speed and the mechanical rotor speed is:

$$\Omega_s = P\Omega_r$$

Therefore, the mechanical rotor speed can be expressed as:

$$\Omega_r = \Omega_s / P \quad (I.7)$$

where ( $P$ ) is the number of pole pairs.

This relation explains the use of the gain ( $P/J$ ) in the Simulink model. The electromagnetic torque and load torque are first used to obtain the angular acceleration. Then, an integrator block is used to calculate the rotor speed. The output of this integrator gives the mechanical rotor speed ( $\Omega_r$ ), while the electrical speed can be obtained from ( $\Omega_s = P \cdot \Omega_r$ ).

### I.3.1.4 Complete state-space form of the DC motor model

For a separately excited DC motor, the electrical equations can be expressed in matrix form by taking field current and armature current as state variables. This form is useful for simulation as it separates the voltage inputs, the resistive voltage drops and the speed dependent coupling term.

Definition of the state and input vectors

Definition of the matrices

The current vector is:  $[I] = \begin{bmatrix} I_f \\ I_a \end{bmatrix}$

The voltage vector is:  $[U] = \begin{bmatrix} U_f \\ U_a \end{bmatrix}$

The inductance matrix is:  $[L] = \begin{bmatrix} L_f & 0 \\ 0 & L_a \end{bmatrix}$

The resistance matrix is:  $[R_1] = \begin{bmatrix} R_f & 0 \\ 0 & R_a \end{bmatrix}$

The speed-coupling matrix is:  $[R_2] = \begin{bmatrix} 0 & 0 \\ M_{fd} & 0 \end{bmatrix}$

Compact matrix equation

Using these definitions, the electrical model is written as:

$$\begin{aligned} U_f &= R_f I_f + L_f \frac{dI_f}{dt} \\ U_a &= R_a I_a + \frac{L_a dI_a}{dt} + \Omega_r M_{fd} I_f \\ C_{em} &= M_{fd} I_a \cdot I_f \end{aligned} \quad (I.8)$$

The electrical model can then be written in matrix form as:

$$[L] \frac{d[I]}{dt} = [U] - [R_1][I] - \Omega_r [R_2][I] \quad (I.9)$$

Expanded matrix form

In expanded form, the same equation is:

$$\begin{bmatrix} L_f & 0 \\ 0 & L_a \end{bmatrix} \begin{bmatrix} I_f \\ I_a \end{bmatrix}^* = \begin{bmatrix} U_f \\ U_a \end{bmatrix} - \begin{bmatrix} R_f & 0 \\ \Omega_r M_{fd} & R_a \end{bmatrix} \begin{bmatrix} I_f \\ I_a \end{bmatrix} \quad (I.10)$$

This matrix can be divided into two parts: a constant resistance matrix and a speed-dependent coupling matrix. This separation is useful since the back electromotive force term is dependent on the rotor speed. Thus the global electrical resistance matrix can be expressed as:

$$\begin{bmatrix} R_f & 0 \\ \Omega_r M_{fd} & R_a \end{bmatrix} = \Omega_r \begin{bmatrix} 0 & 0 \\ M_{fd} & 0 \end{bmatrix} + \begin{bmatrix} R_f & 0 \\ 0 & R_a \end{bmatrix} \quad (I.11)$$

The mechanical behavior of the motor is modeled by using Newton's second law on the rotating shaft. The load torque works against the electromagnetic torque generated by the motor. Thus, the change of the rotor speed can be expressed as:

In the Simulink implementation, the mechanical equation is expressed in terms of the electrical angular speed  $\Omega_s = P \cdot \Omega_r$ , which introduces the factor  $P$  in the integrator gain. The equation governing the electrical speed is therefore:

$$\frac{d\Omega_s}{dt} = \frac{P}{J} (M_{fd} I_f I_a - C_r - f \Omega_r) \quad (I.12)$$

where the mechanical rotor speed is recovered as  $\Omega_r = \Omega_s / P$ .

### I.3.1.5 MATLAB/Simulink implementation and simulation results

The mathematical model developed earlier was implemented in MATLAB/Simulink and used to study the dynamic behavior of the separately excited DC motor. The model is built on the basis of the electrical matrix equation, electromagnetic torque and mechanical motion equation.

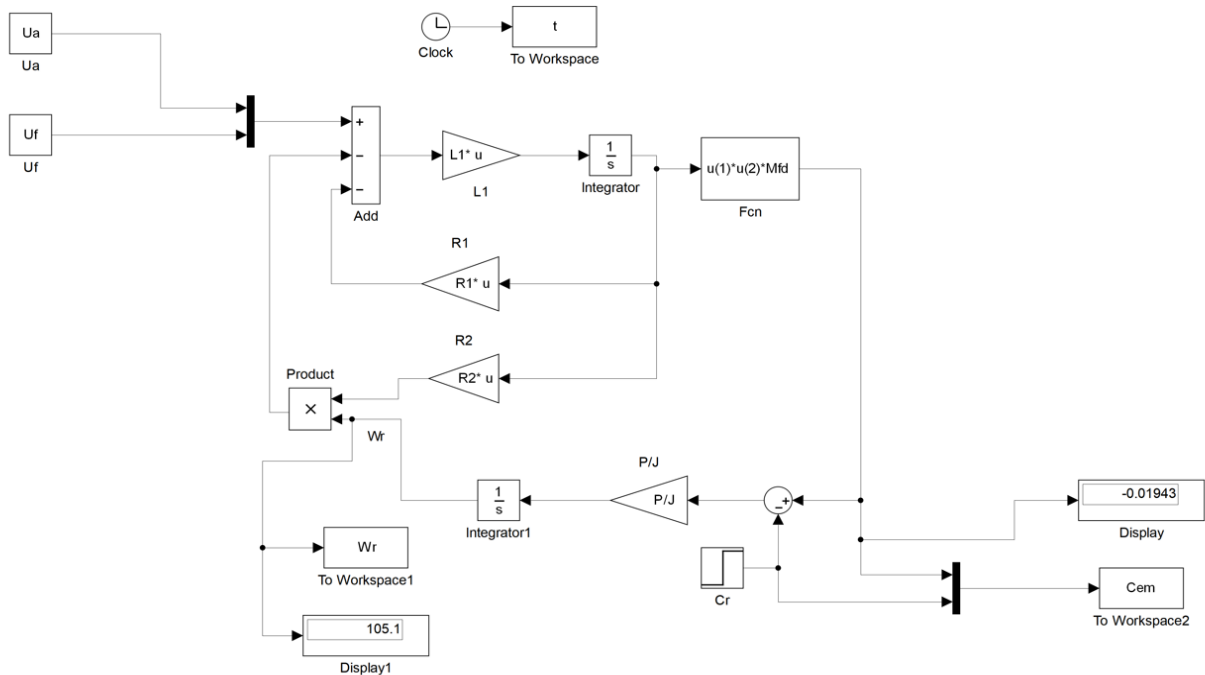


**Fig. I.3:** MATLAB/Simulink model of the separately excited DC motor.

Fig. I.3 shows the MATLAB/Simulink implementation of the separately excited DC motor. The model uses the voltage inputs ( $U_a$ ) and ( $U_f$ ), the resistance and inductance matrices, and the coupling term related to the rotor speed ( $\Omega_r$ ). The current vector is obtained using the inverse inductance matrix and an integrator block. The electromagnetic torque is then calculated from the field current and armature current, while the mechanical part gives the rotor speed.

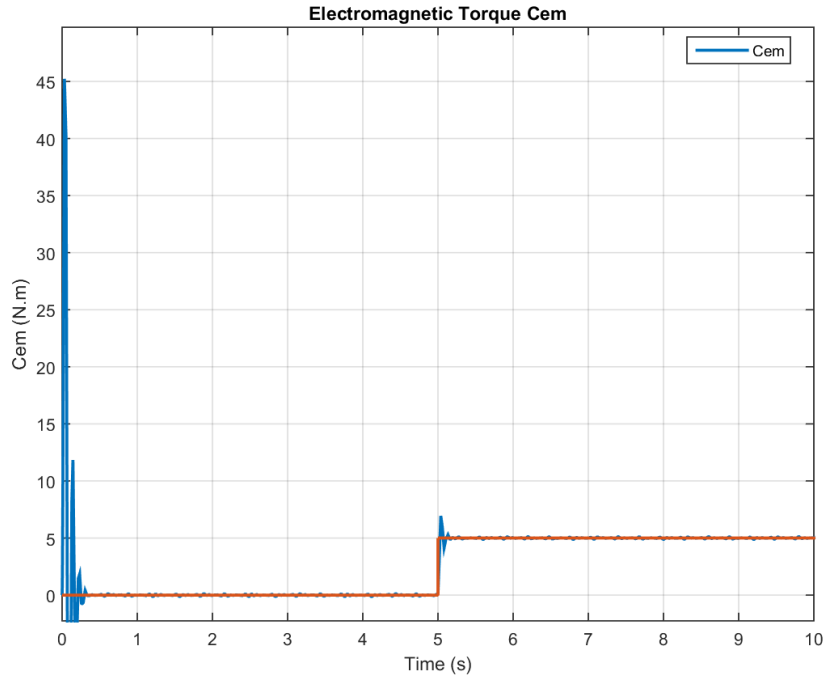


**Fig. I.4:** Electromagnetic torque response of the separately excited DC motor.

Figure I.4 shows the electromagnetic torque response of the motor. At the beginning of the simulation, the torque reaches a high transient value due to the starting conditions. After this short transient period, it decreases and stays close to zero during no-load operation. At  $t = 5$  s, a load torque of  $C_r = 5$  N·m is applied. As a result, the electromagnetic torque increases rapidly to compensate for the applied mechanical load. After a brief transient, it stabilizes at approximately 5 N·m, which is equal to the applied load torque, showing that the motor has reached a new steady-state operating point while remaining stable.

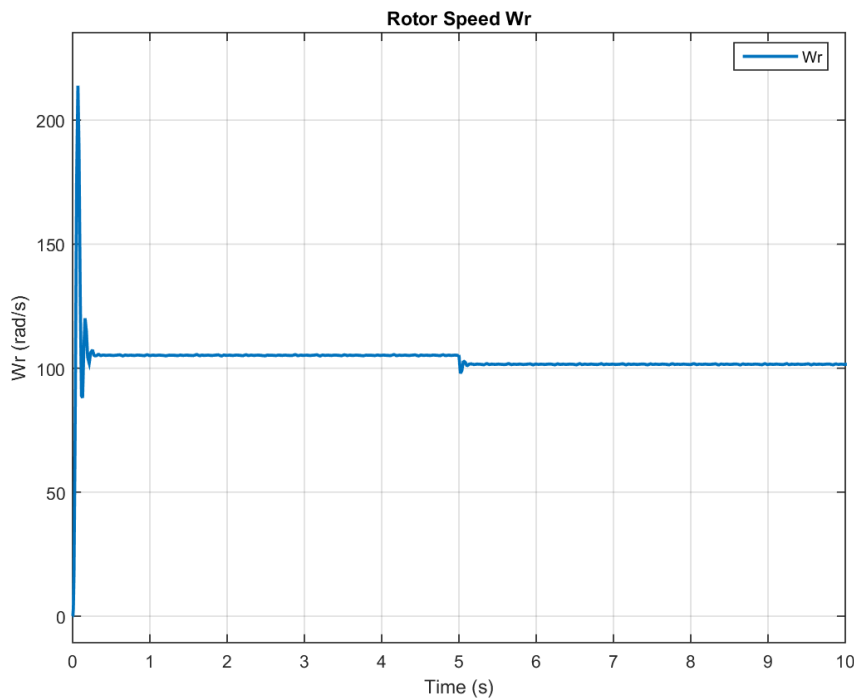


**Fig. I.5:** Rotor speed response of the separately excited DC motor.

Fig. I.5 shows rotor speed response. The velocity increases rapidly in the starting period and reaches a steady-state value. When the load torque is applied at  $t=5$  s, a speed drop happens, then the motor stabilizes again. This kind of behavior confirms that the model predicts the dynamic response of the separately excited DC motor correctly. Simulation results demonstrate that the mathematical model is capable to accurately describe the main operating behavior of the separately excited DC motor, including starting transient, electromagnetic torque production, application of load torque and variation of rotor speed.

## I.4 Induction Motor Fundamentals

Induction motors are one of the most widely used electrical machines in industry due to their rugged and simple construction, reliable performance over a long period and absence of any brush or commutator mechanism, which reduces servicing needs considerably and makes them suitable for prolonged operation. Figure I.6 shows the construction of an induction motor. The torque of the induction motor is produced by the principle of electromagnetic induction. When an alternating voltage is applied to the stator winding a rotating magnetic field at synchronous speed is set up. This rotating field sweeps over the short circuited rotor conductors and induces a voltage and hence a circulating current in the rotor. The interaction between the rotating magnetic field and the induced rotor current creates an electromagnetic torque that rotates the rotor. However, the rotor speed is always slightly less than the synchronous speed. This difference is called slip.



**Fig. I.6:** Construction of an Induction Motor [13]

### I.4.1 Structure of Induction Motor

Induction motor has two main parts; the stator and the rotor. The fixed part of the machine is called the stator. It contains the three phase winding set which is connected to the AC power supply and is the source of the rotating magnetic field that turns the motor. The rotor is the moving part of the machine. The rotor is mounted concentrically inside the stator bore. The winding or bar structure is designed to respond to the stator's rotating field and develop the electromagnetic torque which rotates the mechanical load.

Induction motor rotors are generally classified into two main types, as shown in Fig. I.7:

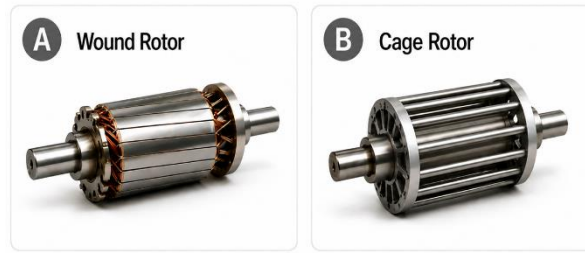


**Fig. I.7:** Types of Rotor: (A) Wound Rotor (B) Squirrel Cage Rotor

The most commonly used is the squirrel cage rotor. Its construction is simple and rugged. It consists of a laminated cylindrical core with aluminium or copper conducting bars. These bars are short-circuited at both ends by end rings and thus form a closed circuit. The wound rotor (aka slip-ring rotor) has three-phase windings similar to the stator. These windings are connected through slip rings, so that an external resistance can be added in the rotor circuit. This increases the starting torque and provides better control of the motor speed but increases the complexity and cost [9].

### I.4.2 Operating Principle

The induction motor is based on the principle of electromagnetic induction. A magnetic field is produced when the stator is connected to a balanced three-phase supply which rotates continuously in the machine at synchronous speed. This field passes the conductors of the rotor and induces an EMF in them by electromagnetic induction. This induced EMF causes a circulating current to flow in the rotor circuit. The rotor bars are shorted electrically through the end rings. The synchronous speed of the rotating magnetic field is a function of supply frequency and number of poles and is given by:

$$N_s = \frac{60 f}{P} \quad (\text{I.13})$$

Where  $N_s$  is the synchronous speed (rpm),  $f$  is the supply frequency (Hz), and  $P$  is the number of pole pairs. Note that some references express this formula using the total number of poles; here  $P$  denotes pole pairs, consistent with the convention adopted throughout this work.

Therefore, a magnetic field is also created in the rotor. The electromagnetic torque is generated by interaction of the stator rotating magnetic field and the rotor magnetic field, which causes the rotor to rotate in the same direction as the stator field. The rotor runs at a speed that is slightly less than the synchronous speed of the rotating field. slip is the difference in speed and it is a necessary condition for the induction process to happen. If the rotor achieved synchronous speed there would be no relative motion between the stator field and the rotor conductors. Under this condition no EMF would be induced and no rotor current would circulate and the torque production would cease completely. This physical limitation explains why the induction motor is always operated below synchronous speed in practice [10].

### I.4.3 Classification of Induction Motors

Induction motor can be classified on several criteria depending upon their construction and operating characteristics.

Induction motors are classified into squirrel cage and wound rotors according to the rotor construction as described in Section I.4.1.

Another important classification concerns the number of phases of the power supply:

#### ➤ Single-Phase Induction Motor

Single phase induction motor is driven by single phase AC supply. Widely used in household and light duty applications like fans, washing machines and small pumps. These motors provide a compact and low cost design but their efficiency and starting torque are inherently less than three phase induction motors.

#### ➤ Three-Phase Induction Motor

The workhorse of industry is the three-phase induction motor, driven by a balanced three-phase supply. It provides higher efficiency, better steady state operation and higher starting torque than single phase designs. It is the most used type of induction motor in industry, due to its reliability and robustness.

### I.4.4 Modeling of the Permanent Capacitor Single-Phase Induction Motor (MAMCP)

#### I.4.4.1 Operating Principle

The single phase induction motor works on a different principle than the three phase induction motor. The main winding (MW) is fed directly from a single phase network and therefore itself cannot create a rotating magnetic field. But it produces a pulsating magnetic field which can be mathematically divided into two rotating fields of equal amplitude, rotating at the same speed but in opposite directions, one in the forward direction, in line with the rotor rotation, and the other in the reverse direction.

In the motor configuration studied here, both the main winding (MW) and the auxiliary winding (AW) are placed along two orthogonal axes. The rotor winding consists of two identical short-circuited windings, offset 90 electrical degrees from each other, as in the conventional three-phase asynchronous motor. This representation allows the motor to be analyzed in a unified electromagnetic framework.

The reference frame adopted for the study of the electromagnetic processes within the machine is the stationary  $\alpha$ - $\beta$  system, fixed to the stator. In this frame, it is assumed that the magnetomotive force (M.M.F.) developed by each stator phase has a sinusoidal spatial distribution along the rotor circumference. This assumption simplifies the mathematical formulation while preserving the

essential dynamic behavior of the machine, and forms the basis for the state-space model developed in the following section [21].

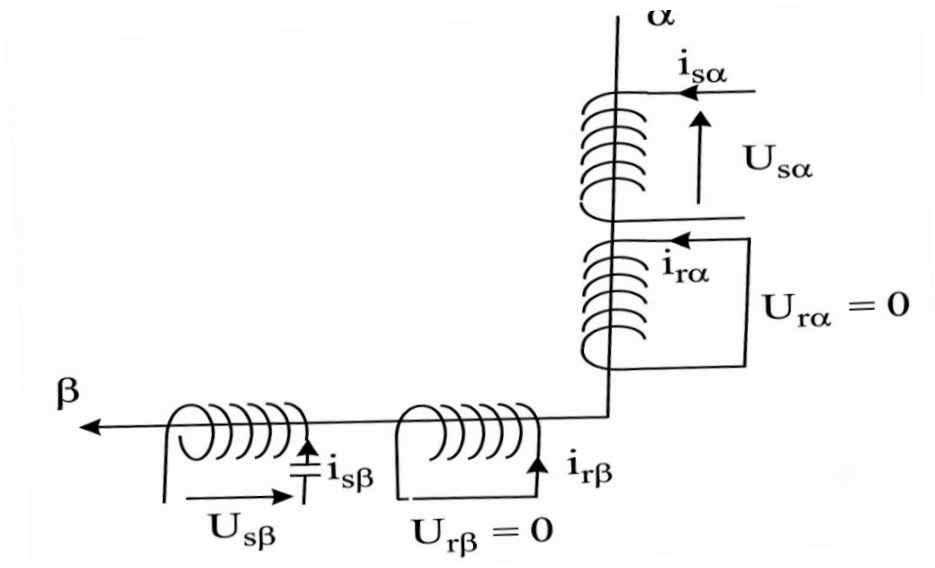


**Fig. I.8:** Representation of the induction machine with capacitor in the  $\alpha$ - $\beta$  reference frame [21].

#### I.4.4.2 Voltage Equation System

The voltage equation system of the single-phase induction motor in the stationary  $\alpha$ - $\beta$  reference frame is given by [20]

$$\begin{aligned}
 U_{s\alpha} &= r_{s\alpha} I_{s\alpha} + \frac{d\Psi_{s\alpha}}{dt} \\
 U_{s\beta} &= r_{s\beta} I_{s\beta} + \frac{d\Psi_{s\beta}}{dt} \\
 U_{r\alpha} &= 0 = r_{r\alpha} I_{r\alpha} + \frac{d\Psi_{r\alpha}}{dt} + \Omega_s \Psi_{r\beta} \\
 U_{r\beta} &= 0 = r_{r\beta} I_{r\beta} + \frac{d\Psi_{r\beta}}{dt} - \Omega_s \Psi_{r\alpha} \\
 0 &= -I_{s\beta} + C \frac{dU_C}{dt} \quad (\text{capacitor equation})
 \end{aligned} \tag{I.14}$$

Where:

$r_{s\alpha}, r_{s\beta}$  : main and auxiliary winding resistances

$r_{r\alpha}, r_{r\beta}$  : rotor resistances (referred to the stator)

$\Omega_s = P \cdot \Omega_r$ : synchronous electrical angular frequency (stator angular frequency)

$U_C$  : voltage across the series capacitor  $C$  of the auxiliary winding

#### I.4.4.3 Flux equations

The flux equations of the machine are expressed as follows[20]:

$$\begin{aligned}
 \Psi_{s\alpha} &= L_{s\alpha} \cdot I_{s\alpha} + M_{\alpha} \cdot I_{r\alpha} \\
 \Psi_{s\beta} &= L_{s\beta} \cdot I_{s\beta} + M_{\beta} \cdot I_{r\beta} \\
 \Psi_{r\alpha} &= L_{r\alpha} \cdot I_{r\alpha} + M_{\alpha} \cdot I_{s\alpha} \\
 \Psi_{r\beta} &= L_{r\beta} \cdot I_{r\beta} + M_{\beta} \cdot I_{s\beta}
 \end{aligned} \tag{I.15}$$

where :

$L_{s\alpha}, L_{s\beta}$  : Stator winding self-inductances of the main and auxiliary windings

$L_{r\alpha}, L_{r\beta}$ : Rotor inductances referred to the main winding turns

$M_{\alpha}$  : Magnetizing inductance of the M.W

$M_{\beta}$ : Magnetizing inductance of the auxiliary winding

By substituting the flux equations into the voltage equations, we obtain:

$$\begin{aligned}
 U_{s\alpha} &= r_{s\alpha} \cdot I_{s\alpha} + L_{s\alpha} \cdot \frac{dI_{s\alpha}}{dt} + M_{\alpha} \cdot \frac{dI_{r\alpha}}{dt} \\
 U_{s\beta} &= r_{s\beta} \cdot I_{s\beta} + L_{s\beta} \cdot \frac{dI_{s\beta}}{dt} + M_{\beta} \cdot \frac{dI_{r\beta}}{dt} + U_c \\
 U_{r\alpha} &= r_{r\alpha} \cdot I_{r\alpha} + L_{r\alpha} \cdot \frac{dI_{r\alpha}}{dt} + M_{\alpha} \cdot \frac{dI_{s\alpha}}{dt} + \Omega_s \cdot L_{r\beta} \cdot I_{r\beta} + \Omega_s \cdot M_{\beta} \cdot I_{s\beta} \\
 U_{r\beta} &= r_{r\beta} \cdot I_{r\beta} + L_{r\beta} \cdot \frac{dI_{r\beta}}{dt} + M_{\beta} \cdot \frac{dI_{s\beta}}{dt} - \Omega_s \cdot L_{r\alpha} \cdot I_{r\alpha} - \Omega_s \cdot M_{\alpha} \cdot I_{s\alpha} \\
 0 &= -I_{s\beta} + C \cdot \frac{dU_c}{dt}
 \end{aligned} \tag{I.16}$$

The last equation represents the voltage across the capacitor connected in series with the auxiliary winding. This capacitor serves as a starting device. This system of equations can be written in the form

$$[U] = [R] \cdot [I] + [L_{\alpha\beta}] \cdot \frac{d[I]}{dt} \tag{I.17}$$

Where:  $[U]$  is the voltage vector in the  $\alpha$ - $\beta$  reference frame

$$\begin{aligned}
 \begin{bmatrix} U_{s\alpha} \\ U_{s\beta} \\ U_{r\alpha} \\ U_{r\beta} \\ 0 \end{bmatrix} &= \begin{bmatrix} r_{s\alpha} & 0 & 0 & 0 & 0 \\ 0 & r_{s\beta} & 0 & 0 & 0 \\ 0 & (1/a) \cdot M_\beta \cdot \Omega_s & r_{r\alpha} & (1/a) \cdot L_{r\beta} \cdot \Omega_s & 0 \\ -a \cdot M_\alpha \cdot \Omega_s & 0 & -a \cdot L_{r\alpha} \cdot \Omega_s & r_{r\beta} & 0 \\ 0 & 0 & 0 & 0 & 0 \end{bmatrix} \times \begin{bmatrix} I_{s\alpha} \\ I_{s\beta} \\ I_{r\alpha} \\ I_{r\beta} \\ U_c \end{bmatrix} \\
 + \begin{bmatrix} L_{s\alpha} & 0 & M_\alpha & 0 & 0 \\ 0 & L_{s\beta} & 0 & M_\beta & 0 \\ M_\alpha & 0 & L_{r\alpha} & 0 & 0 \\ 0 & M_\beta & 0 & L_{r\beta} & 0 \\ 0 & 0 & 0 & 0 & C \end{bmatrix} \times \begin{bmatrix} I_{s\alpha} \\ I_{s\beta} \\ I_{r\alpha} \\ I_{r\beta} \\ U_c \end{bmatrix} \frac{d}{dt} & \quad (I.18)
 \end{aligned}$$

With :  $[X]^* = d[X]/dt$

where the transformation coefficient  $a$  is defined by:

$$a = \frac{N_a K_{ba}}{N_p K_{pb}}$$

where :

$N_p$  : Actual number of turns of the main winding

$N_a$  : Actual number of turns of the auxiliary winding

$K_{pb}$  : Winding factor of the main winding

$K_{ba}$  : Winding factor of the auxiliary winding

Where :

$C$  : Capacitance of the capacitor in series with the auxiliary winding .

#### I.4.4.4 Electromagnetic Torque :

The electromagnetic torque for the induction machine is given by

$$C_e(\alpha\beta) = P [M_\alpha \cdot a \cdot (I_{s\alpha} + I_{r\alpha}) \cdot I_{r\beta} - (M_\beta / a) \cdot (I_{s\beta} + I_{r\beta}) \cdot I_{r\alpha}] \quad (I.19)$$

The factor  $a$  appears asymmetrically in the two axis terms because the two stator axes are referred to different

#### I.4.4.5 Mechanical Equation of Motion

$$J \frac{d\Omega_r}{dt} = C_{em} - C_r - f\Omega_r \quad (I.20)$$

#### I.4.4.6 Simulation and Results

Based on the state-space formulation derived in the previous section, the mathematical model of single phase induction motor was implemented in MATLAB/Simulink. The simulation was performed for 10 s, with a load torque  $C_r = 1 \text{ N}\cdot\text{m}$  applied at  $t = 5 \text{ s}$ .

The Simulink model shown in Fig. I.9 implements the state-space equations using matrix gain blocks, an integrator, and a nonlinear torque function (Fcn block). The input voltages  $U_{sa}$  and  $U_{sb}$  are generated by two sinusoidal sources, with the auxiliary winding connected through a capacitor. The rotor speed is fed back through a gain block to compute the speed-dependent coupling terms, and the results are logged to the MATLAB workspace using "To Workspace" blocks.

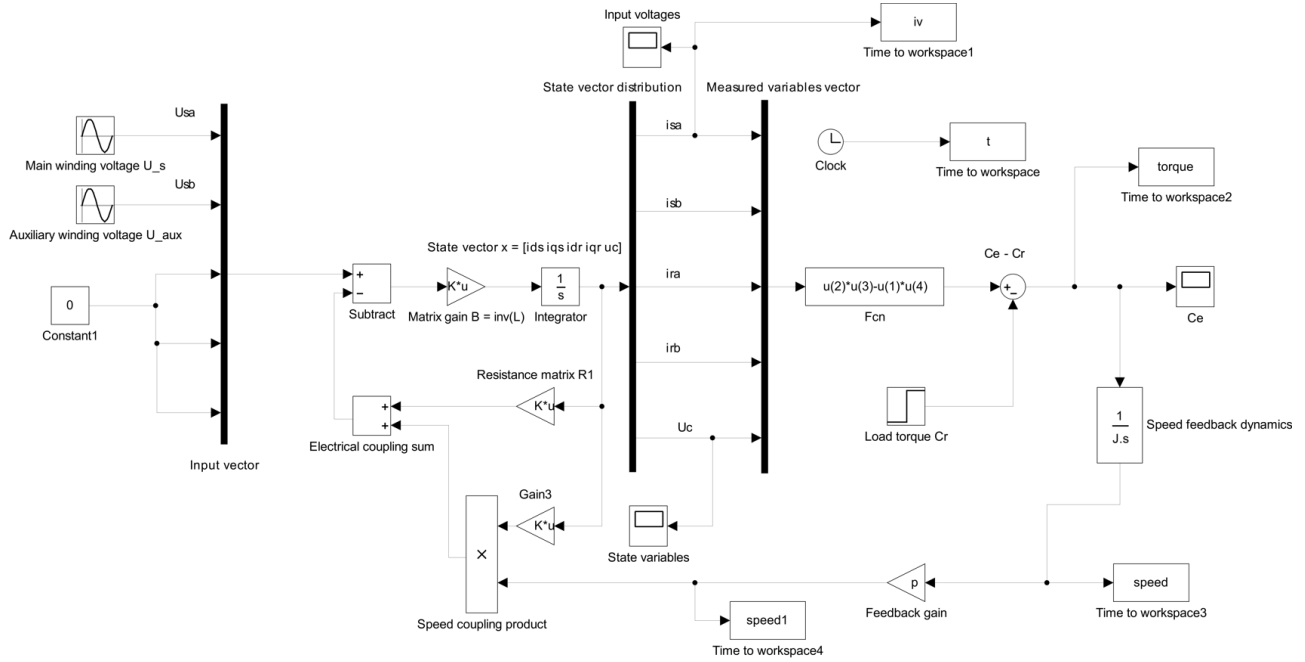


**Fig. I.9:** Simulink model of the single-phase induction motor

### ➤ Rotor Speed Response

The response of the motor rotor speed is given in I.10. The motor goes from standstill to a steady state speed of about 156.1 rad/sec ( $\approx 1491$  RPM) in about 1 sec. The speed response is smooth and has no appreciable overshoot which is a feature of an open loop single phase induction motor under no load conditions. After the load torque is applied at  $t=5$  s, a small and short speed drop is observed, after which the speed goes back to its previous value, which confirms the self-regulating behavior intrinsic to the motor.

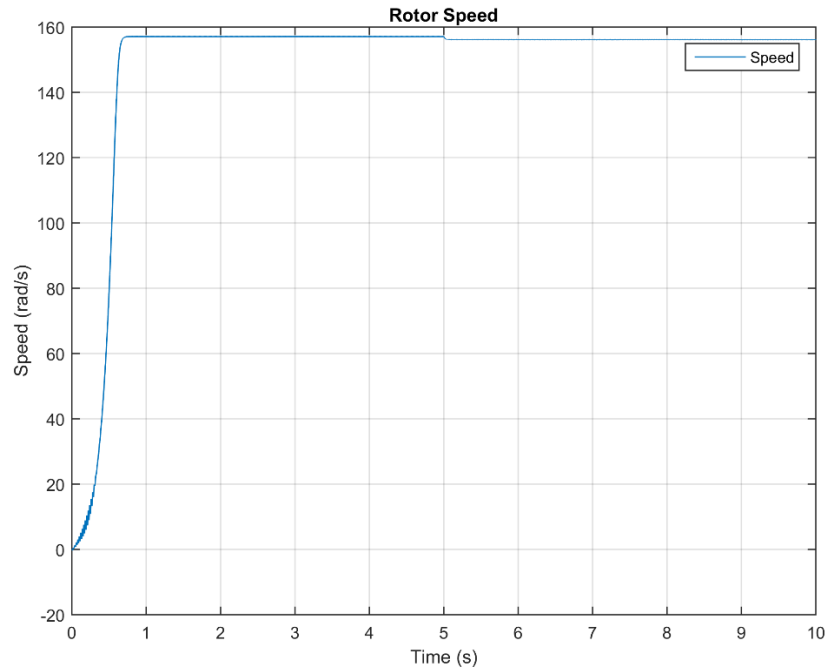


**Fig. I.10:** Rotor speed response of the single-phase induction motor

#### ➤ Electromagnetic Torque Response

As shown in Fig. I.11, the electromagnetic torque presents large oscillations during the starting transient (0 - 1 s) with a maximum value of about 62 N·m. This behaviour is normal at motor start up where the slip is large and the rotor currents produce large pulsating torques. At steady state, the torque approaches a small positive value close to zero for no-load conditions as the motor comes to steady state. At  $t = 5$  s the application of the load torque causes a short negative dip before settling to around 2 N·m, which is the torque needed to overcome the applied load.

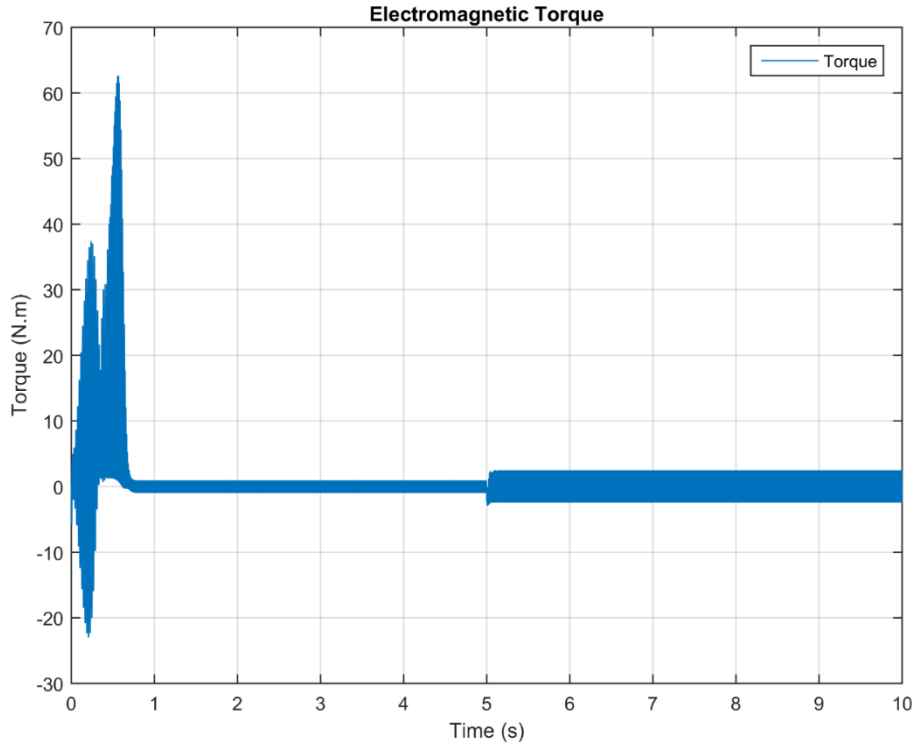


**Fig. I.11:** Electromagnetic torque response of the single-phase induction motor

The simulation results validate the state-space model and its ability to reproduce accurately the expected dynamic behaviour of the single-phase induction motor. The model gives a good basis for the speed controller design presented in the following sections.

#### I.4.5 Modeling of the Three-Phase Induction Machine – Mathematical Model

The mathematical model of the three-phase induction motor is needed to study its dynamic behavior and to prepare its implementation in MATLAB/Simulink. In this work the machine is modeled in the rotating dq reference frame. This model allows the transformation of the three-phase variables into two equivalent components: the direct-axis component d and the quadrature-axis component q.

This representation allows to simplify the analysis of the machine and to study the main electrical and mechanical quantities such as the stator voltages, the stator currents, the rotor speed, the electromagnetic torque and the load torque.

The rotor windings of a squirrel-cage induction motor are short-circuited. Thus the rotor voltages are zero. The dq model is derived from the voltage equations, flux equations, electromagnetic torque equation and mechanical equation.

##### I.4.5.1 Electrical voltage equations in the dq reference frame

The stator and rotor voltage equations in the dq reference frame are given by [21]:

$$\begin{aligned} u_{ds} &= R_s I_{ds} + \frac{d\Psi_{ds}}{dt} - \Omega_s \Psi_{qs} \\ u_{qs} &= R_s I_{qs} + \frac{d\Psi_{qs}}{dt} + \Omega_s \Psi_{ds} \end{aligned} \quad (I. 21)$$

$$u_{dr} = 0 = R_r I_{dr} + \frac{d\Psi_{dr}}{dt} - \Omega_g \Psi_{qr}$$

$$u_{qr} = 0 = R_r I_{qr} + \frac{d\Psi_{qr}}{dt} + \Omega_g \Psi_{dr}$$

with:  $\omega_s = 2\pi f$  and  $\Omega_g = \Omega_s - P\Omega_r$

where :

- $\Psi_{ds}$  : stator flux linkage along the d-axis
- $\Psi_{qs}$  : stator flux linkage along the q-axis
- $\Psi_{dr}$  : rotor flux linkage along the d-axis
- $\Psi_{qr}$  : rotor flux linkage along the q-axis
- $I_{ds}$  : stator current, d-axis component
- $I_{qs}$  : stator current, q-axis component
- $I_{dr}$  : rotor current, d-axis component
- $I_{qr}$  : rotor current, q-axis component
- $\omega_s = 2\pi f$  is the stator supply angular frequency,  $\Omega_g$  is the slip angular frequency, P is the number of pole pairs, and f is the supply frequency.

#### I.4.5.2 Flux equations

The flux equations of the machine are expressed as follows:

$$\begin{aligned}\Psi_{ds} &= L_s I_{ds} + M I_r \\ \Psi_{qs} &= L_s I_{qs} + M I_{qr} \\ \Psi_{dr} &= L_r I_{dr} + M I_{ds} \\ \Psi_{qr} &= L_r I_{qr} + M I_{qs}\end{aligned}\tag{I.22}$$

with:

$$L_s = L_{ls} + M_s$$

$$L_r = L_{lr} + M_r$$

$$M = \frac{3}{2} M_{sr}$$

Where :

- $L_s$  : total stator inductance
- $L_r$  : total rotor inductance
- $M$  : mutual (magnetizing) inductance between stator and rotor
- $L_{ls}$  : stator leakage inductance
- $L_{lr}$  : rotor leakage inductance
- $M_s$ : stator magnetizing inductance
- $M_r$ : rotor magnetizing inductance
- $M_{sr}$ : maximum mutual inductance between one stator phase and one rotor phase

### I.4.5.3 Introduction of flux equations into voltage equations

By introducing the flux equations (I.22) into the voltage equations (I.21), the following system is obtained:

$$\begin{aligned} u_{ds} &= R_s I_{ds} + L_s \frac{dI_{ds}}{dt} + M \frac{dI_{dr}}{dt} - \Omega_s (L_s I_{qs} + M I_{qr}) \\ u_{qs} &= R_s I_{qs} + L_s \frac{dI_{qs}}{dt} + M \frac{dI_{qr}}{dt} + \Omega_s (L_s I_{ds} + M I_{dr}) \\ u_{dr} &= 0 = R_r I_{dr} + L_r \frac{dI_{dr}}{dt} + M \frac{dI_{ds}}{dt} - \Omega_g (L_r I_{qr} + M I_{qs}) \\ u_{qr} &= 0 = R_r I_{qr} + L_r \frac{dI_{qr}}{dt} + M \frac{dI_{qs}}{dt} + \Omega_g (L_r I_{dr} + M I_{ds}) \end{aligned} \quad (I.23)$$

This system represents the electrical behavior of the three-phase induction motor in the dq reference frame.

### I.4.5.4 Matrix form of the dq model

The previous system can be written in matrix form by separating the derivative terms and the algebraic terms:

$$[u] = [L] \frac{d[I]}{dt} + R[I] \quad (I.24)$$

The complete matrix form is written as follows:

$$\begin{bmatrix} u_{ds} \\ u_{qs} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} L_s & 0 & M & 0 \\ 0 & L_s & 0 & M \\ M & 0 & L_r & 0 \\ 0 & M & 0 & L_r \end{bmatrix} \times \begin{bmatrix} I_{ds} \\ I_{qs} \\ I_{dr} \\ I_{qr} \end{bmatrix} \frac{d}{dt} + \begin{bmatrix} R_s & -\Omega_s L_s & 0 & -\Omega_s M \\ \Omega_s L_s & R_s & \Omega_s M & 0 \\ 0 & -\Omega_g M & R_r & -\Omega_g L_r \\ \Omega_g M & 0 & \Omega_g L_r & R_r \end{bmatrix} \times \begin{bmatrix} I_{ds} \\ I_{qs} \\ I_{dr} \\ I_{qr} \end{bmatrix} \quad (I.25)$$

By using the operator  $S = d/dt$ , the developed matrix form becomes:

$$\begin{bmatrix} u_{ds} \\ u_{qs} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s + L_s S & -\Omega_s L_s & M \cdot S & -\Omega_s M \\ \Omega_s L_s & R_s + L_s S & \Omega_s M & M \cdot S \\ M \cdot S & -\Omega_g M & R_r + L_r \cdot S & -\Omega_g L_r \\ \Omega_g M & M \cdot S & \Omega_g L_r & R_r + L_r S \end{bmatrix} \times \begin{bmatrix} I_{ds} \\ I_{qs} \\ I_{dr} \\ I_{qr} \end{bmatrix} \quad (I.26)$$

This matrix form is useful for the simulation of the machine because it expresses the model using the currents as state variables.

#### I.4.5.5 Relation between the angle and the angular speed

The angle  $\theta$  defining the position of the dq reference frame with respect to the stator is related to the angular speed  $\Omega_s$  by the following equation:

$$\theta = \int \Omega_s dt + \theta_0 \quad (I.27)$$

where  $\theta_0$  is the initial angle.

#### I.4.5.6 Electromagnetic torque and mechanical equation

The electromagnetic torque of the machine can be expressed as:

$$C_{em} = P(\Psi_{ds} I_{qs} - \Psi_{qs} I_{ds}) \quad (I.28)$$

For rotor-flux-oriented control, an equivalent expression can also be written as:

$$C_{em} = P \frac{L_m}{L_r} (\Psi_{dr} I_{qs} - \Psi_{qr} I_{ds}) \quad (I.29)$$

For the Simulink implementation in this chapter, expression (I.28) is used, as it directly involves the stator flux linkages computed from the dq model.

The mechanical equation of the motor is given by:

$$J \frac{d\Omega_r}{dt} = C_{em} - C_r - f\Omega_r \quad (I.30)$$

where  $J$  is the moment of inertia,  $C_{em}$  is the electromagnetic torque,  $C_r$  is the load torque,  $f$  is the viscous friction coefficient, and  $\Omega_r$  is the mechanical rotor speed.

#### I.4.5.7 State-space representation

For simulation purposes, the dq model can be expressed in state-space form. By choosing the current components as state variables, the nonlinear model can be written as:

$$\frac{dX}{dt} = A(\Omega_r)X + BU \quad (I.31)$$

where  $X$  is the state vector,  $U$  is the input voltage vector,  $A(\Omega_r)$  is the state matrix depending on the rotor speed, and  $B$  is the input matrix.

This model is suitable for MATLAB/Simulink implementation because it allows the dynamic behavior of the machine to be studied during starting, steady-state operation, and load variation.

### I.4.5.8 MATLAB/Simulink implementation and simulation results

The previously described mathematical model was implemented in MATLAB/Simulink to study the dynamic behavior of the three phase induction motor.

In this simulation, a load torque step was applied in order to study the behavior of the motor under mechanical load variation. The load torque  $C_r$  was initially equal to 0 N.m. At  $t = 5$  s, it was increased to 8 N.m. This variation allows the effect of the applied load on the electromagnetic torque  $C_{em}$  and on the rotor speed response to be observed.

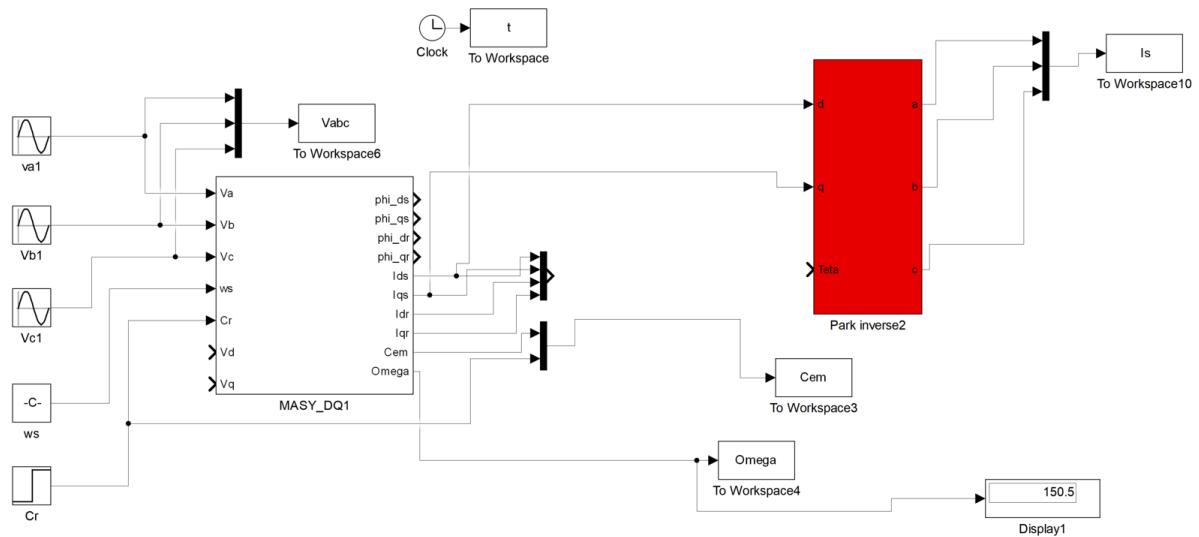


**Fig. I.12:** MATLAB/Simulink model of the three-phase induction motor.

Figure I.12. Global model in MATLAB/Simulink for the simulation of the three-phase induction motor . The model is composed of the dq machine model, transformation blocks, inverter stage, estimator and output blocks for observing the main variables.

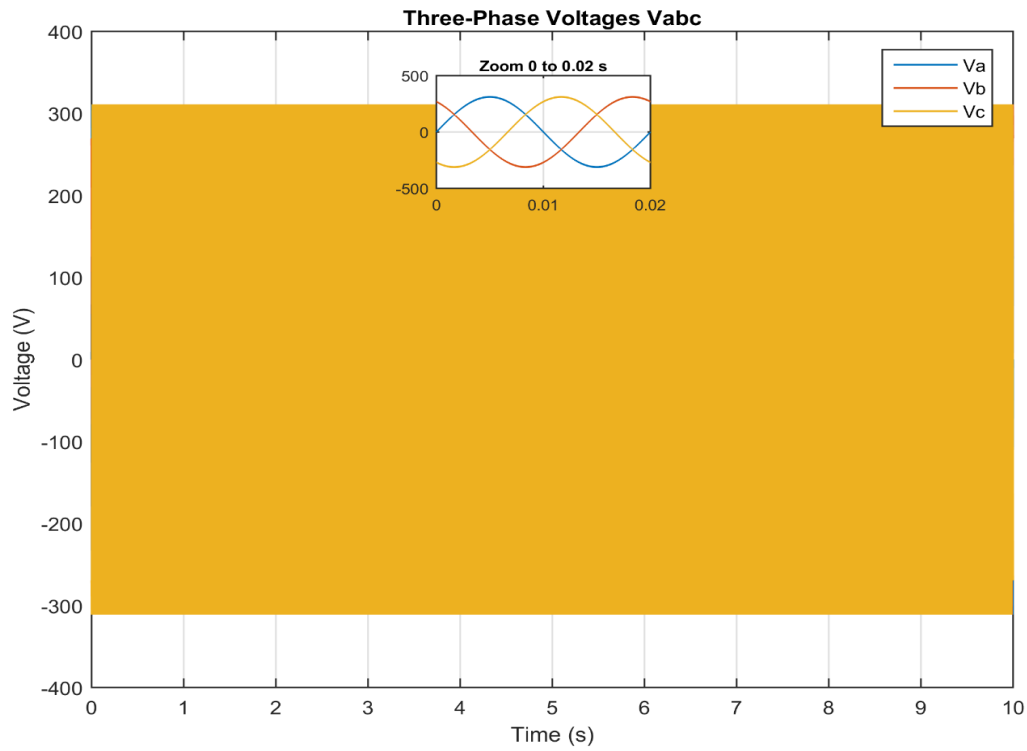


**Fig. I.13:** Three-phase stator voltages Vabc.

Fig. I.13 depicts the three phase stator voltages applied to the motor. The sinusoidal voltages  $V_a$ ,  $V_b$  and  $V_c$  are displaced from each other by 120 electrical degrees. The zoomed view reveals that the supply voltage is balanced.

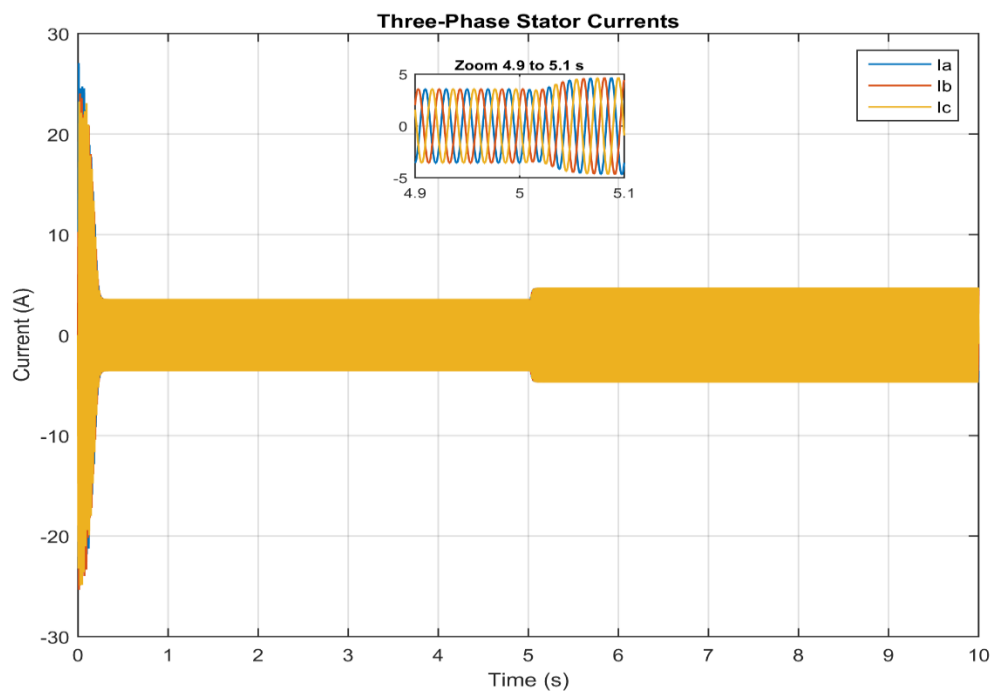


**Fig. I.14:** Three-phase stator currents  $I_a$ ,  $I_b$ , and  $I_c$  with zoom around 5 s.

Fig. I.14 shows the three-phase stator currents. At the beginning of the simulation, a transient current appears due to the starting condition of the motor. After this transient period, the currents become more stable. Around 5 s, a variation appears due to the change in the operating condition.

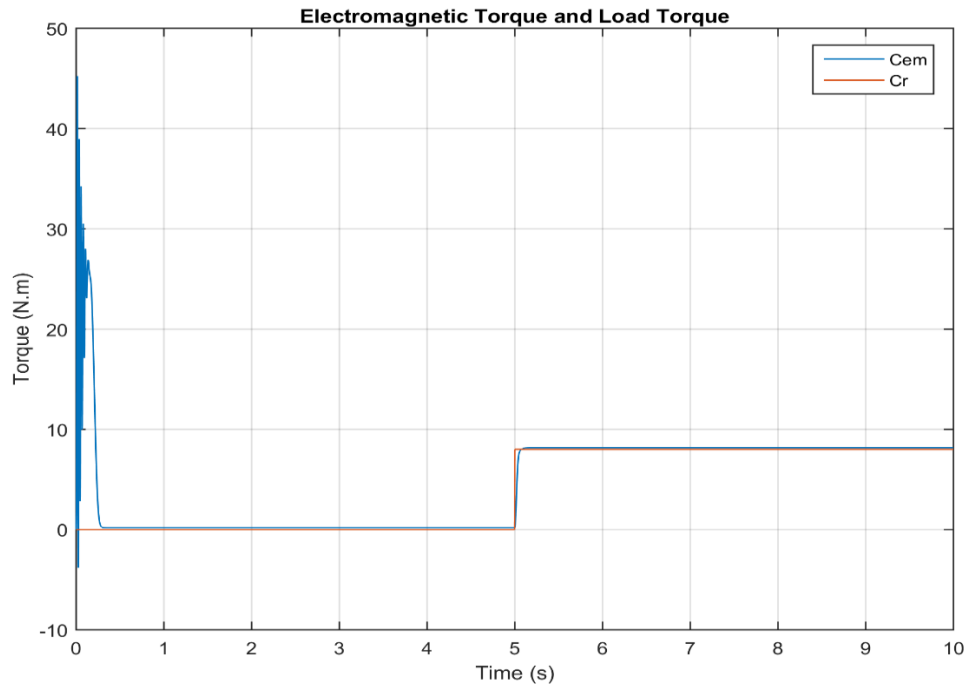


**Fig. I.15:** Electromagnetic torque  $C_{em}$  and load torque  $C_r$ .

Fig. I.15 illustrates the electromagnetic torque and the load torque. At startup, the electromagnetic torque presents a high transient value in order to accelerate the motor. After the starting period, the torque decreases. At  $t = 5$  s, the load torque is applied, and the electromagnetic torque increases to balance the mechanical load.

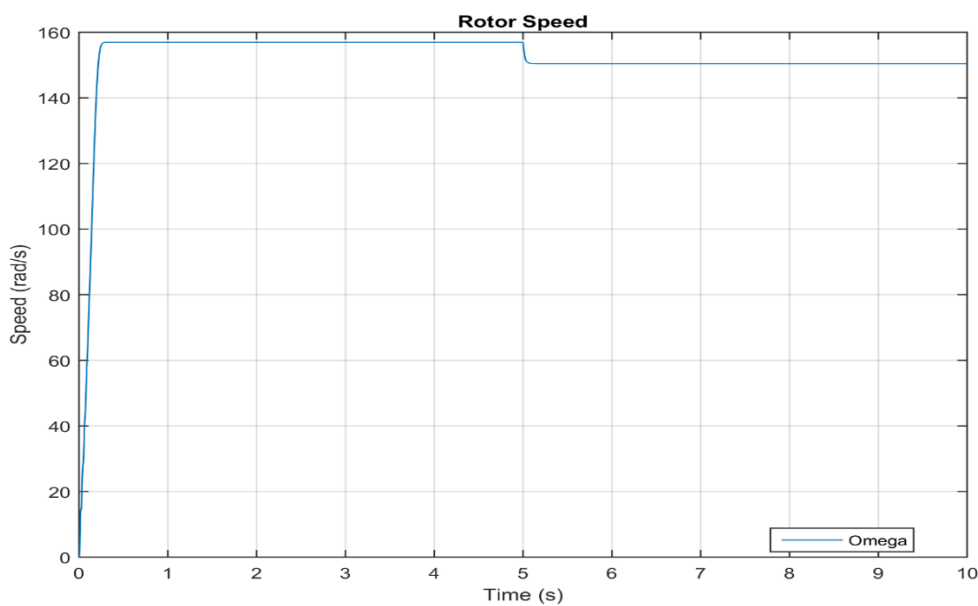


**Fig. I.16:** Rotor speed response of the three-phase induction motor.

Fig. I.16 shows the rotor speed response. The speed increases rapidly during the starting period and then reaches steady-state operation. When the load torque is applied at  $t = 5$  s, a small speed drop appears before the motor stabilizes again.

These simulation results confirm that the dq model correctly represents the main dynamic phenomena of the three-phase induction motor, including starting transient, steady-state operation, load torque application, torque variation, and speed response. Therefore, this model provides a suitable theoretical and simulation basis for the control studies developed in the following chapters.

## I.5 Speed Control Methods of Electric Motors

Modern industrial and domestic drive systems have a critical requirement of accurate control of the motor speed. The method of controlling the speed of the motor depends on the type of motor and the performance required.

There are various speed control techniques and they can be classified based on the motor type and the control approach. One common classification of these methods is inverter control and phase control. These methods are extensively used to various motor systems for effective and flexible speed regulation..

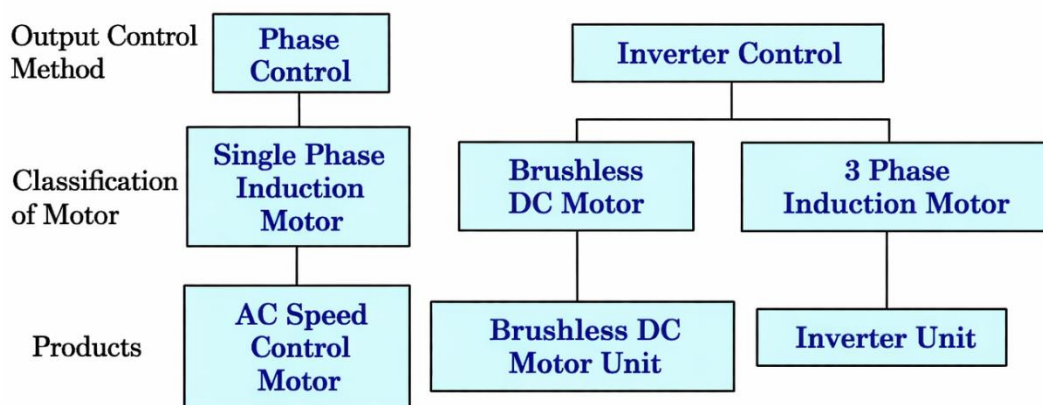


Fig. I.17: Classification of speed control methods of electric motors [11]

### I.5.1 DC Motor Control

The linear characteristics of the DC motor make its speed control quite simple. The speed of a DC motor can be controlled by changing any of the following:

- **Armature voltage control:** The speed is proportional to the applied armature voltage.
- **Field flux control:** The speed of the motor can be controlled by varying the field current.
- **Armature resistance control:** Speed can be decreased by adding resistance in armature circuit.

These methods are very popular due to their simplicity and efficiency in applications where accurate speed control is required.

The relationship between armature voltage and speed of a **DC shunt motor** is illustrated in Fig. I.18.

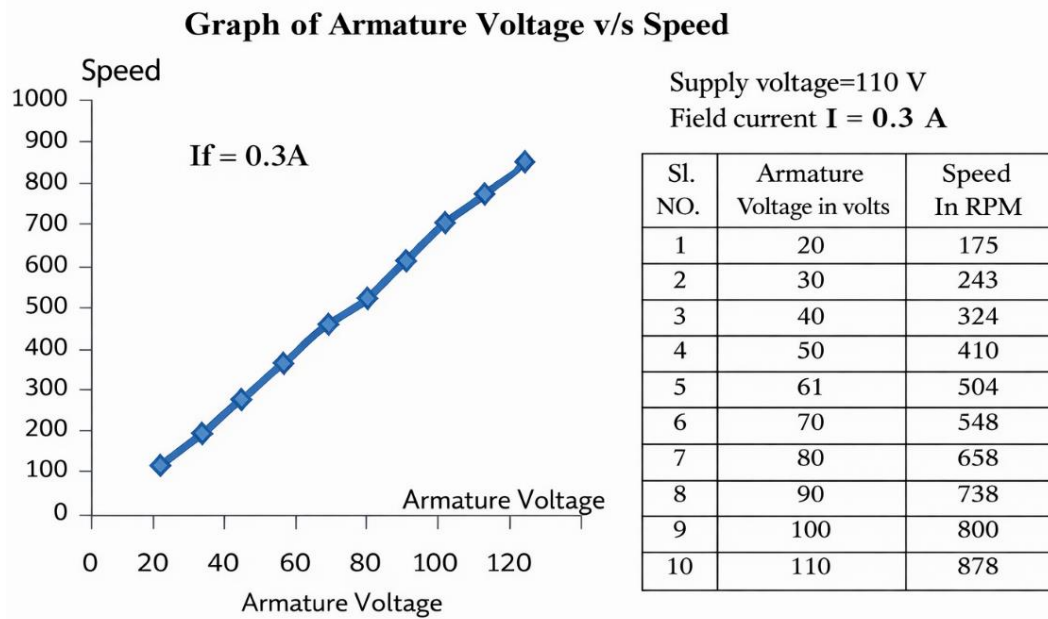


**Fig. I.18:** Experimental relationship between armature voltage and speed of a DC shunt motor [12]

### I.5.2 Induction Motor Control

Speed control of induction motor is more difficult than that of DC motor owing to its nonlinear characteristics and dependency on several parameters. The speed of rotation of an induction motor is mainly determined by the supply frequency, the number of pole pairs and the slip value.

Their low cost and maintenance requirements, robustness and effective speed control techniques are essential to improve their performance and adaptability under different operating conditions.

The speed control methods of induction motors can be classified as stator side control methods and rotor side control methods.

#### a) Stator Side Control Methods

These methods control the motor speed by modifying the stator supply parameters.

- **Stator voltage control:**

The speed control is obtained by changing the amplitude of stator supply voltage. Lower voltage means lower torque and therefore lower speed. It is a simple method but not that efficient and is mainly for low power applications.

- **Frequency control (V/f control):**

This is the most widely used method in the modern applications. The inverter varies the frequency of the supply. The voltage is varied in proportion to maintain the V/f ratio. The air-gap flux remains at its rated value, thus ensuring stable operation over the entire speed range.

- **Pole changing method:**

Speed control is achieved by changing the connection of the stator winding to change the number of effective poles. This method provides discrete speed levels rather than continuous control.

### b) Rotor Side Control Methods

These methods are mainly applied to wound rotor (slip ring) induction motors.

- **Rotor resistance control:**  
Slip rings are used to add extra resistance to the rotor circuit. Increasing resistance increases slip and decreases speed. The starting torque is improved with this method, but the power losses are greater.
- **Cascade control method:**  
In this method two induction motors are connected for speed control. It is used very rarely because of its complexity and cost.

## I.6 Classical Control Methods

Classical control methods are widely used in electric motor drive systems due to their simplicity, reliability and ease of implementation. These methods are based on linear control theory and are commonly used in closed-loop systems to control the motor speed by minimizing the error between the reference and the actual value.

Proportional integral (PI) and Proportional Integral Derivative (PID) are the most popular classical controllers used in industrial applications for DC and induction motor control.

### I.6.1 PI Controller

The Proportional-Integral (PI) controller is one of the most common control strategies used in motor speed control systems. It employs both proportional and integral actions to give fast response and zero steady state error.

The proportional part acts on the instantaneous error and makes the transient response faster. The integral part acts on the steady state offset by integrating the error over time.

The output of the PI controller is given by:

$$u(t) = K_p e(t) + K_i \int e(t) dt \quad (I.32)$$

Where:

- $K_p$ : proportional gain
- $K_i$ : integral gain
- $e(t)$ : error signal

In many motor control applications, the PI controller is preferred because it provides a good trade-off between performance and simplicity without the complexity of derivative action.

### I.6.2 PID Controller

The Proportional-Integral-Derivative (PID) controller is an extension of the PI controller, which adds a derivative term to improve the dynamic response of the system.

The derivative term is a damping term which acts on the rate of change of error to reduce the overshoot and improve the stability of the closed loop.

The PID controller is described by:

$$u(t) = K_p e(t) + K_i \int e(t) dt + K_d \frac{de(t)}{dt} \quad (I.33)$$

Where:

- $K_d$ : derivative gain

The PID controller provides better dynamic performance but is more complex to tune.

## I.7 Intelligent Control

Intelligent control methods were developed to overcome the limitations of classical control techniques especially in nonlinear, complex systems or systems subject to uncertainties and parameter variations. Classical controllers need an accurate analytical model of the plant, intelligent methods do not. Instead, they evolve adaptive strategies using artificial intelligence tools, which can maintain performance across a wide range of operating conditions.

These methods are widely used for the control of electric motors where the behavior of the system can change because of load changes, disturbances and nonlinear characteristics. Intelligent controllers are more flexible and robust than conventional PI and PID controllers.

### I.7.1 Fuzzy Logic Control

Fuzzy Logic Control (FLC) is one of the most widely used intelligent control techniques. It's not based on exact mathematical equations but on human reasoning and the rules of language.

In a fuzzy control system, input variables like error and change of error are converted into fuzzy values using membership functions. A set of if-then rules are then used to determine the control action. In the last step the fuzzy output is translated into a crisp control signal

Fuzzy logic controllers are particularly well suited for motor control problems as they handle nonlinearities and uncertainties well without requiring an exact model of the system.

As shown in Fig. I.19 [16] The FLC structure includes four functional blocks: the fuzzification interface, the rule base, the inference engine and the defuzzification module.

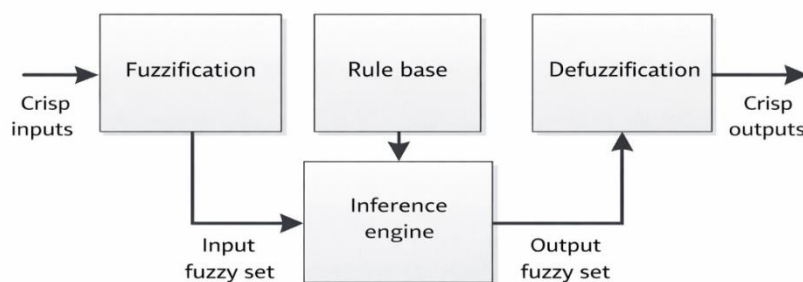


**Fig. I.19:** Structure of a Fuzzy Logic Controller [16]

The fuzzification block converts the numerical inputs to fuzzy sets, and then the inference engine evaluates the applicable rules in order to produce a fuzzy output. This fuzzy result is then fed to the defuzzification block to get the final crisp control signal to be sent to the plant.

### **I.7.2 Advantages of Intelligent Control**

Intelligent control techniques offer several advantages over conventional control techniques such as:

- Ability to handle non-linear and complex systems
- reduced reliance on precise mathematical models
- High robustness against disturbances and parameter variations
- Better adaptability to changing operating conditions

However, these methods are usually more computationally intensive and more complex in design and implementation.

### **I.8 Conclusion**

classical and intelligent control techniques. Classical controllers such as PI and PID are popular because of their simplicity and effectiveness in many industrial applications . However, their performance may be limited to nonlinear systems and uncertainties.

To overcome these drawbacks, intelligent control techniques such as fuzzy logic control have been proposed. These methods are more adaptive and robust and are suitable for modern control applications.

In the next chapter we will be interested in the practical implementation and analysis of the chosen control strategy.

**CHAPTER II :**  
**ARCHITECTURE OF AN ELECTRIC**  
**DRIVE SYSTEM**

## II. INTRODUCTION

An electric drive system, also commonly referred to as a motor control system, is an integrated assembly designed to convert electrical energy into mechanical energy with a high degree of precision. A typical configuration consists of a power source, one or more power electronic converters, an electric motor with a mechanical load connected to its shaft, a sensing system, and a central controller. The fundamental objective of the system is to regulate mechanical quantities, such as rotational speed and torque, by processing data from sensors and intervening through control signals often in the form of pulse-width-modulation (PWM). This chapter explores the comprehensive architecture of these systems, detailing the functional stages and modeling methodologies required to achieve stable and high-performance operation across various motor types [1].

### II.1 General Overview of Variable Speed Drives

Variable speed drives (VSDs) are essential power electronic systems designed to control the rotational speed and torque of electric motors by adjusting the electrical power parameters supplied to them. These systems are of crucial importance in matching the motor performance to the specific requirements of a mechanical load, thus improving the process control and achieving a significant energy optimization in modern industrial automation. The fundamental principle of a VSD involves converting the fixed frequency and voltage of a power source into variable values, allowing for smooth acceleration and deceleration. This capability improves efficiency and extends the operational life of mechanical components by reducing physical stresses and preventing high inrush currents during startup. Consequently, VSD technology is indispensable across a wide range of applications, from simple conveyor belts and pumps to high-precision robotic manufacturing lines [11].

### II.2 Classification of Variable Speed Drives

Variable speed drives (VSDs) are classified according to two main criteria: the type of electrical machine being controlled and the power conversion technique employed. They are broadly divided into two categories: direct current (DC) drives and alternating current (AC) drives. DC drives were historically dominant in industrial applications due to their simple and well-established control methods, while AC drives have progressively replaced them thanks to advances in power electronics and digital control. The selection of a drive type depends on several factors, including the required speed and torque performance, the available power supply, and the constraints of the application environment.

#### II.2.1 DC Variable Speed Drives

DC variable speed drives are specifically designed to manage the speed and torque of DC motors, which include **series-wound, shunt-wound, and compound-wound** configurations. These drives operate by adjusting either the armature voltage or the field excitation current to achieve the desired rotational speed. Although AC technology has become more dominant, DC drives are still widely utilized in applications requiring **high starting torque** and superior speed-torque characteristics, such as in steel mills, mines, and electric traction systems.

The fundamental speed of a DC motor is determined by the following mathematical relationship:

$$N = \frac{U_a - I_a R_a}{K\Phi} \quad (\text{II. 1})$$

Where:

- $N$ : Rotational speed.
- $U_a$ : Applied armature voltage.
- $I_a R_a$ : Voltage drop across the armature circuit resistance.
- $\Phi$ : Magnetic flux per pole.
- $K$ : A constant related to the motor's construction (number of poles, conductors, etc.).

This equation indicates that the speed is **directly proportional to the back electromotive force (EMF)**, where  $e = U_a - I_a R_a$ , and **inversely proportional to the magnetic flux per pole**. This dependency allows for two primary methods of speed control: armature voltage control (for speeds below the rated value) and flux control (for speeds above the rated value by weakening the field current). Structurally, the architecture of these drives is often represented by an electrical equivalent circuit that separates the stator field and rotor armature components to facilitate precise modeling and control.

### II.2.2 AC Variable Speed Drives

AC variable speed drives, commonly referred to as **Variable Frequency Drives (VFDs)**, are the most widely used systems in modern industrial automation, primarily because of the robustness and cost-effectiveness of **three-phase induction motors**. These drives control the motor's speed and torque by modulating both the **frequency and the voltage** of the alternating current supplied to the stator windings. This allows for efficient operation across a vast range of industrial loads, from fans and pumps to complex conveyor systems.

The rotational speed of an induction motor is intrinsically linked to the supply frequency and the motor's physical construction. The **synchronous speed** ( $N_s$ ) of the rotating magnetic field is defined by the following mathematical relationship:

$$N_s = \frac{60 \cdot f}{P} \quad (\text{II. 2})$$

Where:

- $N_s$  : Synchronous speed in revolutions per minute (RPM).
- $f$ : Supply frequency in Hertz (Hz).
- $P$ : Number of pole pairs.

In practice, the rotor speed ( $N$ ) is always slightly less than the synchronous speed due to **slip** ( $s$ ), which is necessary for torque production through electromagnetic induction. The actual speed is expressed as:

$$N = N_s \cdot (1 - s) \quad (\text{II. 3})$$

Modern AC drives often utilize **scalar control (V/f control)** to maintain a constant magnetic flux. The **origin of this relationship** is found in the fundamental Electromotive Force (EMF) equation of an AC machine:

$$E_1 = 4.44 \cdot f \cdot N_1 \cdot \Phi_m \quad (\text{II. 4})$$

Where  $E_1$  is the induced voltage,  $N_1$  is the number of stator turns per phase, and  $\Phi_m$  is the maximum magnetic flux per pole. By neglecting the relatively small voltage drop across the stator resistance and leakage reactance, the applied stator voltage  $V_1$  is approximately equal to the induced EMF ( $V_1 \approx E_1$ ). Consequently, the relationship becomes:

$$\Phi_m \propto \frac{V_1}{f} \quad (\text{II. 5})$$

To maintain a constant flux density and prevent the motor's magnetic field from reaching **saturation** at lower frequencies, the applied voltage must be varied in direct proportion to the frequency ( $V_1/f = \text{constant}$ ). This ensures that the motor develops a constant rated torque throughout its base operating range.

For high-precision tasks, **Vector Control** or **Field Oriented Control (FOC)** is employed to overcome the nonlinear dependencies of induction motors. The **mathematical origin of this control** lies in the **Clarke-Park transformations**, which convert the three-phase stationary current vectors (a,b,c) into a two-dimensional rotating reference frame (d-q). In this rotating frame, the tasks of flux and torque generation are decoupled into two independent DC quantities:

- **Direct-axis current ( $I_d$ )**: Used to produce and control the magnetization flux, acting similarly to the field current of a DC motor.
- **Quadrature-axis current ( $I_q$ )**: This current is responsible for generating and controlling the electromagnetic torque independently.

By aligning the d-axis with the rotor flux, the induction motor is controlled like a **separately excited DC machine**. This enabling architecture allows for high dynamic performance, full torque at zero speed, and precision that was traditionally only possible with DC drive systems.

### II.2.3 Conventional and Intelligent Variable Speed Drives

Variable Speed Drives can be broadly classified into two categories according to the complexity and adaptability of their control logic: **conventional systems** and **intelligent systems**.

**Conventional drives** usually rely on standard linear control strategies, mainly **Proportional-Integral (PI)** or **PID controllers**. These systems are very effective in stable processes where the system parameters remain almost constant. However, their performance strongly depends on the availability of an accurate mathematical model of the motor. Since they are designed to operate in a linear way, they may lose precision when dealing with **non-linearities** or sudden parameter changes, which are common in industrial environments.

In contrast, **intelligent drives** use advanced computational algorithms such as **Fuzzy Logic** or **Artificial Neural Networks** to control the motor's behavior. The main advantage of these systems is their high **robustness**. Unlike conventional methods, they do not need an exact mathematical representation of the machine to work effectively. Instead, they can keep good **stability** and **performance** even when the system is affected by unpredictable **load disturbances** or variations in motor parameters during operation.

The advantage of **intelligent control** becomes especially clear in the **dynamic response** of the system. While a conventional PI controller may produce speed **overshoot** and oscillations before reaching the steady state, an intelligent controller, such as one based on **Fuzzy Logic**, can usually reach the target speed faster and with little or no overshoot, giving a smoother transition.

This behavior is illustrated in Fig. II.1, where the **fuzzy control response** is faster and presents zero overshoot compared with the **linear control response** [19].

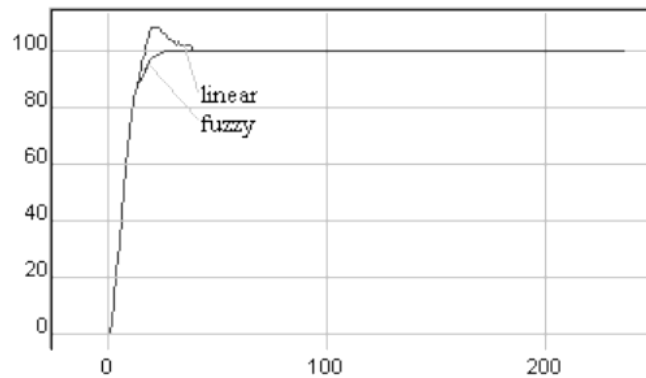


**Fig. II.1:** Speed Response Comparison Between Linear and Fuzzy Control [19].

### II.3 Global Structure of an Intelligent Electric Drive System

An intelligent electric drive system is designed to convert electrical energy into controlled mechanical motion. As shown in Fig. II.2, its general structure includes a power source, power converter, electric motor, mechanical load, and motor controller. These elements work together through electrical and mechanical connections, supported by measurement and feedback signals.

The power source supplies the required electrical energy, while the converter adjusts it according to the motor needs. The electric motor then transforms this controlled electrical energy into mechanical power to drive the load.

The motor controller is the main control unit of the system. It receives feedback signals such as voltage  $u$ , current  $i$ , speed  $\Omega_r$ , and position  $\theta$ , then generates PWM signals to control the converter. This closed-loop operation improves the accuracy, stability, and dynamic performance of the drive system.

Protection functions may also be integrated into the controller or converter to protect the system against faults such as overcurrent, overvoltage, and overheating.

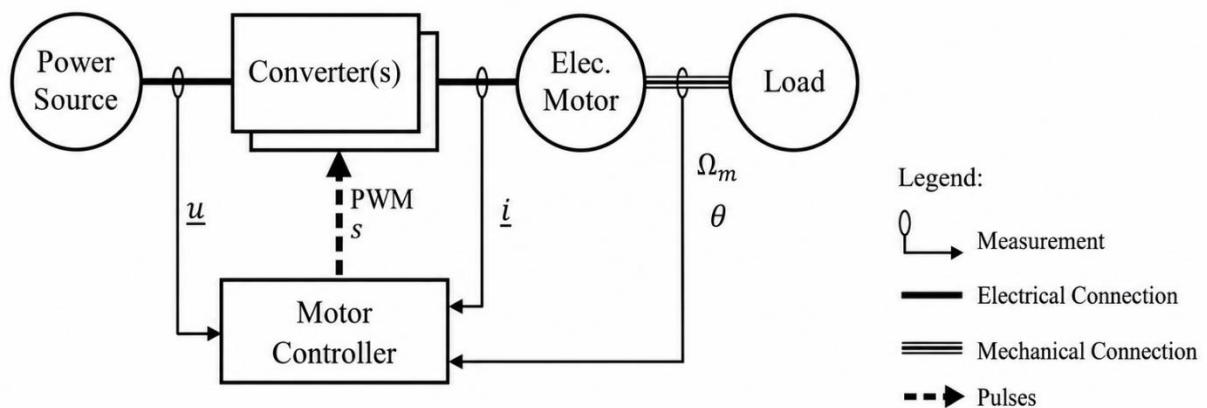


**Fig. II.2:** Global structure of an intelligent electric drive system [1].

### II.3.1 Power Stage

The power stage represents the physical foundation of the electric drive system. Its main function is to process and deliver the required electrical energy to the motor. This energy may come from different sources, such as the three-phase AC grid or a DC source, including batteries or the output of a rectifier.

Depending on the type of motor used, the power stage relies on specific power electronic converters, such as DC/DC choppers or DC/AC inverters, to convert the electrical energy into the suitable form required by the motor. These static converters are designed to handle high power levels while maintaining the efficiency needed in industrial applications.

### II.3.2 Control Stage

The control stage is the part responsible for managing the behavior of the electric motor. Its main role is to make the motor follow the desired operating conditions, such as speed, torque, or position. To do this, the controller receives reference values and compares them with the actual values measured by the sensors.

According to the error between the desired and measured values, the controller generates suitable control signals for the power converter. These signals allow the converter to adjust the electrical energy supplied to the motor, which helps improve the system response, stability, and accuracy.

Different control strategies can be used in this stage. Classical controllers, such as PI or PID controllers, are commonly used because they are simple and effective in many applications. However, in intelligent electric drive systems, more advanced methods such as fuzzy logic control can also be applied. Fuzzy control is useful because it can deal with nonlinear behavior and does not always require an exact mathematical model of the motor.

Therefore, the control stage represents the decision-making part of the drive system. It links the reference command, the feedback information, and the power converter in order to obtain a controlled and efficient motor operation.

### II.3.3 Measurement and Feedback Stage

The measurement and feedback stage is necessary for the drive system to operate accurately in closed-loop control. Its main role is to observe the real state of the system and send this information back to the controller. For this purpose, different sensors are used to measure important electrical and mechanical quantities.

On the electrical side, the system can measure variables such as the input voltage ( $u$ ) and the motor current ( $i$ ). These measurements help the controller evaluate the electrical conditions of the converter and the motor. On the mechanical side, the motor speed ( $\Omega_r$ ) and the shaft position ( $\theta$ ) are measured to know how the motor is actually moving.

The feedback signals are then compared with the reference values required by the system. If there is a difference between the desired and actual behavior, the controller adjusts its control signals to correct the error. In this way, the measurement and feedback stage improves the accuracy, stability, and dynamic response of the electric drive system.

### II.3.4 Protection Stage

The protection stage plays an important role in maintaining the safety and reliability of the electric drive system. Its main function is to continuously monitor the system and detect abnormal operating conditions that may affect the motor or the power converter.

This stage generally includes protection circuits capable of identifying faults such as short circuits, overcurrents, undervoltage, and overheating. When a fault is detected, the protection stage can quickly activate protective actions, such as stopping the power converter, in order to prevent hardware damage or motor burnout.

In modern electric drive systems, many of these protection functions are integrated directly into intelligent power modules. This helps simplify the overall system design while improving its safety and reliability.

## II.4 Study of Power Converters

Power converters, also called static converters, are important elements in electric drive systems. Their main role is to adapt the electrical energy coming from the source to the requirements of the motor. In other words, they control electrical quantities such as voltage, current, and frequency in order to obtain the desired motor speed, torque, and direction of rotation.

There are several types of power converters depending on the nature of the input and output energy. For example, a rectifier is used to convert AC voltage into DC voltage, while an inverter converts DC voltage into AC voltage with controlled frequency and amplitude. A DC/DC chopper is used to regulate the value of a DC voltage, especially in DC motor drives or battery-supplied systems.

These converters are based on fast semiconductor switches such as MOSFETs, IGBTs, and diodes. By switching rapidly, they allow precise control of the power delivered to the motor while reducing energy losses. Therefore, power converters improve the efficiency, flexibility, and dynamic performance of intelligent electric drive systems.

### II.4.1 Rectifiers

In an intelligent electric drive system, rectifiers serve as the primary interface when the system is energized by an alternating current (AC) grid. Their fundamental role is the conversion of AC voltage into a stable direct current (DC) output, which typically feeds the DC link or directly supplies the armature of a DC motor. These converters are classified based on their topology into single-phase or three-phase units, and further into uncontrolled (diode-based) or controlled (thyristor-based) configurations. While simple diode bridges provide a fixed DC voltage, controlled rectifiers allow for precise regulation of the output power by electronically delaying the firing pulses to the semiconductor switches.

The output voltage of a controlled rectifier is determined by the firing angle ( $\alpha$ ), which represents the time delay between the natural commutation point and the moment the thyristor is triggered into conduction. In industrial applications involving high-power drives, the three-phase bridge configuration is highly preferred because it provides a more continuous current flow, exhibits lower harmonic distortion, and ensures a symmetrical load on the AC supply lines.

For a three-phase fully controlled converter, the average output DC voltage ( $V_0$ ) can be calculated using the following relationship:

$$V_0 = \frac{3 \cdot V_{Lm}}{\pi} \cos \alpha \quad (\text{II. 6})$$

Where:

- $V_{Lm}$  is the peak value of the line-to-line AC voltage.
- $\alpha$  is the thyristor firing angle.

By adjusting the firing angle  $\alpha$  between  $0^\circ$  and  $90^\circ$ , the converter operates in rectifier mode, where power is transferred from the AC source to the DC motor. When the firing angle becomes greater than  $90^\circ$ , the average output voltage becomes negative, and the converter may operate in inverter mode. In this case, the motor can work as a generator during regenerative braking, allowing energy to be returned from the motor to the AC supply, provided that the drive conditions allow reverse power flow.

As shown in Fig. II.3, the three-phase fully controlled rectifier consists of six thyristors T11 to T16 connected in a bridge configuration. This arrangement allows full control of the average DC output voltage delivered to the motor by varying the firing angle  $\alpha$ .

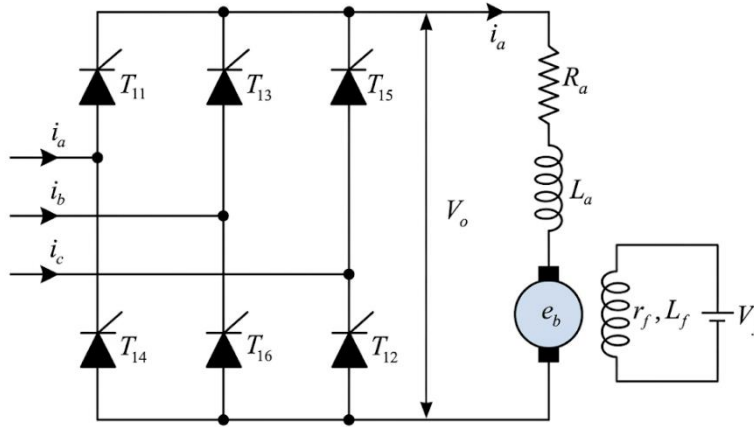


**Fig. II.3:** Three-phase fully controlled rectifier supplying a DC motor drive [17].

## II.4.2 DC-DC Converters

DC-DC converters, widely known as choppers, are static power electronic circuits used to convert a constant DC input voltage into a controlled variable DC output voltage. In an intelligent electric drive system, these converters play an important role in regulating the electrical power supplied to the motor. In this sense, they can be considered the DC equivalent of a transformer in AC systems, since they adapt the voltage level according to the needs of the load [17].

The operation of DC-DC converters is mainly based on high-speed semiconductor switches, such as MOSFETs, IGBTs, or GTOs. These switches are controlled using Pulse Width Modulation (PWM), which adjusts the output voltage by changing the ON time of the switch. As shown in Fig. II.4, when the switch is ON, the input DC voltage is applied to the load, and when it is OFF, the load is disconnected from the source, producing a pulsed output voltage.

By varying the duty cycle, the converter “chops” the constant input voltage into pulses and controls their average value. This average value determines the DC output voltage applied to the motor,

making DC-DC converters useful for controlling the armature voltage, motor speed, and dynamic response of DC motor drives.

Modern DC-DC converters may use hard-switching or soft-switching techniques. Soft-switching converters help reduce switching losses, electrical stress, and electromagnetic interference. Depending on the application, they can also be designed for single-quadrant or multi-quadrant operation, allowing direction control and regenerative braking.

The basic operation of a DC-DC converter is governed by the duty cycle, usually denoted by  $\alpha$ , which represents the ratio between the switch ON time  $t_{on}$  and the total switching period  $T$ :

$$\alpha = \frac{t_{on}}{T} \quad (II. 7)$$

where  $t_{on}$  is the time during which the switch remains ON, and  $T$  is the total switching period.

For a basic step-down converter, also called a Buck chopper, the average output voltage is given by:

$$V_0 = \alpha V_{in} \quad (II. 8)$$

where  $V_{in}$  is the input DC voltage,  $V_0$  is the average output DC voltage, and  $\alpha$  is the duty cycle. Therefore, by increasing the duty cycle, the average output voltage increases. In contrast, reducing the duty cycle decreases the voltage supplied to the motor.

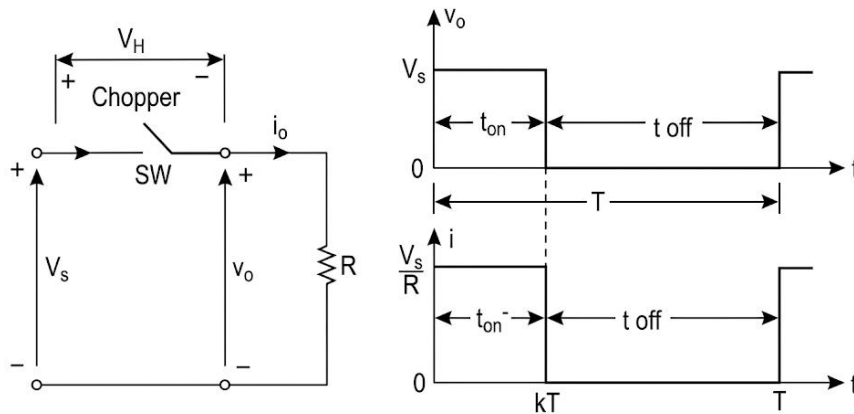


**Fig. II.4:** Basic DC-DC chopper circuit and switching operation [12].

#### II.4.2.1 Mathematical Modeling of the Series Chopper (Buck Converter)

To analyze and control the series chopper (buck converter), we first establish its mathematical model based on the state-space representation. The state vector is defined by the energy-storing elements in the circuit:

$$[X] = \begin{bmatrix} I_L \\ V_c \end{bmatrix} \quad (II. 9)$$

where  $I_L$  is the current flowing through the inductor and  $V_c$  is the voltage across the capacitor, which also represents the output voltage .

Assuming the converter operates in continuous conduction mode and neglecting the series resistance of the inductor, the average state equation of the system is given by:

$$\frac{d}{dt}x(t) = A \cdot x(t) + B \cdot u(t) \quad (II. 10)$$

$$y(t) = C \cdot x(t) \quad (\text{II. 11})$$

The state matrix A, the input matrix B, and the output matrix C are defined as follows:

$$A = \begin{bmatrix} 0 & -\frac{1}{L} \\ \frac{1}{C} & -\frac{1}{RC} \end{bmatrix}, \quad B = \begin{bmatrix} \frac{E}{L} \\ 0 \end{bmatrix}, \quad C = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad (\text{II. 12})$$

In this model,  $\alpha$  represents the duty cycle, while E is the input DC voltage. Therefore, the input voltage applied to the motor is given by  $u = E \cdot \alpha$ . From these matrices, the differential equations of the system can be developed as follows:

$$\frac{dI_L}{dt} = \frac{E\alpha - V_c}{L} \quad (\text{II. 13})$$

$$\frac{dV_c}{dt} = \frac{I_L}{C} - \frac{V_c}{R \cdot C} \quad (\text{II. 14})$$

**Extended Third-Order Model (Loaded Chopper)** When an additional inductive load is connected (or when using a second-order output filter), the system extends to a third-order state-space model. The state vector is expanded to include the load current ( $I_{ch}$ ), leading to the following state matrix equation:

$$\frac{d}{dt} \begin{bmatrix} I_L \\ V_c \\ I_{ch} \end{bmatrix} = \begin{bmatrix} 0 & -\frac{1}{L} & 0 \\ \frac{1}{C} & 0 & -\frac{1}{C} \\ 0 & \frac{1}{L_{ch}} & -\frac{R_{ch}}{L_{ch}} \end{bmatrix} \begin{bmatrix} I_L \\ V_c \\ I_{ch} \end{bmatrix} + \begin{bmatrix} \frac{\alpha E}{L} \\ 0 \\ 0 \end{bmatrix} \quad (\text{II. 15})$$

#### II.4.2.2 Simulink Implementation (Unloaded and Loaded)

To validate the theoretical mathematical modeling, the system is simulated using MATLAB /Simulink based on the state-space equations.

##### ➤ Unloaded Chopper (Open-Loop)

The simulation is carried out over a total duration of 10 seconds. The duty cycle is initially set to  $\alpha=1$  for the first 5 seconds, corresponding to an effective input voltage of  $u=\alpha \cdot E=240$  V, then stepped down to  $\alpha=0.5$ , which reduces the input voltage to  $u=120$  V, in order to observe the system's dynamic response to a sudden change in the operating point.

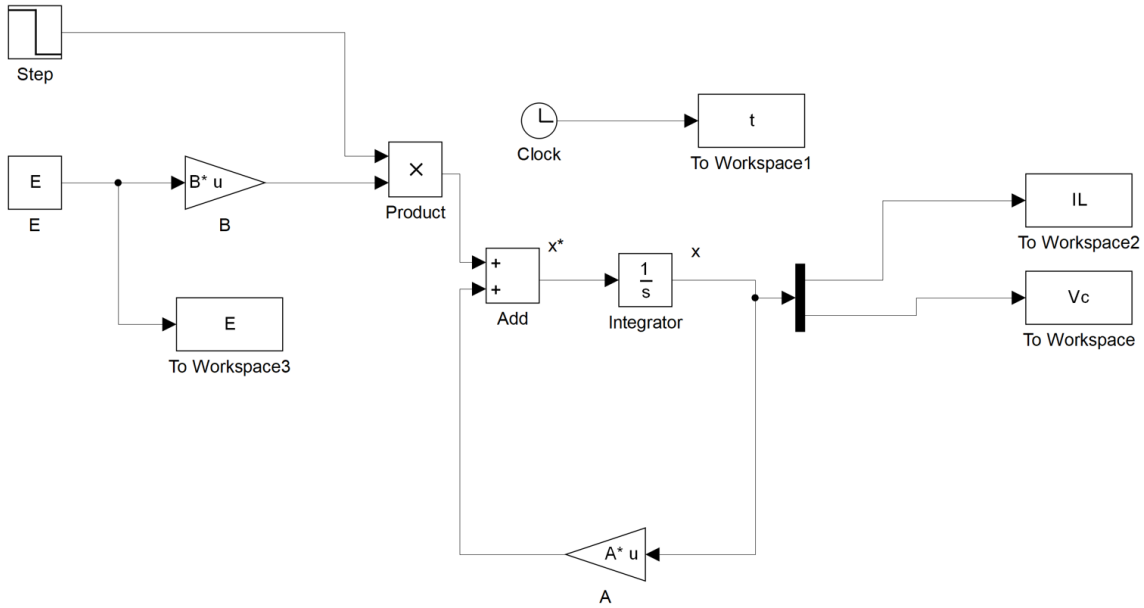


**Fig. II.5:** Simulink block diagram of the open-loop chopper (Unloaded).

This diagram implements the state-space model using an integrator block. The input voltage ( $E$ ) is scaled by matrix  $B$  and multiplied by the duty cycle ( $\alpha$ ), while the feedback term  $A \cdot X$  is added to form the state derivative. The resulting state variables  $I_L$  and  $V_c$  are exported to the MATLAB workspace for visualization.

The corresponding simulation results representing the waveforms of the output voltage  $V_c(t)$  and the inductor current  $I_L(t)$  under unloaded conditions are illustrated below:

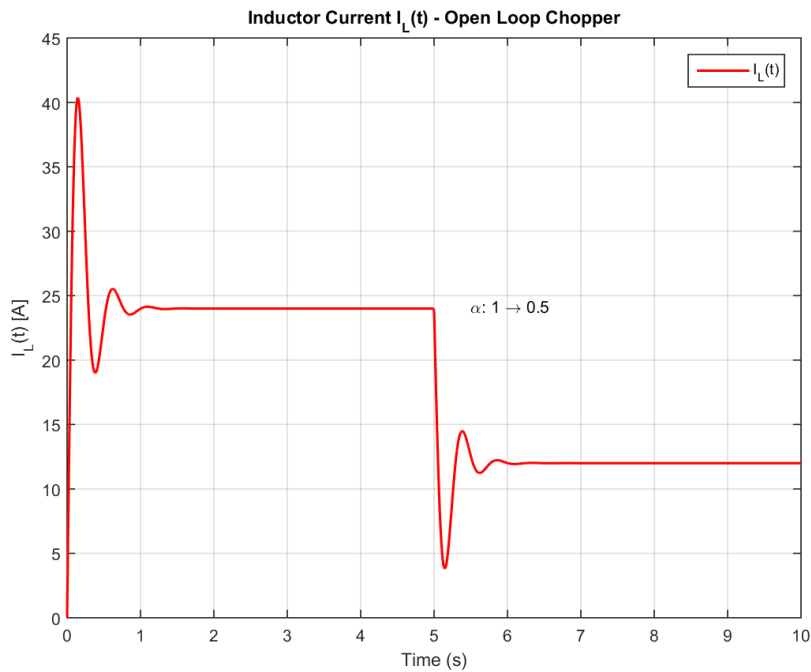


**Fig. II.6:** Waveform of the inductor current  $I_L(t)$  for the unloaded chopper.

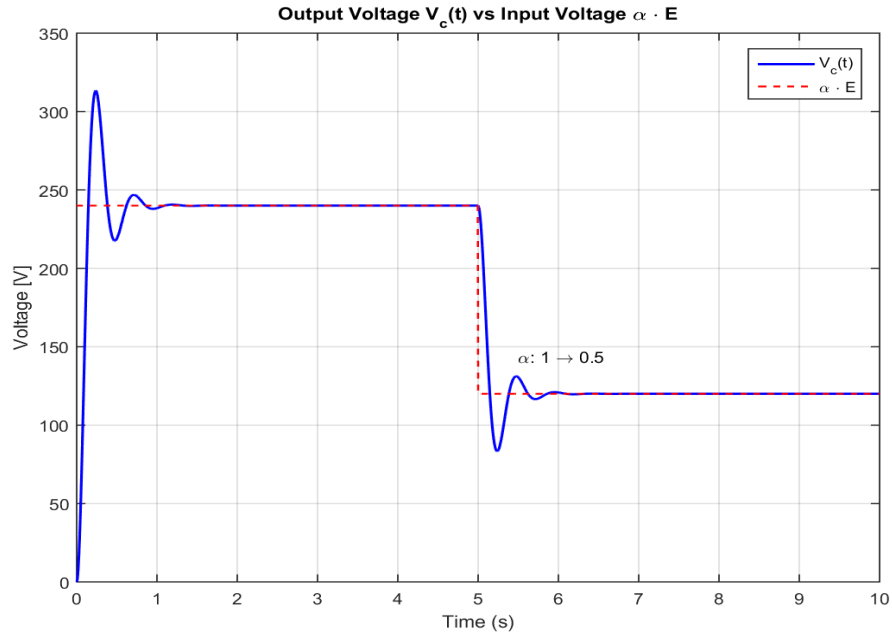


**Fig. II.7:** Waveform of the output voltage  $V_c(t)$  for the unloaded chopper.

### ➤ Loaded Chopper — $R_{ch} = 50 \Omega$

To investigate the effect of the load resistance on the system behavior, the simulation is carried out with  $R_{ch} = 50 \Omega$ . Both open-loop (no regulation) and closed-loop (PI regulation) configurations are tested under the same operating conditions, and the results are compared below.

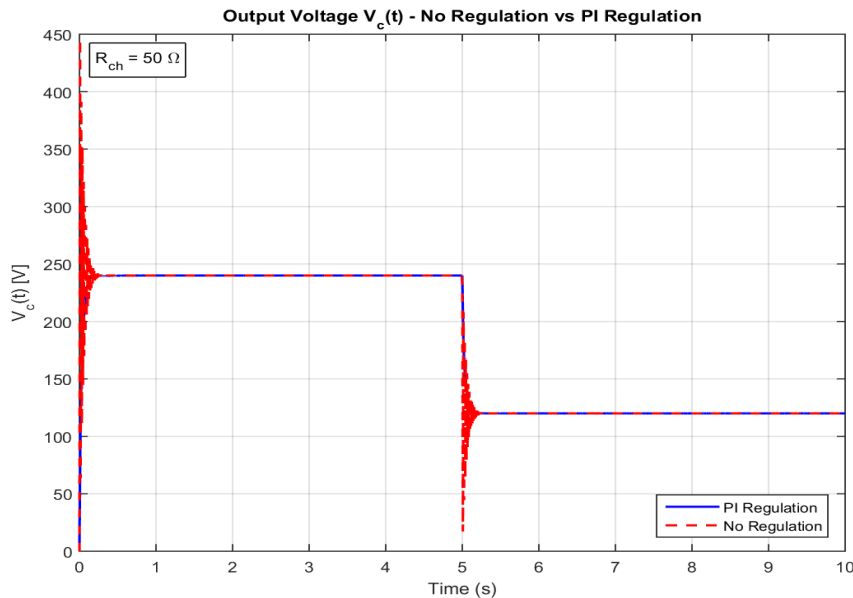


**Fig. II.8:** Output voltage  $V_c(t)$  — No Regulation vs PI Regulation ( $R_{ch} = 50 \Omega$ ).

At light load, both responses are nearly identical. The voltage settles at  $V_C = \alpha \cdot E$ , confirming that the system is naturally well-behaved at high resistance..

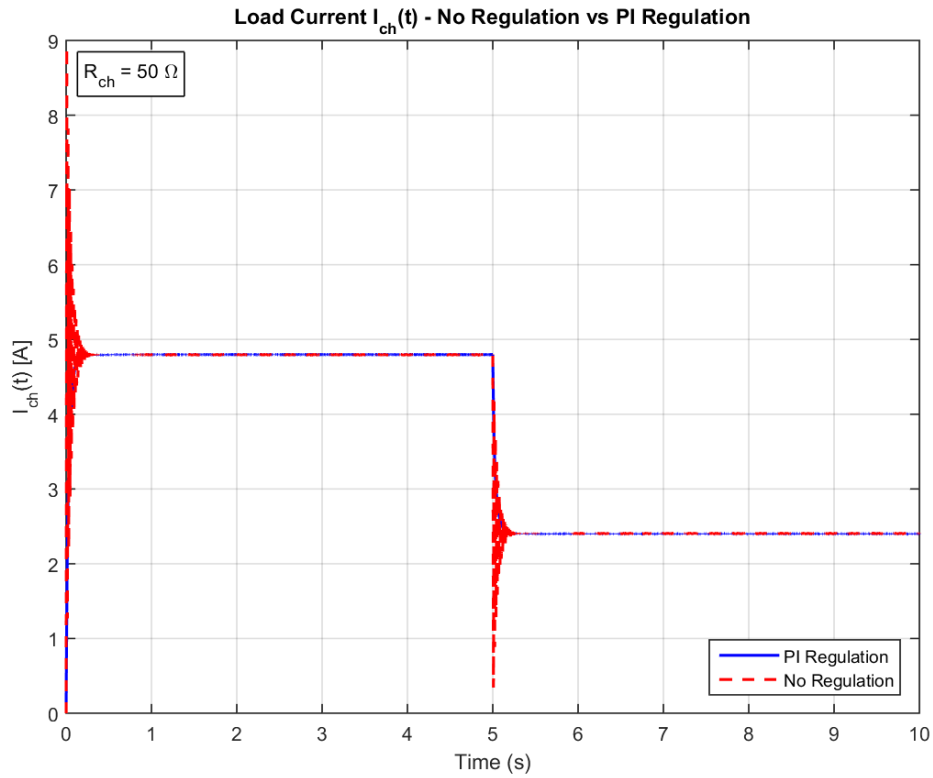


**Fig. II.9:** Load current  $I_{ch}(t)$  — No Regulation vs PI Regulation ( $R_{ch} = 50 \Omega$ ).

The load current follows  $I_{ch} = V_c/R_{ch}$ , settling at 4.8 A then 2.4 A with minimal difference between the two configurations.

### ➤ Loaded Chopper — $R_{ch} = 5 \Omega$

The simulation is repeated with  $R_{ch} = 5 \Omega$ , representing a heavier load condition. The duty cycle is stepped from  $\alpha = 1$  to  $\alpha = 0.5$  at  $t = 5$  s.

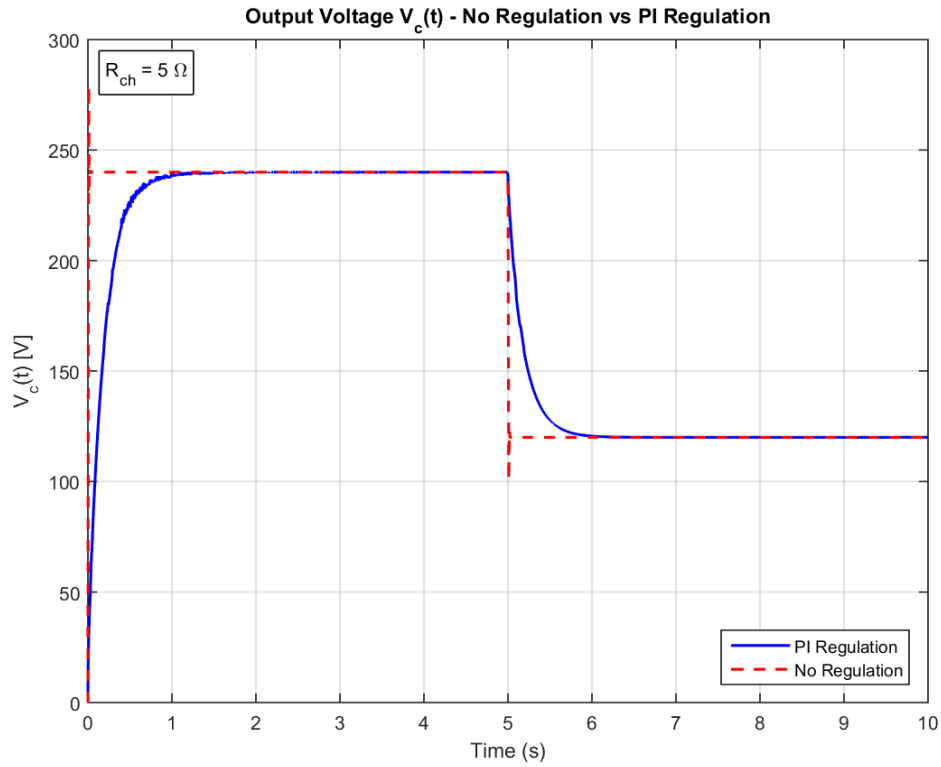


**Fig. II.10:** Output voltage  $V_c(t)$  — No Regulation vs PI Regulation ( $R_{ch} = 5 \Omega$ ).

Both responses settle at the same steady-state values of 240 V and 120 V. However, after the duty cycle step at  $t = 5$  s, the open-loop voltage drops abruptly while the PI-regulated response transitions more smoothly, demonstrating better dynamic performance under heavy load.

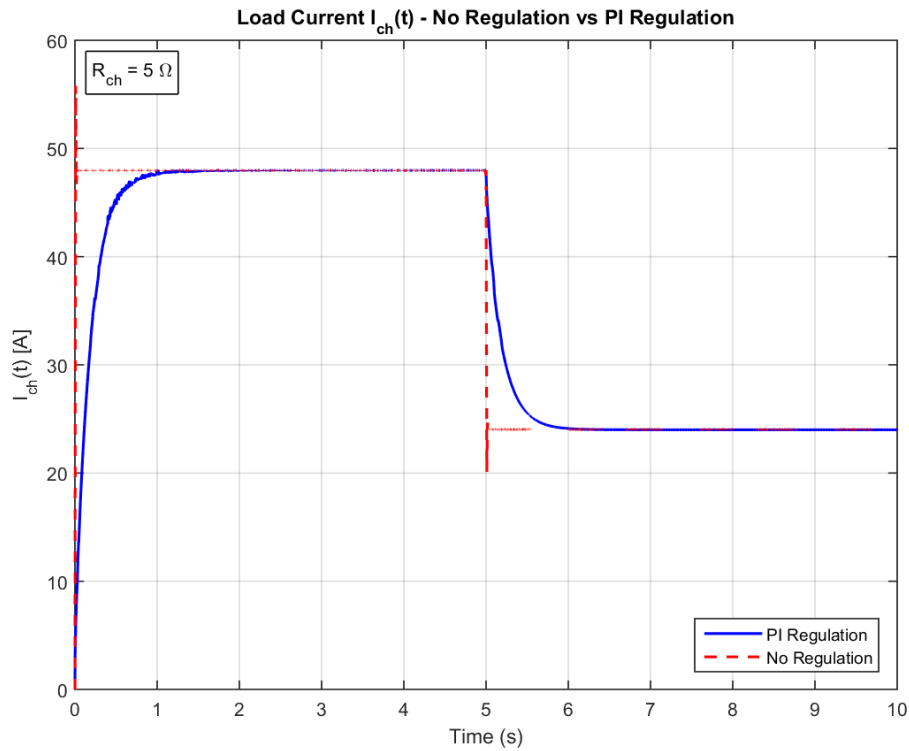


**Fig. II.11:** Load current  $I_{ch}(t)$  — No Regulation vs PI Regulation ( $R_{ch} = 5 \Omega$ ).

The steady-state load current reaches  $I_{ch} = V_c/R_{ch} \approx 48$  A then 24 A. The PI controller reduces the transient at  $t = 5$  s and ensures a smoother settling compared to the open-loop case, confirming its effectiveness under demanding load conditions.

### II.4.3 Inverters

Inverters, or DC-AC converters, are power electronic converters used to transform a DC voltage into an AC voltage with controlled amplitude and frequency. Depending on the application, inverters can be classified into **single-phase inverters** and **three-phase inverters**. Single-phase inverters generate one alternating voltage and are generally used in low-power or domestic applications. Three-phase inverters, on the other hand, generate three AC voltages shifted by  $120^\circ$ , which makes them more suitable for industrial electric drives and AC motor control.

In electric drive systems, especially for asynchronous machines, the three-phase voltage source inverter is the most commonly used structure. As shown in Fig. II.12, it consists of a DC voltage source  $V_{dc}$  and six semiconductor switches S1 to S6, arranged in three inverter legs. Each leg is connected to one motor phase and produces one phase voltage:  $V_{aN}$ ,  $V_{bN}$ , and  $V_{cN}$ . The anti-parallel diodes connected across the switches allow the circulation of the inductive motor current during switching transitions.

The motor phases are represented by their resistance  $R$ , inductance  $L$ , and back electromotive forces  $e_a$ ,  $e_b$ , and  $e_c$ . The inverter switches are controlled in a complementary way, meaning that the two switches of the same leg must not be turned on at the same time in order to avoid a short circuit across the DC link. By applying a suitable switching strategy, the inverter can generate the required three-phase voltage system for motor operation.

To obtain output currents close to sinusoidal waveforms, Pulse Width Modulation (PWM) is usually used. PWM controls the duty cycle of the switches in order to synthesize the desired AC voltages at the motor terminals. By varying the output frequency, the inverter controls the speed of the motor. In scalar control, the  $V/f$  ratio is often kept constant, while in advanced control methods such as Field Oriented Control (FOC), the inverter follows the reference voltage vectors generated by the control algorithm.

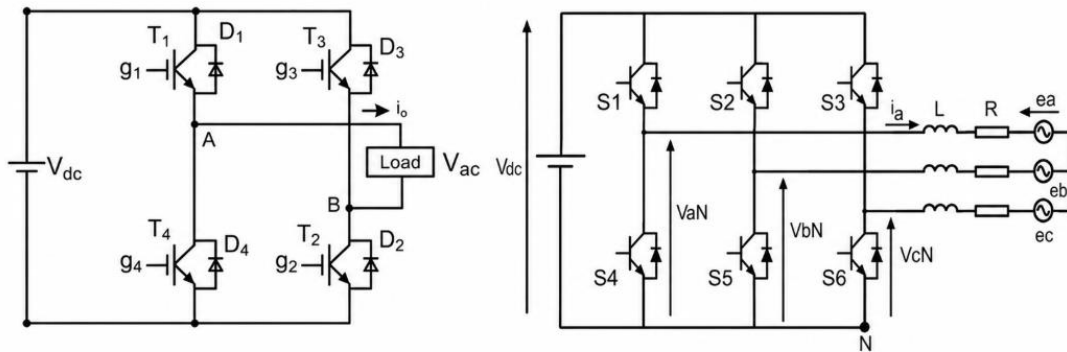


**Fig. II.12:** Single-phase and three-phase voltage source inverter topologies for AC load and machine supply.

### II.4.3.1 Single-Phase Inverter

The single-phase H-bridge inverter feeds a single-phase load, generally through an LC output filter. Taking the inductor current and the capacitor voltage as state variables ( $x_1$  and  $x_2$ ), the system is described by the two general state equations:

$$L \frac{di_L}{dt} = V_{inv} - V_C - r_L i_L \quad (\text{II.16})$$

$$C \frac{dv_C}{dt} = i_L - i_{load} \quad (\text{II.17})$$

where  $L$  is the filter inductance,  $C$  the filter capacitance,  $i_L$  the filter inductor current,  $V_C$  the capacitor (output) voltage,  $r_L$  the internal resistance of the filter inductor,  $V_{inv}$  the inverter output voltage, and  $i_{load}$  the load current.

They express, respectively, Kirchhoff's voltage law along the inductor loop and the current balance at the output node.

Depending on the load connected at the inverter output, the state-space model takes one of the following forms:

➤ **No-load operation:**

$$\begin{bmatrix} \frac{dx_1}{dt} \\ \frac{dx_2}{dt} \end{bmatrix} = \begin{bmatrix} \frac{-r_L}{L} & \frac{-1}{L} \\ \frac{1}{C} & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} \frac{1}{L} \\ 0 \end{bmatrix} V_{inv} \quad (\text{II.18})$$

where  $x_1$  and  $x_2$  are the state variables of the system, namely the filter inductor current and the capacitor voltage, respectively.

➤ **Resistive load R:**

$$\begin{bmatrix} \frac{dx_1}{dt} \\ \frac{dx_2}{dt} \end{bmatrix} = \begin{bmatrix} \frac{-r_L}{L} & \frac{-1}{L} \\ \frac{1}{C} & \frac{-1}{RC} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} \frac{1}{L} \\ 0 \end{bmatrix} V_{inv} \quad (\text{II.19})$$

where  $R$  is the load resistance.

➤ **RL load** — the filter inductor and its series resistance are denoted  $L_f$  and  $r_f$ , and the load inductor  $L_{ch}$ :

$$\begin{bmatrix} \frac{dx_1}{dt} \\ \frac{dx_2}{dt} \\ \frac{dx_3}{dt} \end{bmatrix} = \begin{bmatrix} \frac{-r_f}{L_f} & \frac{-1}{L_f} & 0 \\ \frac{1}{C} & 0 & \frac{-1}{C} \\ 0 & \frac{1}{L_{ch}} & \frac{-R}{L_{ch}} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} + \begin{bmatrix} \frac{1}{L_f} \\ 0 \\ 0 \end{bmatrix} V_{inv} \quad (\text{II.20})$$

where  $L_f$  and  $r_f$  are the inductance and internal resistance of the filter inductor,  $L_{ch}$  the load inductance,  $R$  the load resistance, and  $x_3$  the third state variable (the load current).

The voltage delivered by the bridge is linked to the switch control through the switching function:

$$v_{inv} = S(t) V_{dc} \quad (II.21)$$

$$S(t) = 2 \text{ PWM}(t) - 1 \quad (II.22)$$

where  $S(t)$  is the switching state derived from the PWM signal, with  $\text{PWM}(t) \in \{0, 1\}$ .

### II.4.3.1.1 Simulink Implementation

To validate the theoretical mathematical modeling, the system is simulated using MATLAB/Simulink based on the state-space equations derived in Section II.4.3.1.

#### ➤ No-Load Inverter

The simulation is carried out over a total duration of 5 seconds. The reference voltage amplitude is initially set to 60 V RMS for the first 2.5 seconds, corresponding to a peak reference of  $V_{ref} = 60\sqrt{2} \cdot \sin(100\pi t)$  V, then stepped up to 220 V RMS at  $t = 2.5$  s, in order to observe the system's dynamic response to a sudden change in the operating point.

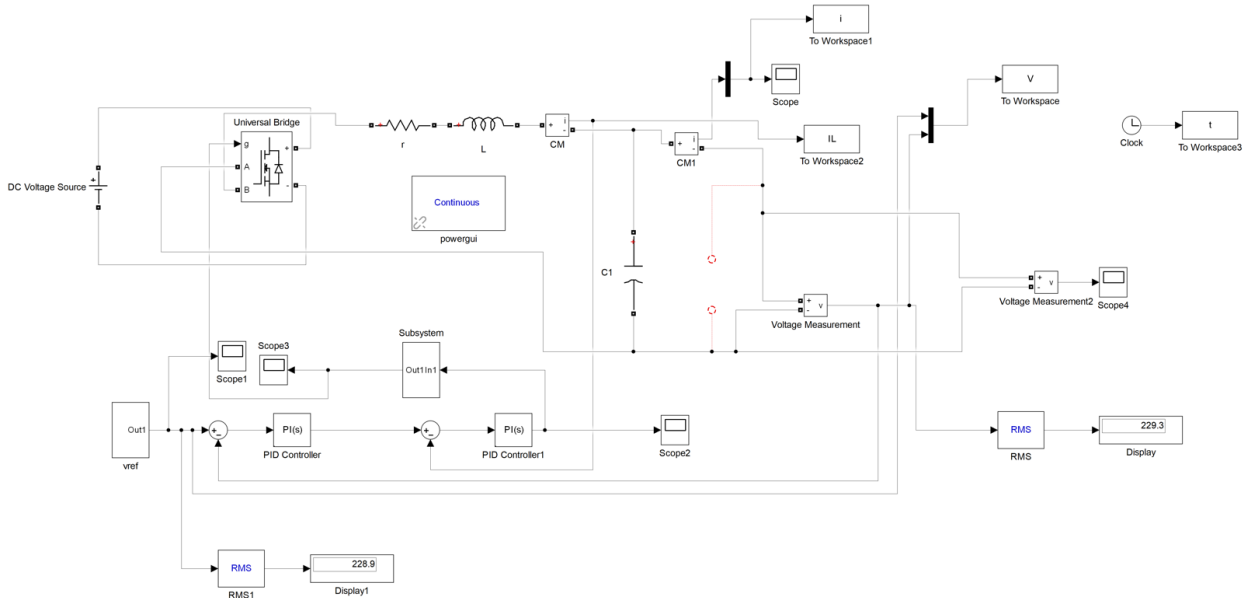


**Fig. II.13:** Simulink block diagram of the closed-loop single-phase inverter (No-Load).

This diagram implements the H-bridge inverter model using a Universal Bridge block (MOSFET/Diodes, 2 arms) supplied by a DC voltage source of  $V_{dc} = 400$  V. The PWM signal is generated by comparing the sinusoidal reference with a triangular carrier at 5000 Hz. The output LC filter parameters are  $L = 1$  mH and  $r = 0.3 \Omega$ . A double closed-loop PI control strategy is used, consisting of an outer voltage loop ( $K_p = 10$ ,  $K_i = 100$ ) and an inner current loop ( $K_p = 6.28$ ,  $K_i = 3947.8$ ), with no load connected at the output. The output voltage  $V_C(t)$  and the inductor current  $I_L(t)$  are exported to the MATLAB workspace for visualization.

The corresponding simulation results representing the waveforms of the output voltage  $V_C(t)$  and the inductor current  $I_L(t)$  under no-load conditions are illustrated below:

This diagram implements the H-bridge inverter model using a Universal Bridge block (MOSFET/Diodes, 2 arms) supplied by a DC voltage source of  $V_{dc} = 400$  V. The PWM signal is generated by comparing the sinusoidal reference with a triangular carrier at 5000 Hz. The output LC filter parameters are  $L = 1$  mH and  $r = 0.3 \Omega$ , with no load connected at the output. A double closed-loop PI control strategy is applied, consisting of an outer voltage loop ( $K_p = 10$ ,  $K_i = 100$ ) and an inner current loop ( $K_p = 6.28$ ,  $K_i = 3947.8$ ). The output voltage  $V_C(t)$  and the inductor current  $I_L(t)$  are exported to the MATLAB workspace for visualization.

The corresponding simulation results representing the waveforms of the output voltage  $V_C(t)$  and the inductor current  $I_L(t)$  under no-load conditions are illustrated below:

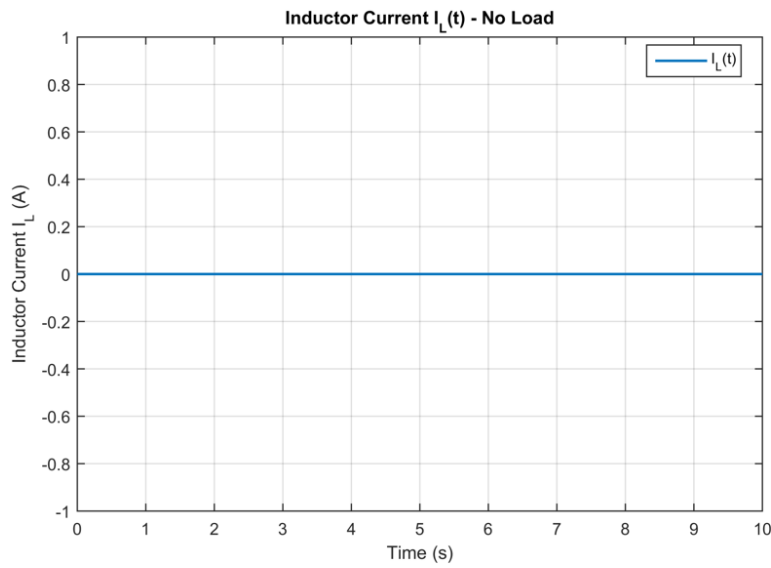


Fig. II.14: Waveform of the inductor current  $I_L(t)$  — No-Load Inverter.

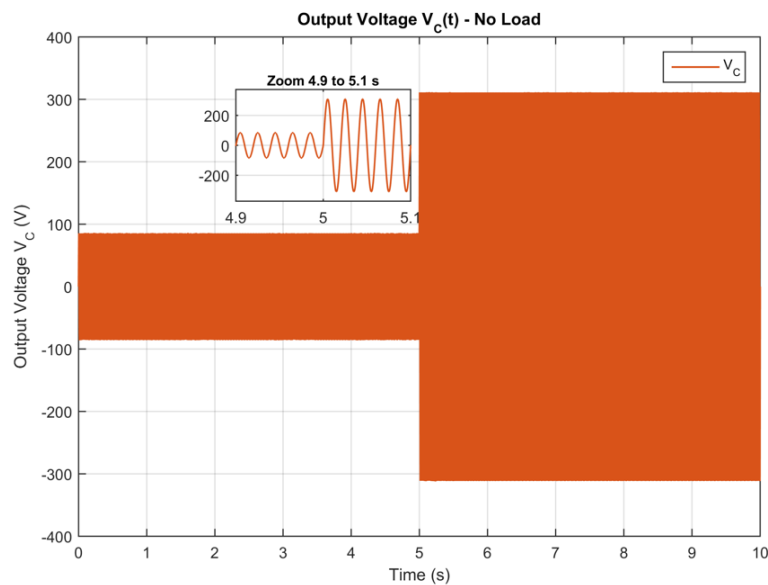


Fig. II.15: Output voltage  $V_C(t)$  under no-load condition with zoom around 5 s.

### ➤ Loaded Inverter — $R = 50 \Omega$

To investigate the effect of the resistive load on the system behavior, the simulation is carried out with  $R = 50 \Omega$  connected at the inverter output. A double closed-loop PI control strategy is applied under the same operating conditions, and the results are presented below.

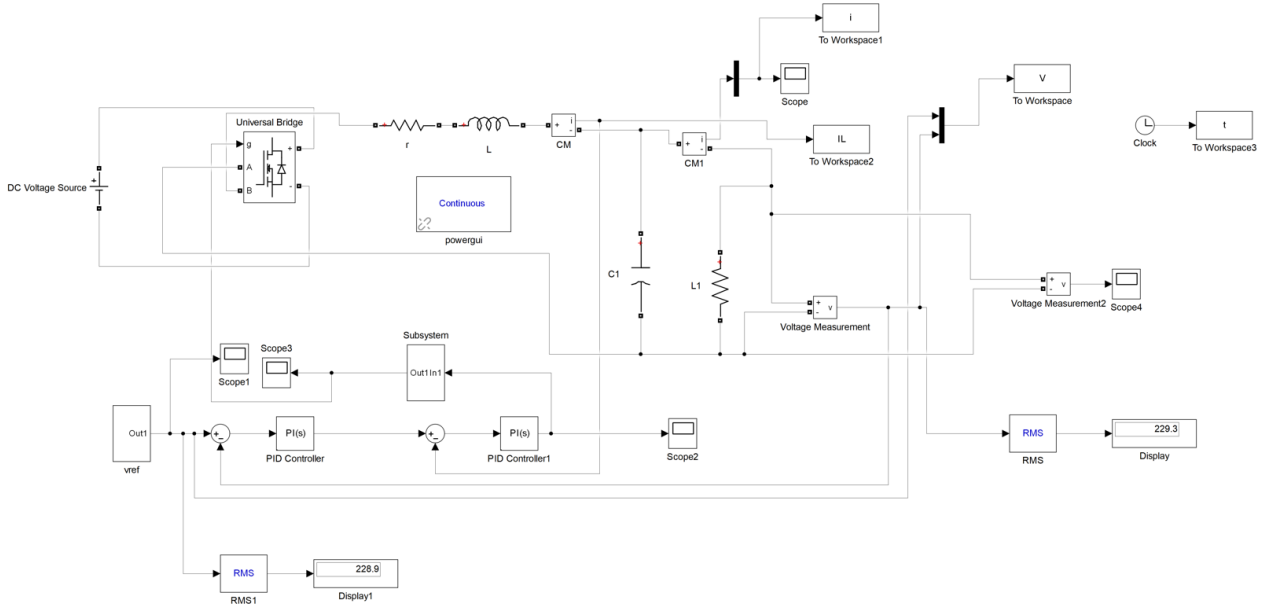


Fig. II.16: Simulink block diagram of the closed-loop single-phase inverter ( $R = 50 \Omega$ ).

This diagram implements the H-bridge inverter model using a Universal Bridge block (MOSFET/Diodes, 2 arms) supplied by a DC voltage source of  $V_{dc} = 400 \text{ V}$ . The PWM signal is generated by comparing the sinusoidal reference with a triangular carrier at 5000 Hz. The output LC filter parameters are  $L = 1 \text{ mH}$  and  $r = 0.3 \Omega$ , with a resistive load of  $R = 50 \Omega$  connected at the output. The double PI controller consists of an outer voltage loop ( $K_p = 10$ ,  $K_i = 100$ ) and an inner current loop ( $K_p = 6.28$ ,  $K_i = 3947.8$ ). The output voltage  $V_C(t)$  and the load current  $I_{Load}(t)$  are exported to the MATLAB workspace for visualization.

The corresponding simulation results are illustrated below:

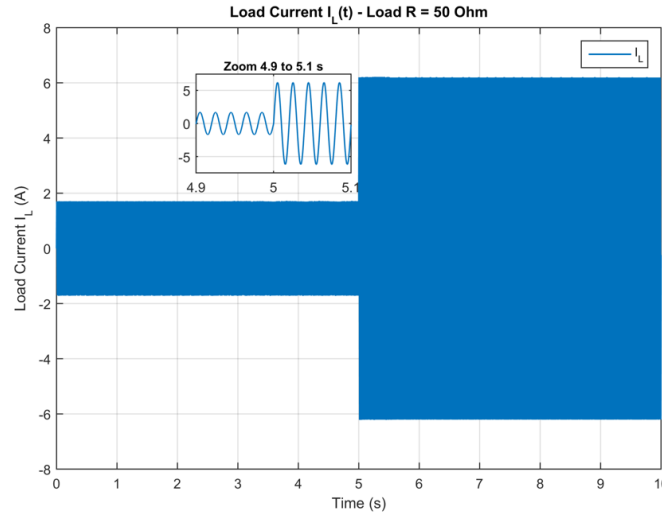


Fig. II.17: Load current  $I_{Load}(t)$  — Loaded Inverter ( $R = 50 \Omega$ ).

The load current follows  $I_{Load} = V_C / R$ , with a peak amplitude of approximately 1.7 A during the first 5 seconds, then stepping up to approximately 6.2 A after  $t = 5$  s. The transition is smooth and well-controlled, confirming the effectiveness of the double PI regulation strategy under resistive loading conditions.

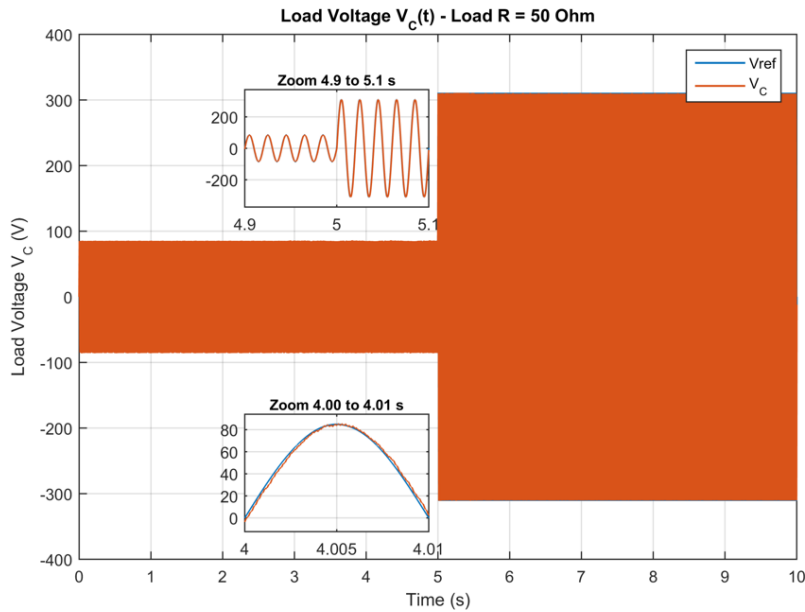


Fig. II.18: Output voltage  $V_C(t)$  — Loaded Inverter ( $R = 50 \Omega$ ).

The output voltage is well-regulated, with a peak amplitude of approximately 85 V (corresponding to 60 V RMS) during the first 5 seconds. At  $t = 5$  s, following the reference step to 220 V RMS, the voltage transitions smoothly to a peak of approximately 311 V, with a very short transient period visible in the zoom window. The double PI controller ensures accurate tracking of the sinusoidal reference under load.

### II.4.3.2 Three-Phase Inverter

The three-phase voltage source inverter is generally supplied by a DC-link voltage obtained from a rectifier and filtered by an LC filter. Before describing the inverter switching stage, the DC-link voltage produced by the diode rectifier can be expressed, by neglecting commutation overlap, as the difference between the maximum and minimum values of the three-phase input voltages:

$$U_{\text{red}}(t) = \text{Max}[V_a, V_b, V_c] - \text{Min}[V_a, V_b, V_c] \quad (\text{II.23})$$

where  $U_{\text{red}}$  is the rectifier output voltage, and  $V_a$ ,  $V_b$  and  $V_c$  are the three-phase input voltages.

The DC-to-AC conversion is described by the transformation matrix  $T$ , which relates the load phase voltages to the inverter leg voltages for a balanced star-connected load:

$$T = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \quad (\text{II.24})$$

where  $T$  is the inverter transformation matrix.

For example, the voltage of phase a is obtained as:

$$V_{\text{an}} = \frac{2}{3}V_{\text{ao}} - \frac{1}{3}V_{\text{bo}} - \frac{1}{3}V_{\text{co}} \quad (\text{II.25})$$

where  $V_{\text{an}}$  is the phase voltage of phase a of the load, and  $V_{\text{ao}}$ ,  $V_{\text{bo}}$  and  $V_{\text{co}}$  are the inverter pole (leg) voltages.

the leg voltages being referenced to the fictitious midpoint of the DC bus. The DC voltage is smoothed by a second-order LC filter, whose transfer function and cut-off frequency are:

$$F(s) = \frac{1}{1 + L_f C_f s^2} \quad (\text{II.26})$$

$$f_c = \frac{1}{\sqrt{L_f C_f}} \quad (\text{II.27})$$

where  $F(s)$  is the filter transfer function,  $L_f$  and  $C_f$  the inductance and capacitance of the LC filter,  $s$  the Laplace variable, and  $f_c$  the filter cut-off frequency.

which attenuates the voltage ripple delivered by the rectifier.

#### II.4.3.2.1 Simulink Implementation

To validate the theoretical operation of the three-phase voltage source inverter, the system is simulated using MATLAB/Simulink. The objective of this simulation is to observe the behavior of the inverter output voltages and currents under different operating conditions. In this part, the inverter is studied without closed-loop regulation, meaning that no feedback controller is used. The switching signals are generated directly using sinusoidal pulse width modulation (SPWM).

The three-phase inverter is implemented using a Universal Bridge block supplied by a DC voltage source. The SPWM technique is based on comparing three sinusoidal reference signals with a high-frequency triangular carrier. The three reference signals have the same frequency and amplitude, and

they are phase-shifted by  $120^\circ$  with respect to each other. This allows the inverter to generate a balanced three-phase voltage system.

### ➤ No-Load Three-Phase Inverter

In the first simulation case, the three-phase inverter is tested under no-load conditions. This test is performed in order to verify the correct generation of the three-phase output voltages before connecting a load to the inverter. The simulation is carried out over a total duration of 10 seconds. During the first 5 seconds, the inverter operates at the initial reference voltage level. At  $t = 5$  s, a step change is applied to the reference voltage amplitude in order to observe the dynamic response of the inverter.

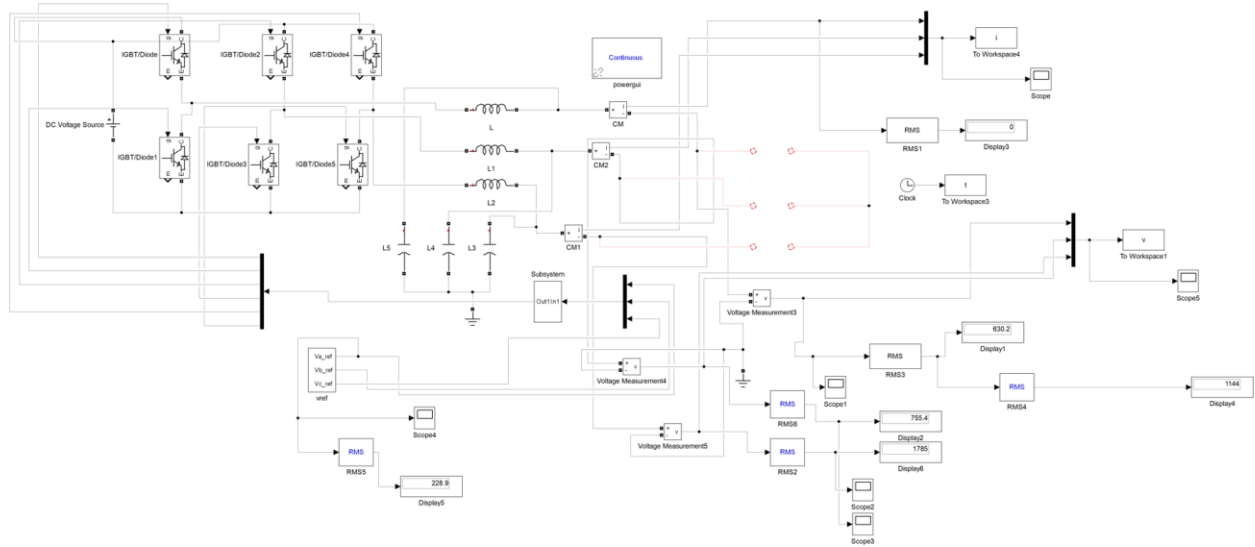


**Fig. II.19:** Simulink block diagram of the three-phase inverter under no-load condition.

This diagram represents the three-phase voltage source inverter operating without load. The inverter is supplied by a DC voltage source and controlled using sinusoidal PWM signals. The three sinusoidal reference voltages are balanced and shifted by  $120^\circ$  from each other, then compared with a triangular carrier signal to generate the switching pulses. Since no load is connected at the output, the main objective of this simulation is to verify the balance and phase displacement of the output voltages. The three-phase output voltages  $V_a(t)$ ,  $V_b(t)$ , and  $V_c(t)$  are exported to the MATLAB workspace for visualization.

The corresponding simulation results representing the three-phase output voltages under no-load conditions are illustrated below.

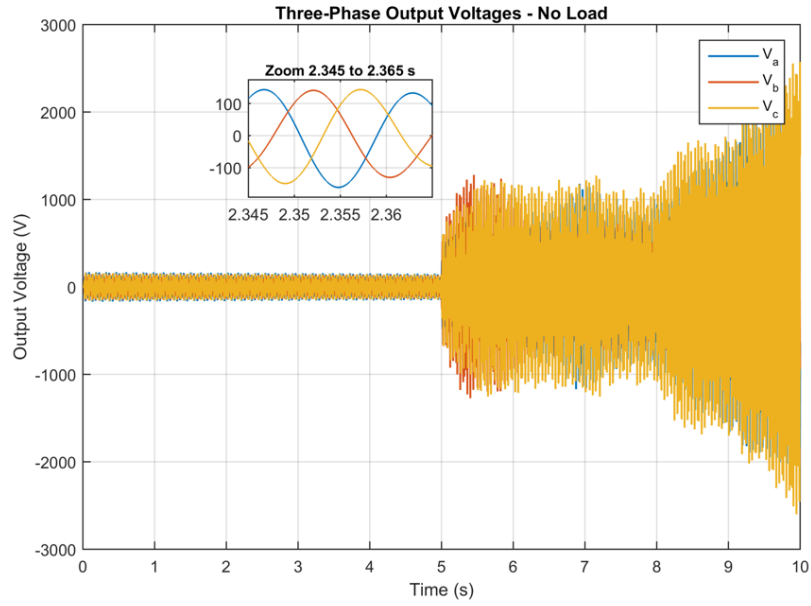


Fig. II.20: Three-phase output voltages  $V_a$ ,  $V_b$ , and  $V_c$  under no-load condition.

The zoomed view shows the phase displacement between the three output voltages. However, after the reference step at  $t = 5$  s, the voltage waveforms present noticeable distortion and a significant increase in amplitude. This behavior is mainly due to the open-loop operation of the inverter, where no feedback regulation is used to correct the output voltage.

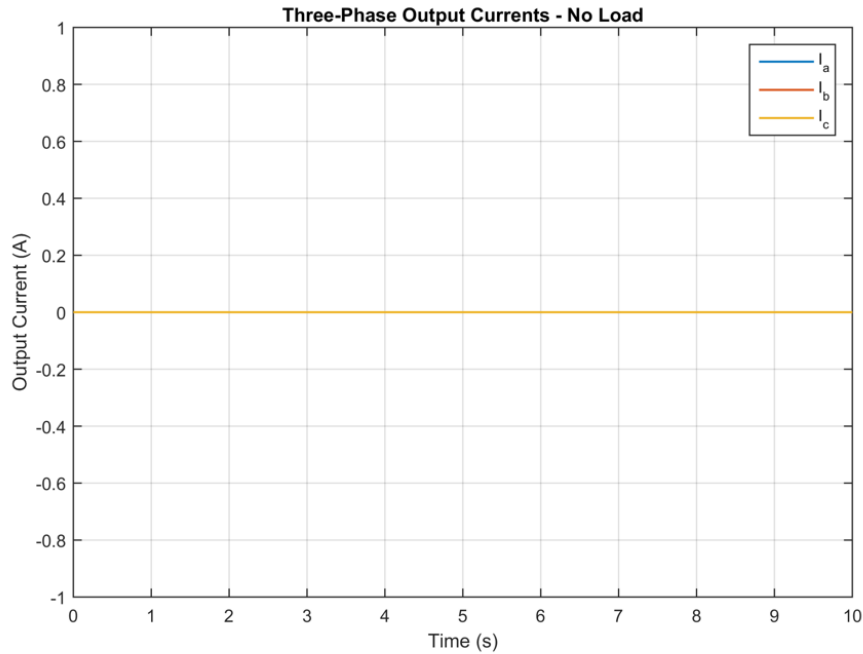


Fig. II.21: Three-phase output currents  $I_a$ ,  $I_b$ , and  $I_c$  under no-load condition.

Under no-load conditions, the three-phase output currents  $I_a$ ,  $I_b$ , and  $I_c$  remain equal to zero because no load is connected to the inverter output.

### ➤ Loaded Three-Phase Inverter — $R = 50 \Omega$

In the second simulation case, a balanced three-phase resistive load is connected to the inverter output in order to study the behavior of the system under loaded conditions. The load resistance is set to  $R = 50 \Omega$  for each phase. The simulation is carried out over a total duration of 10 seconds. The reference voltage amplitude is initially set to 60 V RMS, then increased to 220 V RMS at  $t = 2$  s using a Step block. This test allows the observation of the inverter response when the reference voltage changes under load operation.

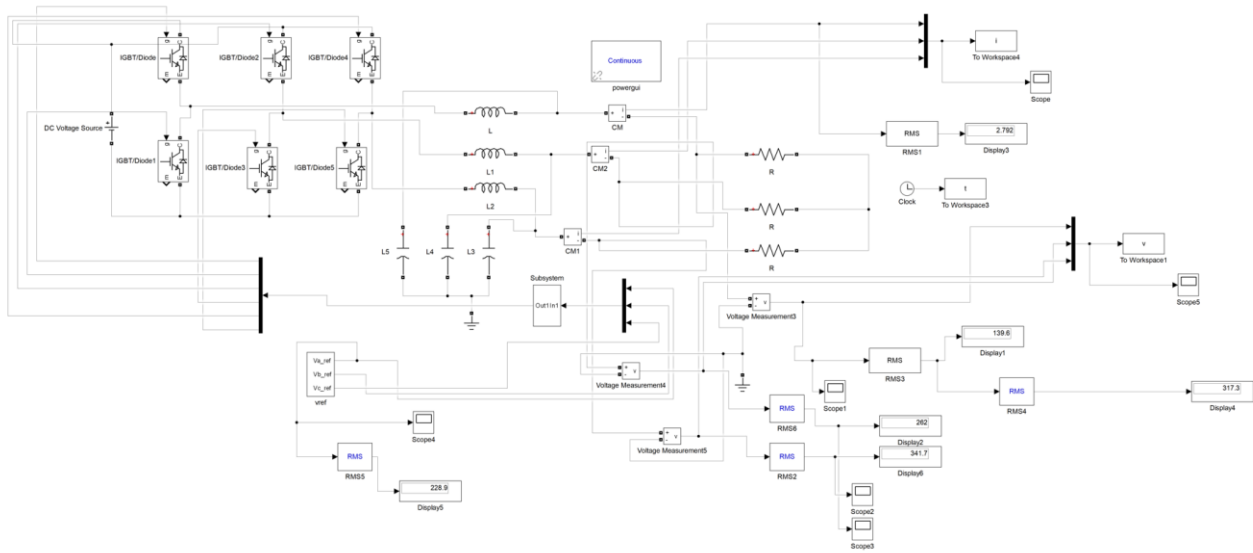


Fig. II.22: Simulink block diagram of the three-phase inverter under load condition  $R = 50 \Omega$ .

The Simulink model represents a three-phase voltage source inverter supplied by a DC voltage source and controlled using sinusoidal PWM. The three reference signals are sinusoidal waveforms shifted by  $120^\circ$  from each other and compared with a triangular carrier signal to generate the switching pulses. A balanced three-phase resistive load of  $R = 50 \Omega$  is connected to the inverter output. Since no closed-loop regulation is used, the inverter operates in open-loop mode. The three-phase output voltages  $V_a(t)$ ,  $V_b(t)$ , and  $V_c(t)$ , as well as the load currents  $I_a(t)$ ,  $I_b(t)$ , and  $I_c(t)$ , are exported to the MATLAB workspace for visualization.

The corresponding simulation results are presented below.

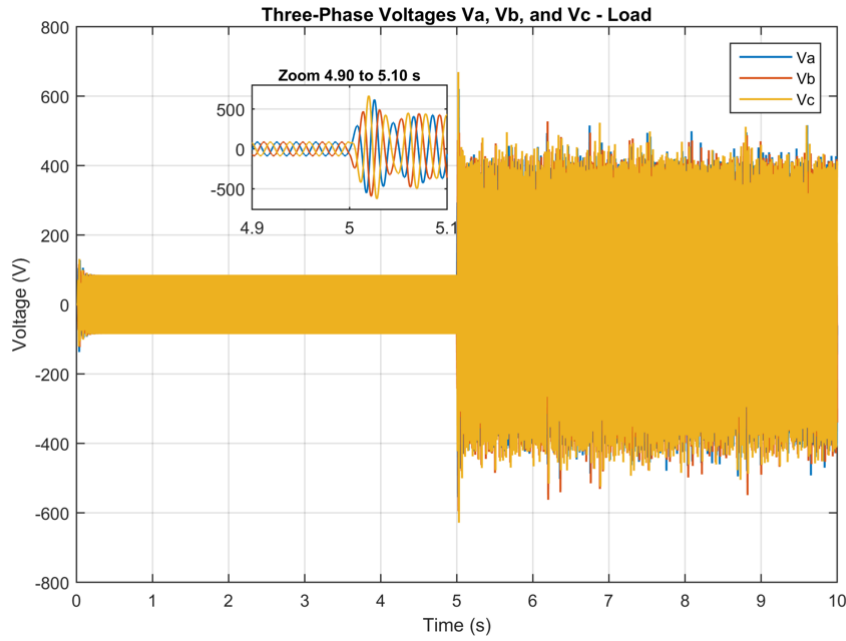


**Fig. II.23:** Three-phase output voltages  $V_a$ ,  $V_b$ , and  $V_c$  under load condition  $R = 50 \, \Omega$  with zoom around 5 s.

The obtained waveforms show the three-phase output voltages under load condition. Before  $t = 5$  s, the voltage amplitude remains relatively low. After the reference step, a clear increase in voltage amplitude is observed, accompanied by noticeable switching ripples due to the open-loop SPWM operation. The zoomed view around 5 s highlights the transient response during the reference voltage change.

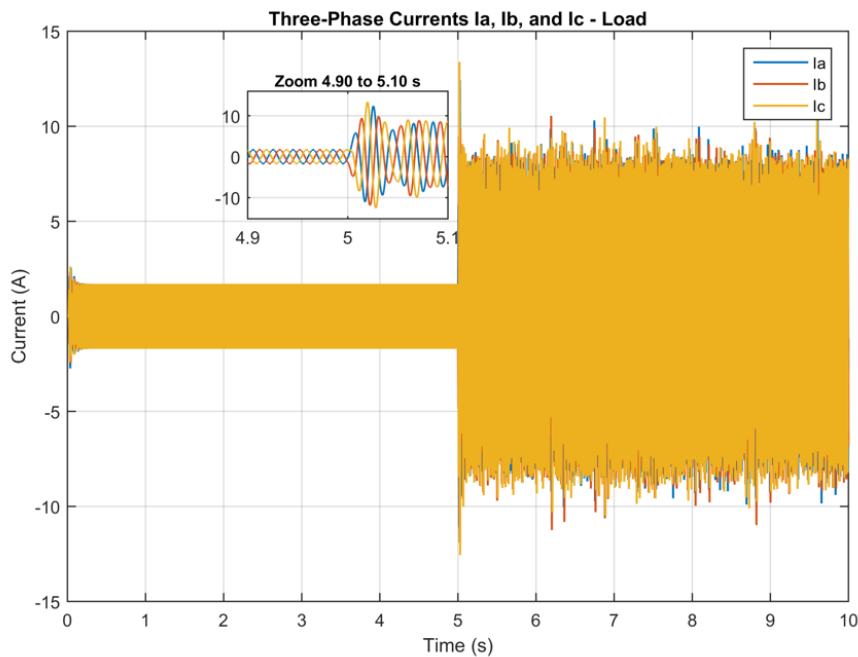


**Fig. II.24:** Three-phase load currents  $I_a$ ,  $I_b$ , and  $I_c$  under load condition  $R = 50 \, \Omega$  with zoom around 5 s.

The obtained waveforms show the three-phase load currents under  $R = 50 \, \Omega$ . Before  $t = 5 \, \text{s}$ , the current amplitude is relatively low. After the reference voltage step, the current amplitude increases clearly due to the increase in the output voltage. The zoomed view around 5 s highlights the transient response of the load currents during the reference change. The presence of switching ripples is due to the open-loop SPWM operation of the inverter.

#### II.4.4 Pulse Width Modulation

Pulse Width Modulation, usually called **PWM**, is one of the most widely used control techniques in power electronic converters. Its main idea is to control the output voltage by switching the power devices ON and OFF at high frequency. Instead of using a continuously variable voltage, the converter applies a series of pulses with constant amplitude, while the width of these pulses is changed according to the required output value.

In an electric drive system, PWM represents the link between the control stage and the power converter. The controller calculates the required control signal, then the PWM block transforms this command into gate signals used to control the switches of the converter. In this way, the motor receives a controlled voltage while the power switches operate mainly in ON/OFF mode, which improves the efficiency of the system.

The basic parameter of PWM is the **duty cycle**. It represents the ratio between the ON time of the switch and the total switching period. It can be written as:

$$\alpha = \frac{t_{\text{on}}}{T} = \frac{t_{\text{on}}}{t_{\text{on}} + t_{\text{off}}} \quad (\text{II. 28})$$

where  $\alpha$  is the duty cycle,  $t_{\text{on}}$  is the ON time,  $t_{\text{off}}$  is the OFF time, and  $T$  is the switching period.

For a simple switching converter, the average output voltage can be expressed approximately as:

$$V_{\text{avg}} = \alpha \cdot V_{\text{dc}} \quad (\text{II. 29})$$

This relation shows that increasing the duty cycle increases the average output voltage, while decreasing it reduces the average output voltage. As shown in **Fig. II.25**, narrow pulses produce a lower average voltage, while wider pulses produce a higher average voltage.

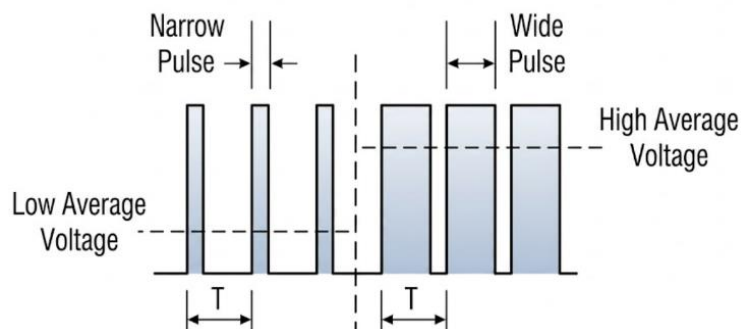


**Fig. II.25:** Effect of duty cycle on the average output voltage [12].

Therefore, PWM allows smooth voltage control without using dissipative elements such as series resistors.

In a three-phase inverter, PWM is used to control the six power switches. A common method consists of comparing a reference signal with a high-frequency triangular carrier signal. When the reference signal is greater than the carrier signal, the corresponding switch is turned ON, and when it becomes lower, the switch is turned OFF. As illustrated in **Fig. II.26**, this comparison process generates a sequence of switching pulses that controls the inverter operation.

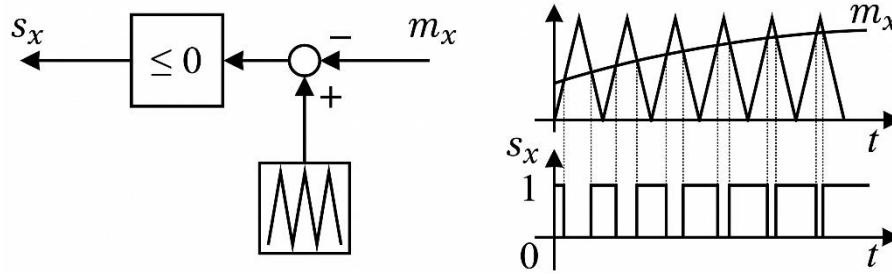


**Fig. II.26:** Principle of PWM signal generation using a carrier comparison method [1].

By changing the reference signal, the inverter can generate controlled output voltages for the three motor phases. In AC motor drives, the objective is not only to obtain an average voltage, but also to produce a waveform that is close to a sinusoidal voltage. This is achieved by varying the duty cycle during each electrical period. The pulses become wider near the peak of the sinusoidal reference and narrower near the zero-crossing. Due to the inductance of the motor windings, the high-frequency switching components are naturally smoothed, and the motor current becomes closer to a sinusoidal waveform.

Thus, PWM is an essential technique in modern electric drives. It allows efficient control of the converter, reduces power losses, and makes it possible to adjust the voltage and frequency applied to the motor according to the required operating conditions.

## II.5 Control System

After studying the power converters used in electric drive systems, it is necessary to explain how these converters are controlled. The control system represents the decision-making part of the intelligent electric drive system. Its main role is to make the motor follow the required operating conditions, such as speed, torque, or position.

As previously shown in **Fig. II.2**, the motor controller receives different feedback signals from the system, such as the input voltage  $u$ , the motor current  $i$ , the mechanical speed  $\Omega_r$ , and the rotor position  $\theta$ . These measured values are compared with the desired reference values in order to determine the error between the required and actual behavior of the motor.

After this comparison, the controller generates suitable control signals for the power converter. In most modern electric drives, these signals are sent in the form of **PWM pulses**, which control the switching devices of the converter. In this way, the converter can adjust the voltage, current, or frequency according to the motor type and control strategy. according to the required operating condition.

Through this control action, the system can improve speed regulation, torque response, accuracy, and stability. The controller can also contribute to protection by monitoring abnormal conditions such as overcurrent, overvoltage, or overheating, and by reducing or stopping the converter operation when necessary.

### II.5.1 Open-Loop Control

Open-loop control is the simplest control mode used in electric drive systems. In this type of control, the controller sends commands to the power converter without using feedback from the actual motor output. In other words, the system does not measure the real speed, torque, or position of the motor, and therefore it cannot automatically correct the error between the desired and actual values.

As shown in **Fig. II.27**, the open-loop structure consists of a reference input, a controller, and a plant or process. The controller generates the actuating signal, which is applied to the system in order to obtain the required output. Since there is no feedback path, the controller does not know whether the motor has really reached the desired operating condition.

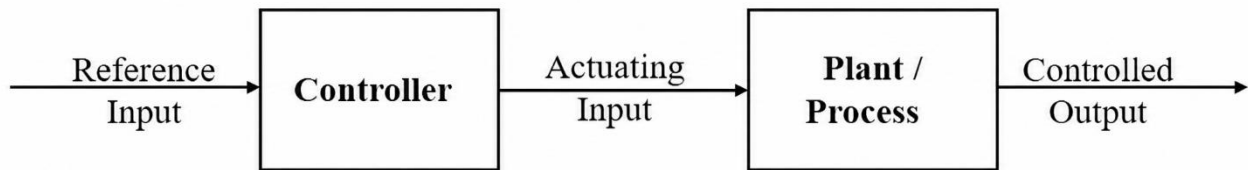


**Fig. II.27:** Basic block diagram of an open-loop control system [18].

The main advantage of open-loop control is its simplicity. It is easy to implement, requires fewer components, and has a relatively low cost because feedback sensors are not needed. For this reason, open-loop control can be used in applications where high precision is not required.

However, this control mode has important limitations. Its accuracy is generally low, especially when the load changes or when the system is affected by disturbances. Since the controller does not receive feedback, it cannot compensate automatically for variations in speed or torque. As a result, the motor performance may deviate from the desired value.

A common example of open-loop control in electric drives is the **V/f control** of an induction motor. In this method, the inverter applies a predefined voltage and frequency to the motor without speed feedback. If the load increases, the actual motor speed may decrease, but the controller continues to send the same command because no feedback is available.

### II.5.2 Closed-Loop Control

To address the limitations inherent in open-loop systems most notably the reduction in motor speed under varying mechanical loads modern high-performance electric drives predominantly adopt closed-loop control strategies. In this configuration, as illustrated in Fig. II.28, the control action is not independent but is continuously determined by the system output. The system therefore operates as a dynamic feedback cycle, constantly monitoring the actual motor performance and applying real-time corrections to maintain the desired operating conditions.

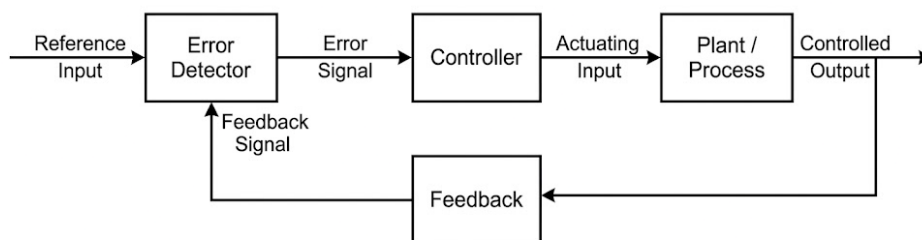


**Fig. II.28:** Basic block diagram of a closed-loop control system [18].

The fundamental characteristic of this architecture is the feedback mechanism. A variety of sensors are employed to acquire real-time data from the drive system; for example, incremental encoders or tachometers are used to measure rotational speed, while Hall-effect sensors are utilized to monitor motor current. As shown in the block diagram, these feedback signals are transmitted to an error detection unit, where the measured output is compared with the reference input (setpoint provided by the user). The resulting difference constitutes the error signal.

Once the error is determined, the controller processes it to generate the appropriate corrective action by adjusting the actuator input, such as modifying the PWM signals supplied to the power converter. For instance, if an increased load causes a drop in motor speed, the controller immediately detects the deviation and responds by increasing the voltage or frequency through the converter, thereby restoring the motor speed to its reference value.

Closed-loop control systems offer significant advantages in terms of accuracy and robustness. They are capable of automatically compensating for:

- External disturbances, such as sudden variations in mechanical load.
- Internal parameter variations, such as changes in motor winding resistance due to temperature rise.

As a result, this feedback-based architecture provides considerably improved stability and reliability compared to open-loop systems. Although closed-loop control introduces additional complexity and higher implementation costs due to the requirement for sensing devices and advanced processing units, its superior performance makes it the preferred standard in modern intelligent motor drive applications.

### **II.5.3 Speed and Current Control**

After studying open-loop and closed-loop control systems, it is important to explain how these principles are applied to the two main controlled variables in an electric drive: speed and current. In high-performance drive systems, these variables are commonly controlled using a cascaded structure, composed of a fast inner current loop and a slower outer speed loop.

The current control loop is directly related to torque production. In DC motors, the electromagnetic torque is proportional to the armature current, while in vector-controlled AC motors it is mainly linked to the torque-producing current component. Therefore, controlling the current allows the drive to control the torque developed by the motor. In addition, this loop plays an important protective role, since motor current can rise quickly during overloads, transients, or stall conditions. By limiting the current reference to a safe value, the controller protects the motor windings and the power electronic converter.

The speed control loop works at a higher level. Its main purpose is to maintain the motor speed close to the desired reference, even when the load torque changes. The reference speed is compared with the measured speed, obtained for example from an encoder or tachometer. The resulting speed error is processed by a PI controller, which generates the current reference for the inner loop. If the motor speed decreases because of an increased load, the controller increases the current reference, allowing the motor to produce more torque and recover its speed.

In general, the cascaded control process starts by comparing the desired and measured speeds. The speed controller then produces a current reference, which is compared with the measured current. The

current controller generates the voltage command, which is finally converted into PWM signals for the power converter. This structure gives the drive a fast torque response, accurate speed regulation, and better protection against abnormal operating conditions. The overall cascaded speed and current control structure of an electric drive is illustrated in **Fig. II.29**.

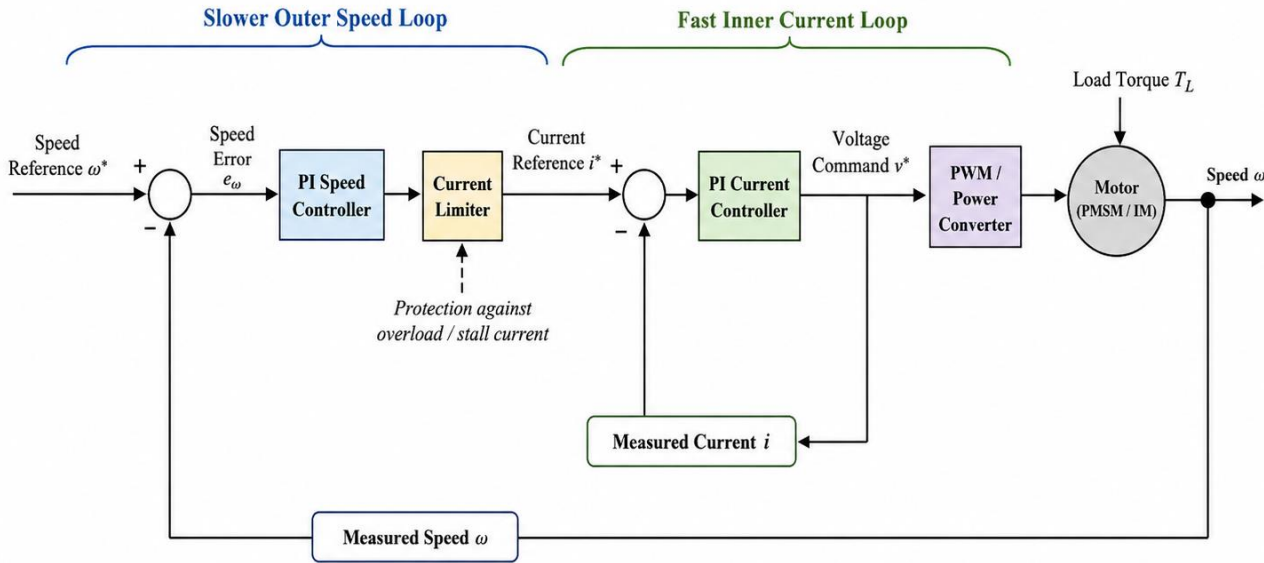


**Fig. II.29:** Cascaded speed and current control structure of an electric drive.

### II.5.4 Intelligent Control Strategy

After discussing the cascaded speed and current control loops, it is important to introduce the role of intelligent control in modern electric drive systems. Classical controllers, such as PI and PID controllers, are widely used because they are simple, reliable, and easy to implement. They provide satisfactory performance in many applications, especially when the system parameters are well known and the operating conditions remain relatively constant. However, their performance may decrease when the drive operates under variable load conditions, parameter changes, or nonlinear effects, which are frequently encountered in practical industrial environments.

In real industrial applications, the behavior of the motor is not always perfectly predictable. For example, the load torque may change suddenly, the winding resistance may vary with temperature, and nonlinear phenomena such as friction, magnetic saturation, or converter switching effects may appear. These factors make the drive system more complex and may reduce the effectiveness of fixed-parameter classical controllers. In such cases, a controller designed only on the basis of a simplified mathematical model may not always ensure the best dynamic response, especially when fast adaptation and robustness are required.

For this reason, intelligent control strategies are introduced to improve the robustness, adaptability, and overall performance of the electric drive. Instead of depending only on an exact mathematical model, these methods can use expert knowledge, data, learning capability, or adaptive mechanisms to make better control decisions. Their main objective is to maintain good speed regulation, reduce overshoot, improve response time, and increase the stability of the system under changing operating conditions. In this way, intelligent control makes the drive more suitable for real applications where disturbances and uncertainties are unavoidable.

Among the most widely used intelligent control methods, fuzzy logic control is one of the most common. It is based on linguistic rules such as “IF the speed error is high, THEN increase the control action.” This approach is especially useful for systems that are difficult to model accurately or that exhibit significant nonlinear behavior. Artificial neural networks can also be used because they are able to learn complex relationships between inputs and outputs from data, which allows them to adapt to system variations. In addition, adaptive and hybrid controllers, such as Fuzzy-PI or Fuzzy-PID controllers, combine the simplicity of classical control with the flexibility and intelligence of advanced techniques, thus providing better performance than conventional regulators alone.

In an intelligent electric drive, sensors remain essential because they provide real-time information about important variables such as speed, current, voltage, and sometimes temperature. These measured signals are processed by a digital controller, such as a DSP, microcontroller, or industrial control unit. Based on this information, the intelligent algorithm evaluates the operating state of the drive and adjusts the control commands sent to the power converter. As a result, the drive becomes more capable of responding effectively to disturbances, load variations, and parameter changes, while maintaining stable and accurate operation.

Therefore, intelligent control can be considered an important improvement in electric drive systems. It does not completely replace classical control, but rather enhances it by making the system more flexible, stable, and suitable for complex industrial applications. The general structure of this intelligent control approach, including the interaction between the reference input, the intelligent controller, the digital control unit, the sensors, the power converter, and the motor drive, is illustrated in Fig. II.30.

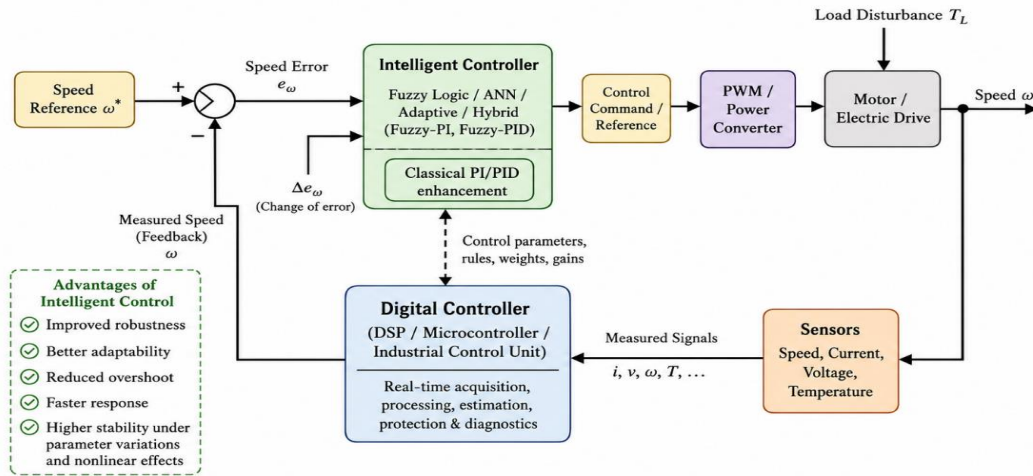


**Fig. II.30:** General structure of an intelligent control strategy for an electric drive system.

## II.6 Sensors and Protection Systems

To implement the control strategies detailed in the previous section, the system requires a continuous flow of information about its current state. Sensors and protection systems fulfill this role by acting as the feedback and safety interface of the intelligent electric drive.

As shown in Fig. II.31, sensors collect the main measured signals, such as speed, current, voltage, and temperature, and send them back to the digital controller. This feedback allows the controller to monitor the motor operation and adjust the power converter according to the required speed or torque reference.

Sensors are used to measure critical parameters and provide the controller with the necessary information to regulate the motor operation with high precision. At the same time, protection systems are integrated to maintain safe and reliable operation by detecting and reacting to abnormal conditions, such as voltage surges, overcurrents, or overheating, which could cause hardware damage if not handled in time.

By combining accurate measurement with fault detection, these systems allow the drive to operate in a more reliable and intelligent way.

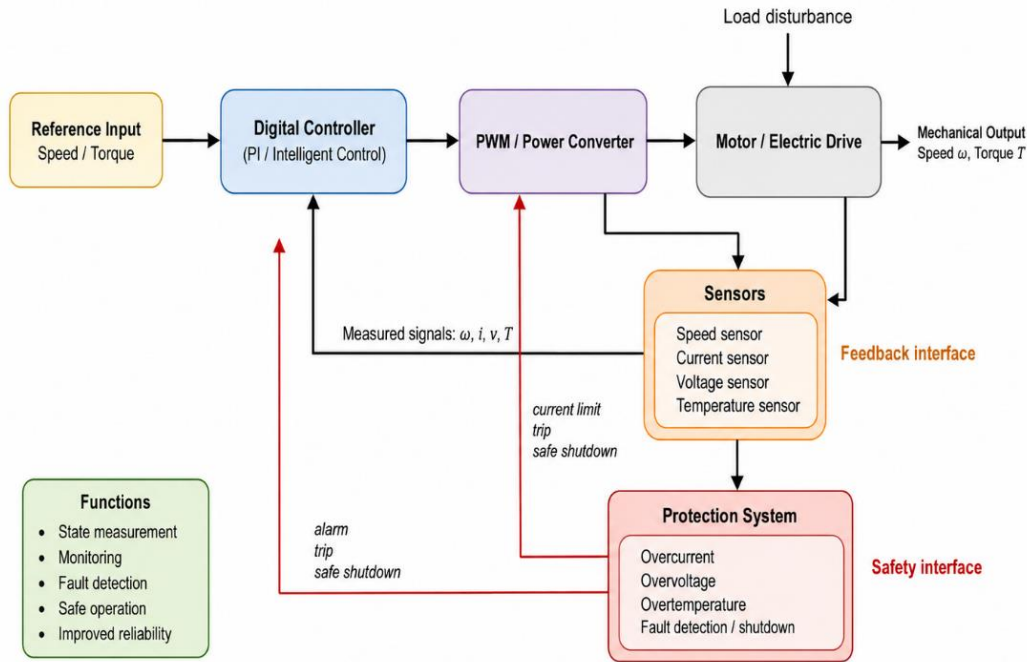


**Fig. II.31:** Main sensor positions and measurement feedback paths in an intelligent electric drive system.

### II.6.1 Current and Voltage Sensors

In order to achieve the intelligent behavior described in the previous section, the system must continuously acquire electrical data to close the control loops and ensure safe operation. Current and voltage sensors are the main components used for this feedback process. They convert high-power electrical quantities into low-level signals that can be interpreted by the digital signal processor (DSP) or the controller for real-time adjustment.

As shown in Fig. II.32, current and voltage sensors are placed at key points of the drive system, such as the DC-link, the converter output, and the motor phase lines. Their measured signals are sent back to the digital controller, which uses them to regulate the power converter and protect the system during operation.

Current sensors play an important role in motor torque regulation and drive protection. In modern drive systems, sensors such as Hall effect transducers or Rogowski coils are used to measure the phase currents with good accuracy. This information is essential because the torque developed by the motor is directly related to the current level.

In addition, current sensors contribute to overcurrent protection. They allow the system to detect short circuits or overloads and send a fault signal to stop the converter or limit the current before hardware

damage occurs. They also help supervise the static converter by monitoring the current flowing through the power semiconductor switches.

Voltage sensors are also important, especially for monitoring the DC-link voltage and the converter output voltage. Measuring the DC bus voltage helps ensure the stability of the power stage and protects the system against abnormal voltage conditions, such as overvoltage or undervoltage.

By providing accurate information about supply and terminal voltages, voltage sensors help the controller manage the PWM signals correctly. Together, current and voltage sensors provide the electrical feedback needed for reliable, safe, and high-performance operation of the intelligent electric drive.

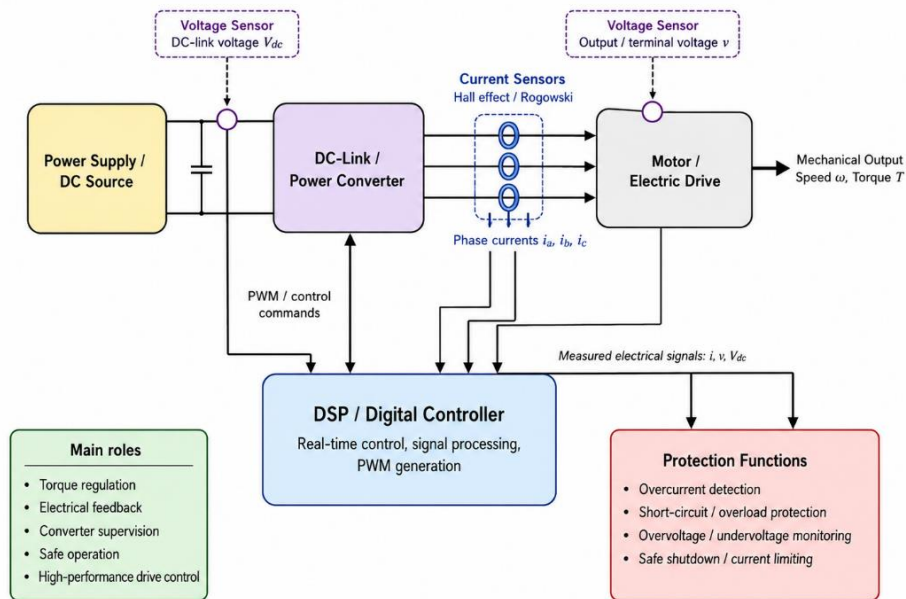


**Fig. II.32:** Current and voltage sensing in an intelligent electric drive system.

## II.6.2 Electrical Protection

In addition to the monitoring and control functions provided by the sensors described in the previous subsections, an intelligent drive must include electrical protection systems to ensure safe and reliable operation. These systems protect the motor, the power electronic converter, and the control hardware from damage when the drive is exposed to electrical stress or fault conditions.

As shown in Fig. II.33, the protection system receives measured signals from the current and voltage sensors, including the DC-link voltage. These signals are continuously monitored and compared with predefined safety limits. When an abnormal condition is detected, a fault signal is sent to the DSP or digital controller.

The protection system is designed to detect several electrical faults that may occur in industrial environments. The most common faults include short circuits, overloads, overcurrent, overvoltage, and undervoltage. Short circuits are especially dangerous because they can quickly damage the power semiconductor switches. Overloads and voltage problems can also affect the performance of the drive and may cause damage to the motor windings or converter components.

Once a fault is detected, the DSP or controller determines the appropriate response. It can limit the current, stop the PWM command signals, or shut down the power converter completely. This fast

reaction helps isolate the fault, protect the internal components, and keep the electric drive within a safe operating range.

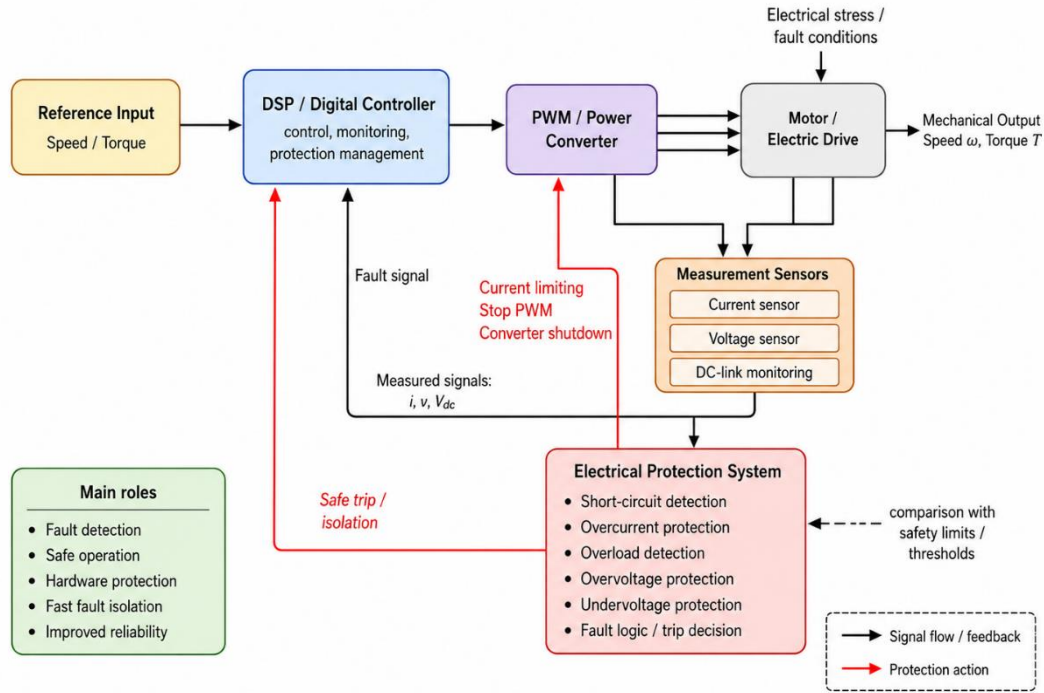


**Fig. II.33:** Electrical protection system in an intelligent electric drive.

### II.6.3 Fault Detection

Fault detection in an intelligent electric drive refers to the ability of the system to identify abnormal operating conditions and possible damage states. This function is very important in modern industrial drives, where reliability and safety are required during continuous operation.

As shown in Fig. II.34, the system uses real-time feedback from current, voltage, speed, and position sensors. By analyzing these measured signals, the controller can distinguish between normal variations during operation and real faults that may affect the motor or the power converter.

In practice, dedicated protection logic circuits are used to detect critical faults such as short circuits, overloads, high temperature, and undervoltage. These signals are compared with predefined safety limits or reference values. Then, the fault information is sent to the Digital Signal Processor (DSP), which evaluates the protection state and determines the type of abnormal condition.

Once a fault is detected, the drive can react according to the fault type. It may generate an alarm, send the fault information to a higher-level control system, limit the current, stop the PWM signals, or shut down the converter. These actions help isolate the fault and protect the internal components.

In this way, fault detection is not limited to simple measurement. It allows the intelligent drive to diagnose its operating condition and react quickly to protect the motor, the converter, and the control hardware.

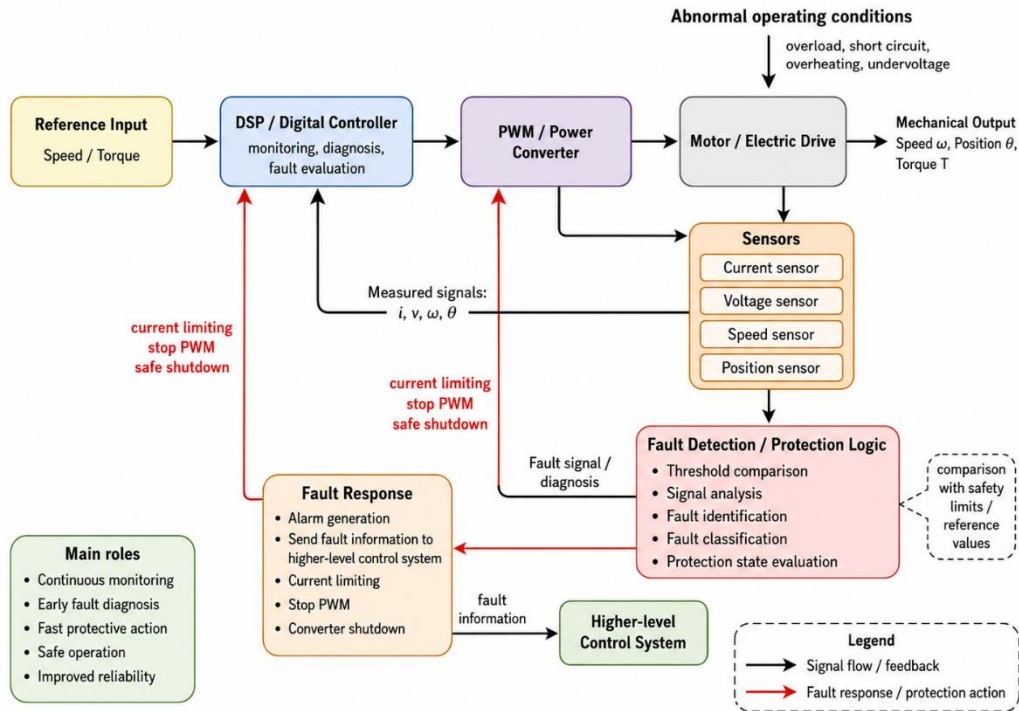


**Fig. II.34:** Fault detection and response process in an intelligent electric drive system.

## II.7 Conclusion

In this chapter, we studied the general architecture of an intelligent electric drive system. By dividing the system into its main stages, namely power conversion, control, measurement, and protection, it becomes easier to understand how electrical energy is converted into controlled mechanical motion.

Power converters, such as rectifiers, DC-DC choppers, and inverters, play an important role in this architecture. They regulate the electrical energy supplied to the motor, mainly through Pulse Width Modulation (PWM). This allows the drive to adapt its output according to the required speed, torque, and load conditions.

The intelligent behavior of the drive is mainly achieved through the integration of closed-loop control, sensing, and protection systems. The control stage allows the system to regulate speed and current, while intelligent control strategies, such as fuzzy logic, can improve the response under nonlinear conditions and load disturbances.

Sensors also represent an essential part of the system. Current, voltage, speed, and temperature sensors provide the real-time information required for feedback control, monitoring, and fault detection. In addition, electrical protection systems help protect the motor, converter, and control hardware against abnormal conditions such as overcurrent, overvoltage, overheating, and short circuits.

Finally, this chapter provides the basic framework for the next part of the thesis. Understanding the role of each component and the interaction between power, control, sensing, and protection stages is

necessary before moving to the detailed modeling, simulation, and analysis of the intelligent electric drive system.

## **CHAPTER III:**

# **MODELING AND SIMULATION OF MOTOR SPEED CONTROL SYSTEMS**

### III.1 Introduction:

Electric motor speed control is essential in industrial applications requiring accuracy, stability, and energy efficiency. DC motors offer simple speed regulation require more maintenance, while AC motors, especially three-phase induction motors, are widely used due to their robustness and reliability. Power electronic converters enable flexible voltage and frequency control for variable-speed operation. This chapter uses MATLAB/Simulink to model and simulate speed control systems for DC, single-phase AC, and three-phase induction motors, presenting their principles, models, control strategies, results, and comparison.

### III.2 Speed Control of a DC Motor:

#### III.2.1 Operating Principle of the DC Motor:

##### A. Construction and Basic Operation:

A DC motor is an electromechanical device that converts direct current electrical energy into mechanical rotational energy. Its operation is based on the interaction between an electric current and a magnetic field. When a current-carrying conductor is placed inside a magnetic field, it is subjected to a mechanical force. This force produces a torque that causes the rotor to rotate.

The basic electromagnetic force acting on a conductor is expressed as:  $F = B \cdot I \cdot l$

where  $F$  is the electromagnetic force [N],  $B$  is the magnetic flux density [T],  $I$  is the current flowing through the conductor [A], and  $l$  is the active length of the conductor inside the magnetic field [m].

##### B. Electromagnetic Torque and Back-EMF:

When the armature conductors carry current in the magnetic field produced by the stator, electromagnetic forces are created. The combined effect of these forces produces the electromagnetic torque of the motor. In the general case, the electromagnetic torque is proportional to the magnetic flux and the armature current:

$$C_{em} = K_T \Phi I_a$$

Since the field current is kept constant in a separately excited DC motor, the magnetic flux is also considered constant. Therefore, the torque equation can be simplified as:

$$C_{em} = K_T I_a$$

where  $C_{em}$  is the electromagnetic torque [N.m],  $K_T$  is the torque constant [N.m/A], and  $I_a$  is the armature current [A].

During rotation, the armature conductors cut the magnetic flux lines. According to Faraday's law, a voltage is induced in the armature winding. This induced voltage acts in the opposite direction to the applied armature voltage and is called the back electromotive force, or back-EMF. In the general case, the back-EMF is proportional to the magnetic flux and the angular speed of the motor:

$$e_b = K_e \Phi \Omega_r$$

Under constant magnetic flux, this relation becomes:

$$e_b = K_e \Omega_r$$

Under constant field flux, the back-EMF constant  $K_e$  and the torque constant  $K_T$  are numerically equal in SI units. Therefore, the following relation is used throughout this chapter:

$$K_e = K_T = K$$

### C. Speed Control Methods

The speed of a DC motor can be controlled by acting on the quantities that appear in its voltage equation. In steady-state operation, the derivative term of the armature current is neglected, and the motor speed can be expressed as:

$$\Omega_r = \frac{U_a - R_a I_a}{K_e}$$

This expression shows that the motor speed depends mainly on the armature voltage  $U_a$ , the armature voltage drop  $R_a I_a$ , and the back-EMF constant  $K_e$ .

In this work, armature voltage control is adopted. The armature voltage is adjusted using a DC-DC chopper controlled by a PWM signal. By changing the PWM duty cycle, the average voltage applied to the armature is modified, which allows the motor speed to be controlled.

#### III.2.2 DC Machine Parameter Identification

In order to build an accurate model of the DC machine, the electrical and mechanical parameters of the machine must be identified experimentally. This section presents the tests carried out to determine the armature and field resistances, the armature and field inductances, the mutual inductance, the electromotive force constant, the friction torque, and the moment of inertia of the system.

##### III.2.2.1 Armature and Field Resistance (DC Current Test)

The armature resistance  $R_a$  and the field resistance  $R_f$  are obtained by applying a DC voltage to each winding and measuring the corresponding current. The resistances are then calculated as follows:

$$R_a = \frac{U_a}{I_{a(n)}} \quad R_f = \frac{U_f}{I_{f(n)}}$$

where  $U_a$  and  $U_f$  are the DC voltages applied to the armature and field windings respectively, and  $I_{a(n)}$  and  $I_{f(n)}$  are the corresponding nominal currents.

##### III.2.2.2 Armature and Field Inductance (AC Current Test)

The inductances are determined by supplying each winding with an AC voltage and measuring the resulting impedance. Knowing the resistance obtained from the DC test, the inductance is deduced as follows:

$$L_a = \frac{1}{\omega} \cdot \sqrt{Z_a^2 - R_a^2} \quad L_f = \frac{1}{\omega} \cdot \sqrt{Z_f^2 - R_f^2}$$

where  $Z_a$  and  $Z_f$  are the armature and field impedances respectively, and  $\omega$  is the angular frequency of the supply.

### III.2.2.3 Mutual Inductance and No-Load Characteristic

The mutual inductance between the field winding and the armature winding is identified by coupling the DC machine (M) to a generator (G) mounted on the same shaft, with the field winding of the generator supplied independently and the induced voltage  $E$  measured at its armature terminals.

To trace the no-load characteristic, the following procedure is followed:

1. The rotational speed  $\Omega$  is set to a fixed value.
2. The field current  $I_f$  is increased progressively from zero to its nominal value.
3. For each value of  $I_f$ , the induced voltage  $E$  at the armature terminals is recorded.
4. The relation between  $E$  and  $I_f$  is expressed as:

$$E = f(I_f) = M \cdot \Omega \cdot I_f$$

The slope  $S$  of the linear part of this characteristic is equal to  $M \cdot \Omega$ , from which the mutual inductance is obtained as:

$$M = \frac{S}{\Omega}$$

In the linear region, the electromotive force can also be expressed as:

$$E = K\phi \cdot \Omega, \quad \text{with} \quad \phi = M \cdot I_f$$

From this relation, the mutual inductance can be deduced as:

$$M = \frac{E}{I_f \Omega}$$

### III.2.2.4 Electromotive Force Constant $K\phi$

The electromotive force constant  $K\phi$  is determined from the no-load test. Since:

$$E = K\phi \cdot \Omega$$

the constant  $K\phi$  is obtained as:

$$K\phi = \frac{E}{\Omega}$$

### III.2.2.5 Friction Torque and Mechanical Losses

The mechanical losses are represented by a friction torque expressed as:

$$C_f = f \cdot \Omega + C_s$$

where  $f$  is the viscous friction coefficient and  $C_s$  is the static (dry) friction torque.

From the no-load test, the electromagnetic torque can be written as:

$$C_{em} = K\phi \cdot I_0 = f \cdot \Omega + C_s$$

The slope of the curve  $C_{em}(\Omega)$  corresponds to the viscous friction coefficient  $f$ , while the intersection of this curve with the vertical axis corresponds to the static friction torque  $C_s$ . The viscous friction coefficient is therefore obtained as:

$$f = \frac{K\phi I_0 - C_s}{\Omega} = \frac{\Delta C_{em}}{\Delta \Omega}$$

### III.2.2.6 Determination of the Moment of Inertia J

The moment of inertia  $J$  of the motor-load assembly is determined experimentally by means of a deceleration (run-down) test, carried out according to the following procedure:

1. The motor is driven up to its nominal speed  $\Omega_N$ .
2. At the instant  $t = 0$  s, the armature supply is interrupted, so that  $I_a = 0$ .
3. The speed curve is recorded during the natural deceleration of the machine.

During this free deceleration phase, the electromagnetic torque is zero ( $C_{em} = 0$ ), and the fundamental equation of motion reduces to:

$$J \frac{d\Omega}{dt} = C_{em} - f\Omega - C_s = -(f\Omega + C_s)$$

Since the mechanical losses (friction torque) are responsible for the deceleration, the moment of inertia can be estimated from the deceleration time and the corresponding loss torque as:

$$J = \frac{C_{pertes} t_{stop}}{\Omega_N} = \frac{K\phi I_0 t_{stop}}{\Omega_N}$$

where  $C_{pertes}$  is the total loss torque,  $t_{stop}$  is the time required for the machine to come to a complete stop, and  $\Omega_N$  is the nominal angular speed from which the deceleration test is started.

### III.2.3 Mathematical Model of the DC Motor

The mathematical model of a separately excited DC motor is constructed from two coupled sets of equations: the electrical equation of the armature circuit and the mechanical equation of the rotor shaft. These two equations are linked through the back-EMF and the electromagnetic torque, forming an electromechanical system whose dynamic behaviour can be fully described by four fundamental equations [7].

### A. Electrical Equation of the Armature Circuit

Applying Kirchhoff's voltage law to the armature circuit, the instantaneous voltage balance is written as [1], [7]:

$$V_a(t) = R_a i_a(t) + \frac{L_a (di_a(t))}{dt} + e_b(t) \quad (\text{III. 1})$$

Where:

- $V_a(t)$  : armature supply voltage (V)
- $R_a$  : armature winding resistance ( $\Omega$ )
- $i_a(t)$  : armature current (A)
- $L_a$  : armature winding inductance (H)
- $e_b(t)$  : back-electromotive force (V)

The armature equivalent circuit is shown in Fig. III. 1. It consists of the armature resistance  $R_a$ , the armature inductance  $L_a$ , and the back-EMF source  $e_b$ , which opposes the applied armature voltage.

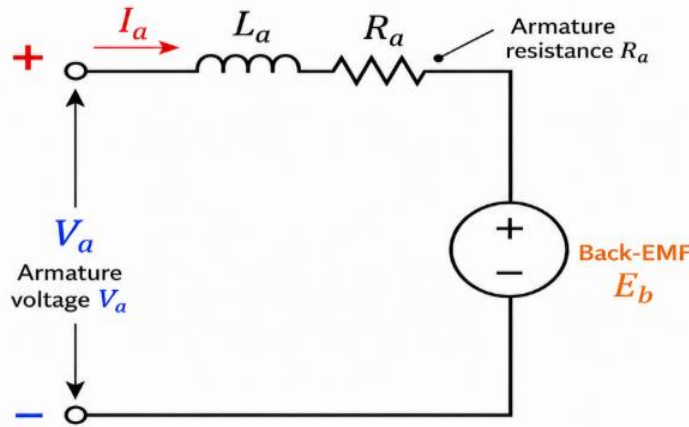


**Fig. III.1:** Equivalent armature circuit of a separately excited DC motor.

### B. Back-EMF and Electromagnetic Torque:

In a separately excited DC motor, the field current is supplied independently and maintained constant during normal operation. Under this condition, the magnetic flux  $\Phi$  is constant, and the motor equations simplify considerably. The back-EMF induced in the rotating armature is proportional to the rotor angular speed [1], [7]:

$$e_b(t) = K_e \Omega_r(t) \quad (\text{III. 2})$$

Where:

- $K_e$  : back-EMF constant (V·s/rad)
- $\Omega_r(t)$  : rotor angular speed (rad/s)

The electromagnetic torque produced by the interaction between the armature current and the magnetic flux is given by [1], [7]:

$$C_{em}(t) = K_t i_a(t) \quad (\text{III. 3})$$

Where:

- $C_{em}(t)$  : electromagnetic torque (N·m)
- $K_t$  : motor torque constant (N·m/A)

It should be noted that under constant magnetic flux, the two motor constants  $K_e$  and  $K_t$  are numerically equal in the SI system ( $K_e = K_t = K$ ). This equality is a well-established result in DC machine theory and holds throughout this chapter.

### C. Mechanical Equation:

Applying Newton's second law of rotation to the motor shaft, the torque balance between the electromagnetic torque, the load torque, the inertial torque, and the friction torque gives [7]:

$$J \frac{d\Omega_r(t)}{dt} + f\Omega_r(t) = C_{em}(t) - C_r(t) \quad (III. 4)$$

Where:

- $J$  : total moment of inertia of the rotor and load (kg·m<sup>2</sup>)
- $f$ : viscous friction coefficient (N·m·s/rad)
- $C_r(t)$ : load torque (N·m)

Equation (III.4) assumes that only viscous (linear) friction acts on the shaft. This is the standard assumption used in DC motor control modeling.

### D. Transfer Functions in the Laplace Domain:

To enable the design of the speed controller and the construction of the Simulink block diagram, equations (III.1) and (III.4) are transformed into the Laplace domain, assuming zero initial conditions.

#### ➤ Electrical subsystem:

Taking the Laplace transform of equation (III.1) and rearranging:

$$I_a(s) = \frac{V_a(s) - e_b(s)}{R_a + L_a s} \quad (III. 5)$$

Equation (III.5) describes a first-order system. Defining the armature electrical time constant  $\tau_a$  as:

$$\tau_a = \frac{L_a}{R_a} \quad (III. 6)$$

Equation (III.5) is rewritten in standard first-order transfer function form:

$$\frac{I_a(s)}{V_a(s) - e_b(s)} = \frac{\left(\frac{1}{R_a}\right)}{1 + \tau_a s} \quad (III. 7)$$

#### ➤ Mechanical subsystem:

Taking the Laplace transform of equation (III.4) and rearranging:

$$\Omega_r(s) = \frac{C_{em}(s) - C_r(s)}{Js + f} \quad (III. 8)$$

Defining the mechanical time constant  $\tau_{mec}$  as:

$$\tau_{mec} = \frac{J}{f} \quad (III. 9)$$

Equation (III.8) is rewritten in standard first-order form:

$$\frac{\Omega(s)}{C_{em}(s) - C_r(s)} = \frac{\left(\frac{1}{f}\right)}{1 + \tau_{mec} s} \quad (III. 10)$$

Equations (III.5) through (III.10) form the complete transfer function model of the separately excited DC motor. The armature and mechanical subsystems are each represented as first-order systems, characterised respectively by the electrical time constant  $\tau_a$  and the mechanical time constant  $\tau_{mec}$ . This model serves as the direct basis for the closed-loop control design in Section III.2.3 and for the Simulink implementation in Section III.2.4.

### III.2.4 Closed-Loop Speed Control Strategy:

The mathematical model developed in Section III.2.2 shows that the speed of the DC motor is mainly affected by the armature voltage and the load torque. In open-loop operation, any variation in the load torque may produce a speed deviation, and the system cannot automatically correct this error. For this reason, a closed-loop speed control strategy is adopted in order to improve speed regulation and disturbance rejection.

In the proposed closed-loop system, the measured motor speed  $\Omega_r(t)$  is continuously compared with the reference speed  $\Omega_{ref}(t)$ . The resulting error is processed by a Proportional-Integral (PI) controller. The output of the controller is then used to adjust the duty cycle of the PWM signal, which controls the DC-DC chopper supplying the motor armature. In this way, the armature voltage is automatically regulated according to the speed error.

#### A. Speed Error and PI Controller

At each instant, the speed error signal is computed as the difference between the reference speed and the measured motor speed:

$$e(t) = \Omega_{ref}(t) - \Omega_r(t) \quad (III. 11)$$

Where:

- $e(t)$  : speed error (rad/s)
- $\Omega_{ref}(t)$  : reference angular speed (rad/s)
- $\Omega_r(t)$  : measured motor angular speed (rad/s)

This error is fed to a Proportional-Integral (PI) controller. The PI controller was selected because it provides zero steady-state error for step reference inputs without the noise amplification problems associated with derivative action. The control output in the time domain is expressed as [7]:

$$u(t) = K_p e(t) + K_i \int_0^t e(\tau) d\tau \quad (III. 12)$$

Where:

- $u(t)$  : PI controller output (dimensionless, duty cycle)

- $K_p$  : proportional gain
- $K_i$  : integral gain

Taking the Laplace transform of equation (III.12), the transfer function of the PI controller is written as:

$$C(s) = K_p + \frac{K_i}{s} \quad (\text{III. 13})$$

or equivalently:

$$C(s) = \frac{K_p s + K_i}{s} \quad (\text{III. 14})$$

Equation (III.13) shows that the PI controller introduces one integrator pole at the origin, which ensures zero steady-state error in response to a step speed reference, and one zero at  $s = -K_i/K_p$ , which contributes to the damping of the closed-loop transient response.

### B. Duty Cycle Limitation and PWM Generation

The output  $u(t)$  of the PI controller directly represents the duty cycle  $\alpha$  of the PWM signal. Since the duty cycle must remain physically meaningful, its value is bounded by a saturation block:

$$0 \leq \alpha \leq 1 \quad (\text{III. 15})$$

When  $u(t)$  exceeds 1, the output is clamped to 1 (maximum armature voltage). When it falls below 0, the output is clamped to 0 (zero armature voltage). This saturation block ensures that the duty cycle remains within its physically valid range  $[0, 1]$ .

The bounded duty cycle  $\alpha$  is then used by the PWM generator to produce a high-frequency switching signal. This PWM signal is applied to the gate of an IGBT transistor, which acts as the controlled switch of the DC-DC chopper [17].

### C. DC-DC Chopper and Armature Voltage Control

The DC-DC chopper is a step-down (Buck) converter that converts the fixed DC supply voltage  $V_{dc}$  into a variable armature voltage  $V_a$ , controlled by the duty cycle  $\alpha$ . In continuous conduction mode, the average armature voltage delivered to the motor is [17]:

$$V_a = \alpha \cdot V_{dc} \quad (\text{III. 16})$$

Where:

- $V_a$ : average armature voltage applied to the motor (V)
- $\alpha$  : duty cycle of the PWM signal (dimensionless,  $0 \leq \alpha \leq 1$ )
- $V_{dc}$  : fixed DC supply voltage (V)

Equation (III.16) shows that increasing the duty cycle ( $\alpha$ ) increases the average armature voltage, which raises the motor speed. In the same way, decreasing ( $\alpha$ ) reduces the armature voltage and

lowers the speed. Therefore, the DC-DC chopper behaves as a variable DC voltage source controlled by the PI controller output.

Under steady-state conditions, the derivative term of the armature current is neglected. By combining the armature voltage equation with the chopper output voltage, the motor speed can be related to the duty cycle as follows:

$$\Omega_r = \frac{\alpha V_{dc} - R_a I_a}{K_e} \quad (\text{III. 17})$$

This equation shows that the motor speed can be controlled through the duty cycle ( $\alpha$ ). Thus, the PI controller indirectly controls the motor speed by adjusting the average armature voltage supplied by the chopper.

#### D. Complete Closed-Loop Control System

Figure III.2 presents the block diagram of the closed-loop PI speed control system for the separately excited DC motor. The system is composed of a reference speed input, a comparator, a PI controller, a saturation block, a PWM generator, a DC-DC chopper, the DC motor, and a speed sensor used for feedback.

The reference speed  $\Omega_{ref}(t)$  is first compared with the measured motor speed  $\Omega_r(t)$ . The difference between these two signals represents the speed error  $e(t)$ . This error is applied to the PI controller, which generates the control signal required to reduce the difference between the reference speed and the actual speed. Before being sent to the PWM generator, the controller output is limited by the saturation block in order to keep the duty cycle within its physical range.

The PWM generator converts the controller output into a duty cycle  $\alpha$ , which drives the power switch of the DC-DC chopper. The chopper then adjusts the average armature voltage  $U_a$  applied to the motor according to the relation shown in the figure. By increasing or decreasing the duty cycle, the armature voltage changes, and consequently the motor speed is regulated.

The motor produces an electromagnetic torque and rotates at the speed  $\Omega_r(t)$ . This speed is measured by the speed sensor and fed back to the comparator, which closes the control loop. The load torque  $C_r$  acts as a disturbance on the mechanical part of the motor. When the load torque changes, the controller automatically modifies the duty cycle and the armature voltage to reduce the speed error and maintain the motor speed close to its reference value.

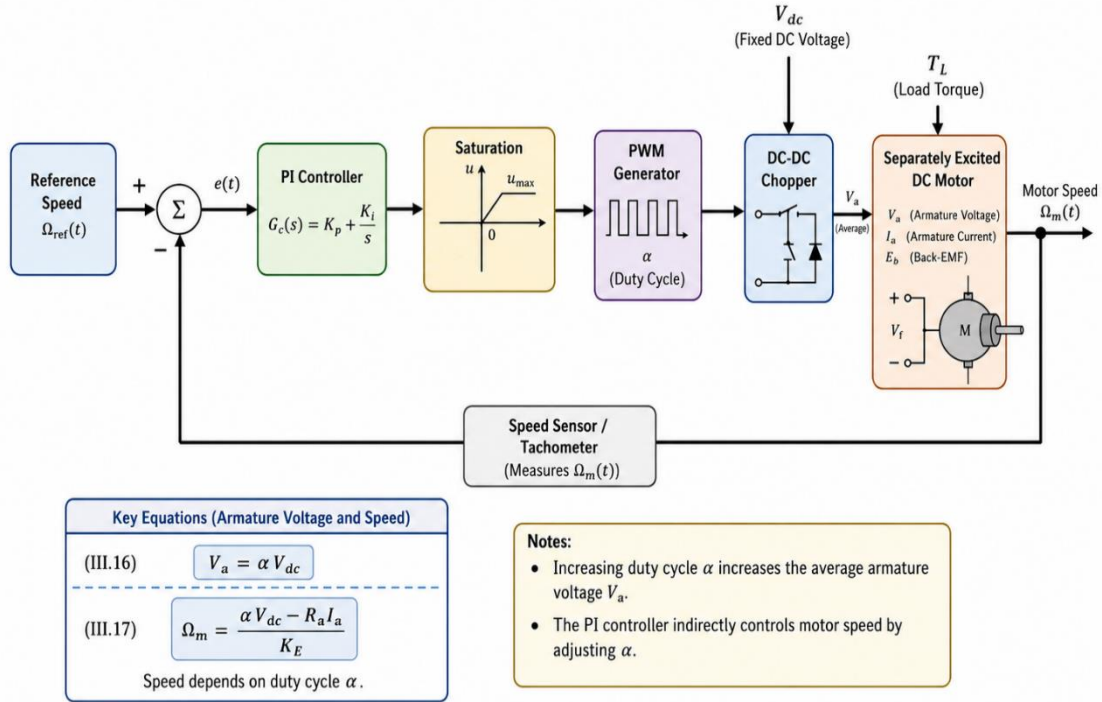


**Fig. III.2:** Block diagram of the closed-loop PI speed control system for the separately excited DC motor.

### E. Calculation of PI Controller Gains:

To calculate the PI controller gains, the mechanical model previously developed in Section III.2.2 is used. Starting from the mechanical torque balance equation given in (III.4):

$$J \frac{d\Omega_r(t)}{dt} + f\Omega_r(t) = C_{em}(t) - C_r(t)$$

For the nominal controller design, the load torque  $C_r(t)$  is considered as an external disturbance and is neglected during the gain calculation. Therefore, equation (III.4) becomes:

$$J \frac{d\Omega_r(t)}{dt} + f\Omega_r(t) = C_{em}(t)$$

Using the electromagnetic torque relation given in (III.3):

$$C_{em}(t) = K_T I_a(t)$$

and applying the Laplace transform under zero initial conditions, the mechanical equation can be written as:

$$(Js + f)\Omega_r(s) = K_T I_a(s)$$

Thus, the simplified mechanical transfer function used for the PI controller design is obtained as:

$$G_m(s) = \frac{\Omega_r(s)}{I_a(s)} = \frac{K_T}{Js + f} \quad (III. 18)$$

where  $K_T$  is the torque constant,  $J$  is the total moment of inertia, and  $f$  is the viscous friction coefficient.

The mechanical time constant of the motor is calculated as:

$$\tau_{mec} = \frac{J}{f} \quad (\text{III. 19})$$

Using the motor parameters:

$$J = 0.05 \text{ kg} \cdot \text{m}^2$$

$$f = 0.01 \text{ N} \cdot \text{m} \cdot \text{s/rad}$$

the mechanical time constant becomes:

$$\tau_{mec} = \frac{0.05}{0.01} = 5 \text{ s}$$

Starting from the PI controller transfer function previously obtained in equation (III.13):

$$G_c(s) = K_p + \frac{K_i}{s}$$

this expression can be rewritten by factoring out the proportional gain ( $K_p$ ):

$$G_c(s) = K_p \left( 1 + \frac{K_i}{K_p s} \right)$$

By introducing the integral time constant ( $T_n$ ), defined as:

$$T_n = \frac{K_p}{K_i}$$

we obtain:

$$\frac{K_i}{K_p} = \frac{1}{T_n}$$

Therefore, the PI controller can be written in the following standard form:

$$C(s) = K_p \left( 1 + \frac{1}{T_n s} \right) \quad (\text{III. 20})$$

In order to cancel the mechanical pole of the motor, the integral time constant of the PI controller is chosen equal to the mechanical time constant:

$$T_n = \tau_{mec} = \frac{J}{f} = 5 \text{ s}$$

The armature electrical time constant is calculated from the armature inductance and resistance:

$$\tau_a = \frac{L_a}{R_a}$$

Using:  $L_a = 0.016 \text{ H}$ ,  $R_a = 0.78 \Omega$

$$\tau_a = \frac{0.016}{0.78} = 0.0205 \text{ s}$$

For the controller design, the desired closed-loop time constant is chosen to be approximately ten times the armature electrical time constant:  $\tau_{CL} \approx 10\tau_a$

Therefore:  $\tau_{CL} \approx 10 \times 0.0205 = 0.205 \text{ s} \approx 0.2 \text{ s}$

Using the simplified mechanical transfer function obtained in (III.18):

$$G_m(s) = \frac{K_T}{Js + f}$$

and the PI controller written in the standard form given in (III.20):

$$C(s) = K_p \left( 1 + \frac{1}{T_n s} \right)$$

the open-loop transfer function is obtained by multiplying the controller by the mechanical model:

$$L(s) = C(s)G_m(s)$$

Therefore:

$$L(s) = K_p \left( 1 + \frac{1}{T_n s} \right) \frac{K_T}{Js + B_m}$$

Since the integral time constant is selected equal to the mechanical time constant:

$$T_n = \tau_{mec} = \frac{J}{B_m}$$

the zero introduced by the PI controller cancels the mechanical pole of the motor. After this simplification, the open-loop transfer function becomes:

$$L(s) = \frac{K_p K_T}{Js}$$

The corresponding closed-loop transfer function can be written as:

$$\frac{\Omega_r(s)}{\Omega_{ref}(s)} = \frac{L(s)}{1 + L(s)} = \frac{\frac{K_p K_T}{Js}}{1 + \frac{K_p K_T}{Js}} = \frac{\frac{K_p K_T}{J}}{s + \frac{K_p K_T}{J}}$$

This expression represents a first-order system. Its closed-loop time constant is therefore:

$$\tau_{CL} = \frac{J}{K_p K_T}$$

By rearranging this equation, the proportional gain is obtained as:

$$K_p = \frac{J}{K_T \tau_{CL}} \quad (III. 21)$$

The torque constant ( $K_T$ ) is obtained from the field-armature mutual inductance and the field current. Since the field voltage is ( $U_f = 150, V$ ) and the field resistance is ( $R_f = 150\Omega$ ), the field current is:

$$I_f = \frac{U_f}{R_f} = \frac{150}{150} = 1A$$

Therefore:

$$K_T = L_{af}I_f = 1.234 \times 1 = 1.234 \text{ N.m/A}$$

This relation follows from DC machine theory, where the torque constant  $K_T$  is determined by the field-armature mutual inductance and the field current. In SI units,  $K_T$  in N.m/A is numerically equal to  $L_{af}I_f$ .

Using  $J = 0.05 \text{ kg.m}^2$ ,  $K_T = 1.234 \text{ N.m/A}$  and  $\tau_{CL} = 0.2 \text{ s}$

$$K_p = \frac{0.05}{1.234 \times 0.2} = 0.2026$$

The integral gain is obtained from:

$$K_i = \frac{K_p}{T_n} \quad (\text{III. 22})$$

or equivalently:

$$K_i = \frac{f}{K_T \tau_{CL}} \quad (\text{III. 23})$$

Thus:

$$K_i = \frac{0.01}{1.234 \times 0.2} = 0.0405$$

Therefore, the final PI controller gains used in the Simulink model are:

$$K_p = 0.2026$$

$$K_i = 0.0405$$

In the Simulink PI Controller block, the controller type is selected as PI, with the derivative gain set to zero. Therefore, the entered parameters are:  $P = 0.2026$ ,  $I = 0.0405$ ,  $D = 0$

Since the PI controller output represents the PWM duty cycle  $\alpha$ , it must be limited between 0 and 1:  $0 \leq \alpha \leq 1$

This saturation ensures that the duty cycle remains physically valid for the PWM signal. In addition, an anti-windup mechanism is used in the Simulink controller to prevent integrator saturation during large speed reference changes.

### III.2.5 Simulation Model in MATLAB/Simulink:

To validate the closed-loop speed control strategy developed in the previous section, a simulation model was built in MATLAB/Simulink. The model represents a separately excited DC motor supplied by a DC-DC chopper and controlled using a PI speed controller. The complete Simulink model is shown in Fig. III.3.

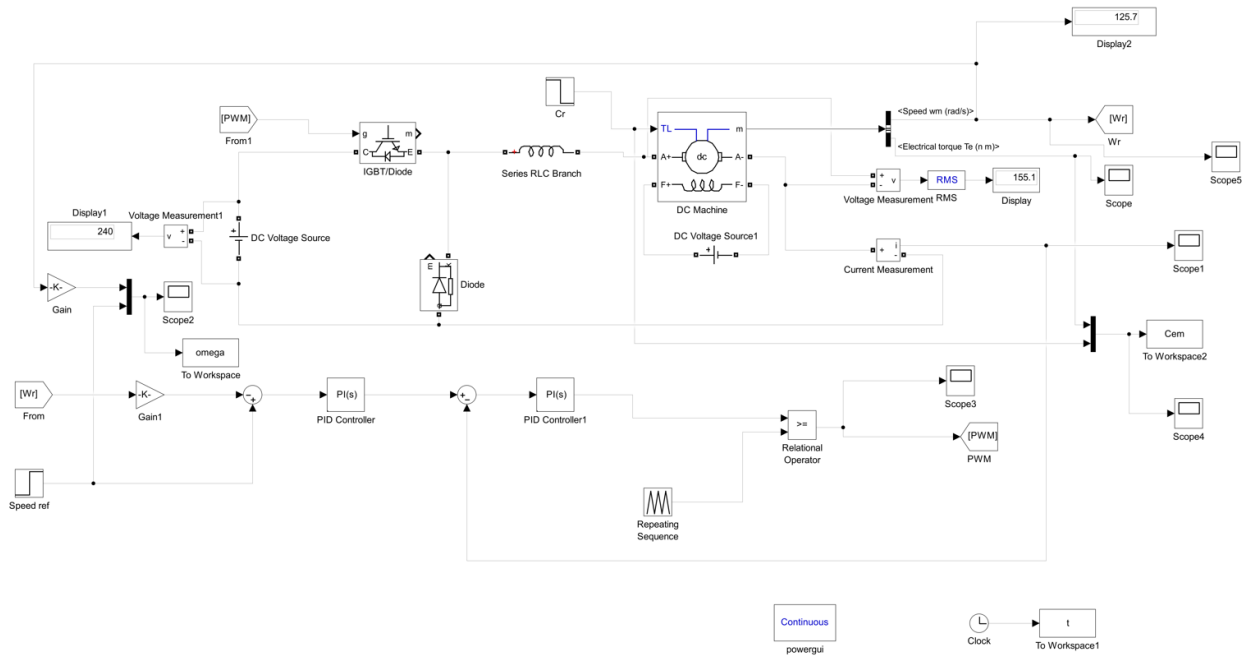


**Fig. III.3:** MATLAB/Simulink model of the closed-loop DC motor speed control system.

The Simulink model is composed of two main parts: the control circuit and the power circuit. The control system is organized in a cascade structure consisting of an outer speed control loop and an inner current control loop.

The reference speed is generated using a Step block named Speed Ref. In this simulation, the reference speed is initially set to 800 rpm and changes to 1200 rpm at ( $t = 5, s$ ). The measured motor speed is fed back and compared with the reference speed through a summing junction to generate the speed error.

The speed error is processed by the first PI controller (speed controller), which generates the reference armature current. This current reference is then compared with the measured armature current in a second summing junction. The resulting current error is applied to a second PI controller (current controller).

The output of the current controller determines the duty-cycle command of the chopper. This signal is compared with a triangular carrier waveform generated by a Repeating Sequence block through a Relational Operator block. The comparison produces the PWM pulses used to drive the gate of the IGBT switch in the power converter.

The power circuit consists of a DC voltage source, an IGBT switch with a freewheeling diode, and a separately excited DC motor. The PWM-controlled chopper regulates the average armature voltage supplied to the motor, thereby controlling its speed according to the reference command.

### III.2.6 Simulation Results and Discussion:

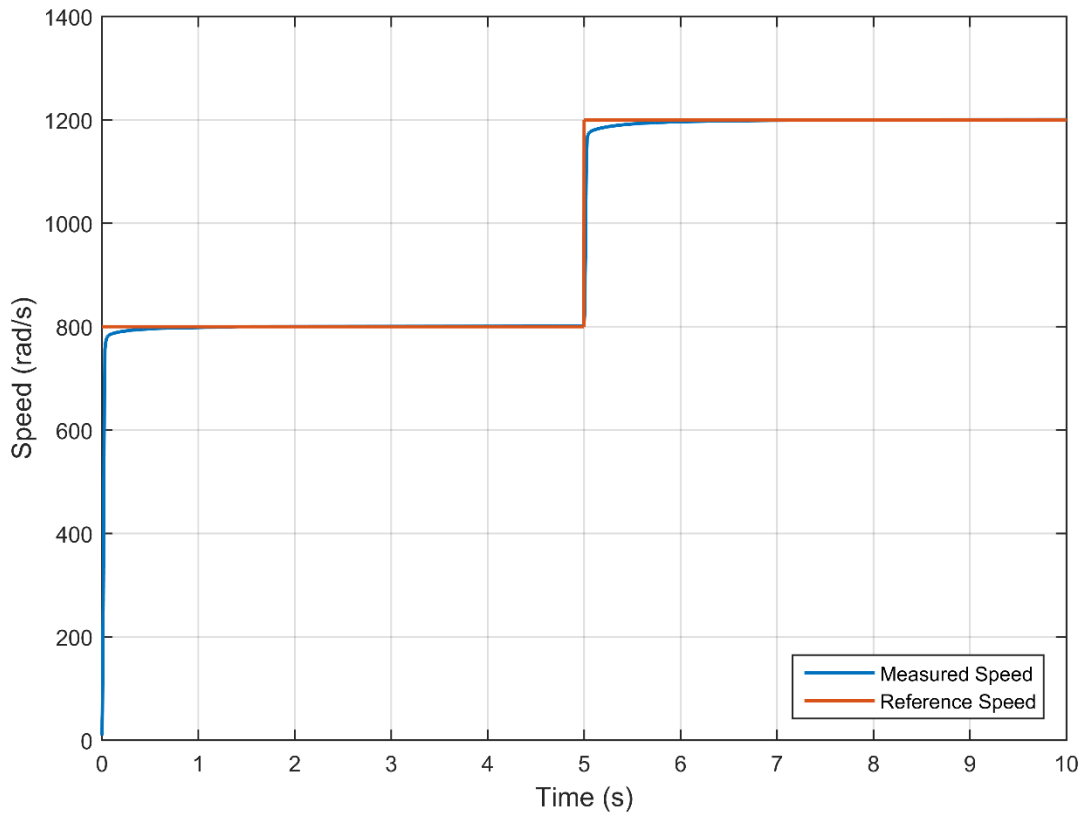


**Fig. III.4:** Motor speed response for a reference speed change from 800 rpm to 1200 rpm.

Fig. III.4 illustrates the speed response of the separately excited DC motor obtained through MATLAB/Simulink simulation. The reference speed is initially set to 800 rpm. Starting from rest, the motor accelerates rapidly and reaches the desired speed with a very small transient deviation. After the transient period, the measured speed closely follows the reference and remains stable around the commanded value.

At  $t = 5$  s, the reference speed is increased from **800 rpm to 1200 rpm**. The motor responds quickly to this change and accelerates smoothly toward the new operating point. The measured speed tracks the reference accurately and settles at the desired value within a short time, with negligible overshoot and no noticeable steady-state error.

The close agreement between the measured speed and the reference speed demonstrates the effectiveness of the PI speed controller. By adjusting the PWM duty cycle of the DC–DC chopper, the controller regulates the armature voltage and ensures accurate speed tracking over the entire operating range. These results confirm the good dynamic performance and stability of the closed-loop speed control system.

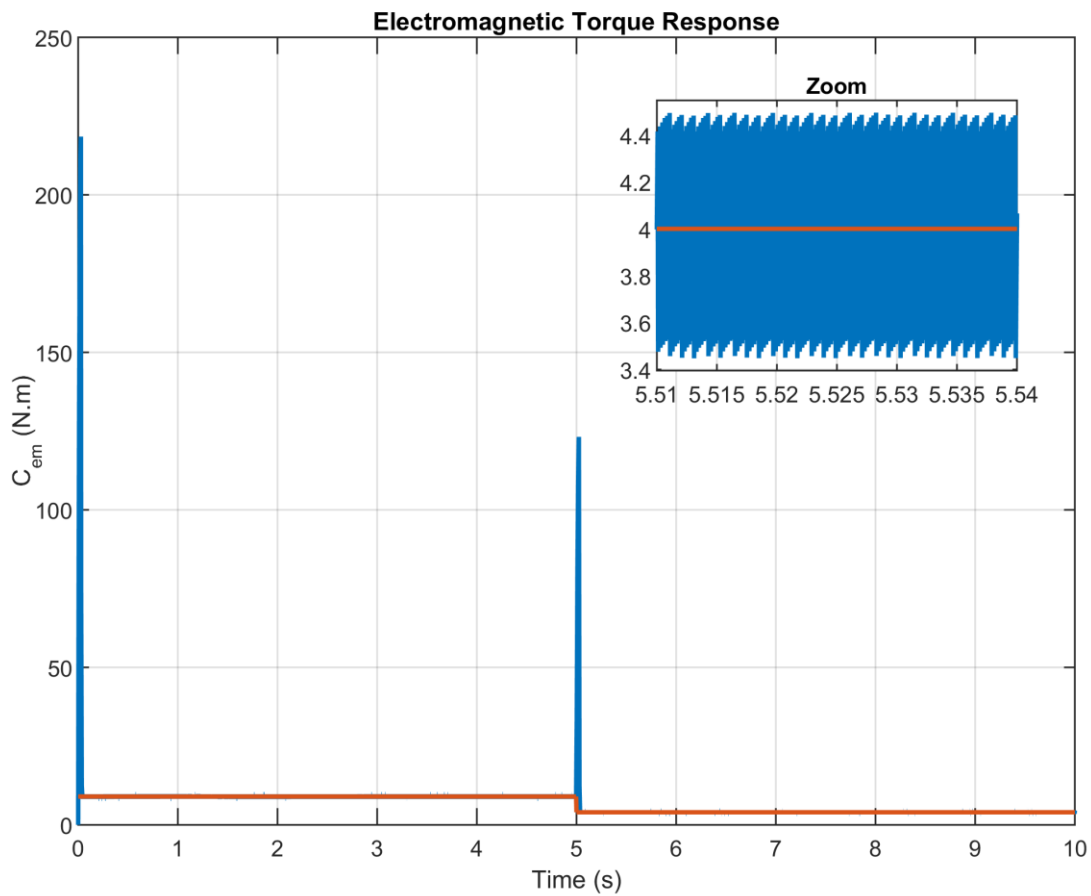


**Fig. III.5:** Electromagnetic torque response under a load torque change from 9 N·m to 4 N·m.

The electromagnetic torque response of the separately excited DC motor is shown in Fig. III.5. Initially, the load torque is set to 9 N·m, and the motor develops the required electromagnetic torque to balance the applied load and maintain the reference speed. At  $t = 5$  s, the load torque is reduced from 9 N·m to 4 N·m. As a result, the electromagnetic torque decreases rapidly and reaches a new steady-state value corresponding to the reduced load condition.

The transient response is short and well damped, indicating that the control system effectively compensates for load variations. The zoomed section highlights the small torque ripple produced by the PWM switching process. Despite these high-frequency oscillations, the average electromagnetic torque remains close to the required value, ensuring stable motor operation.

These results demonstrate the ability of the PI-controlled DC drive to maintain satisfactory dynamic performance and quickly adapt to sudden changes in load torque while preserving speed regulation and system stability.

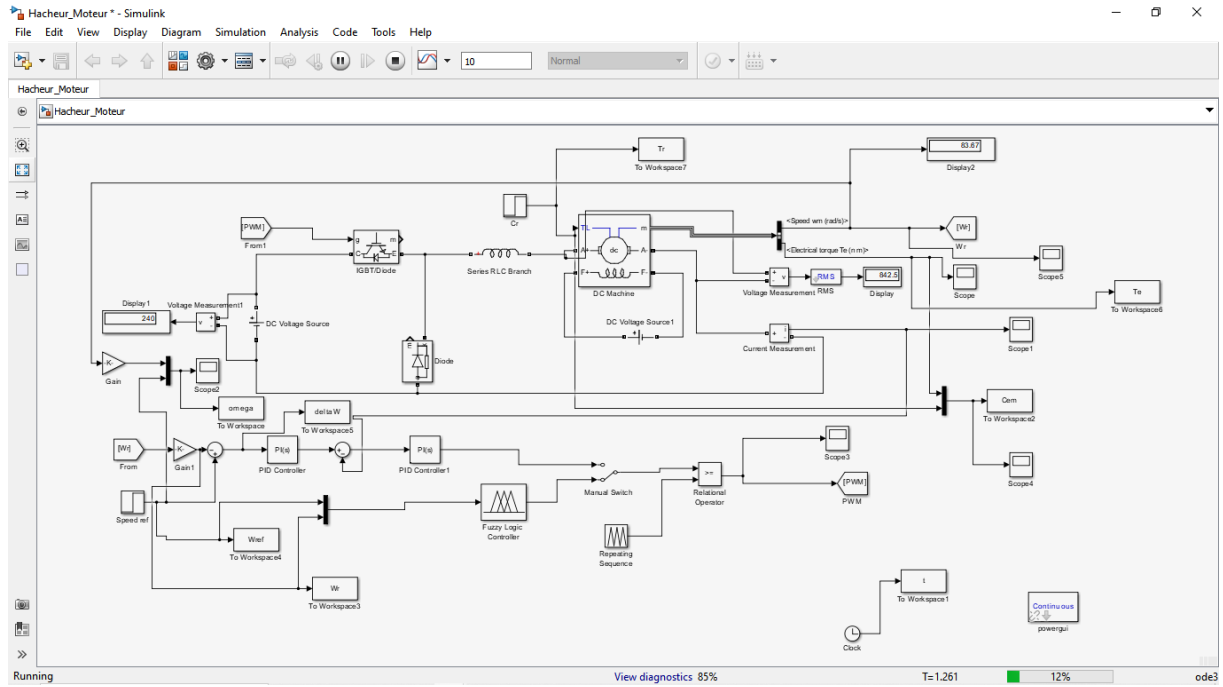


**Fig. III.6:** MATLAB/Simulink model of the closed-loop DC motor speed control system ANFIS/PI.

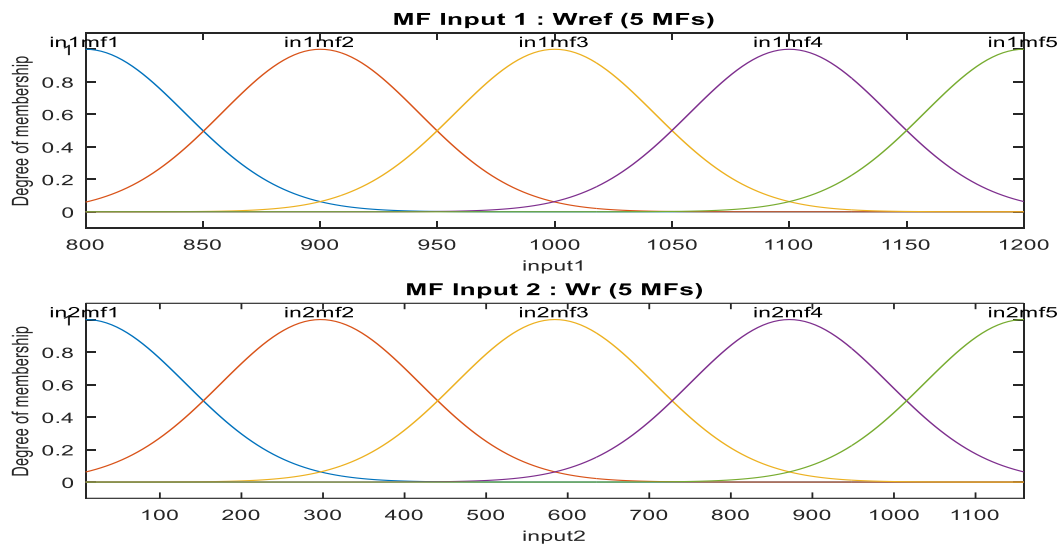


**Fig. III.7:** (ANFIS input membership functions)

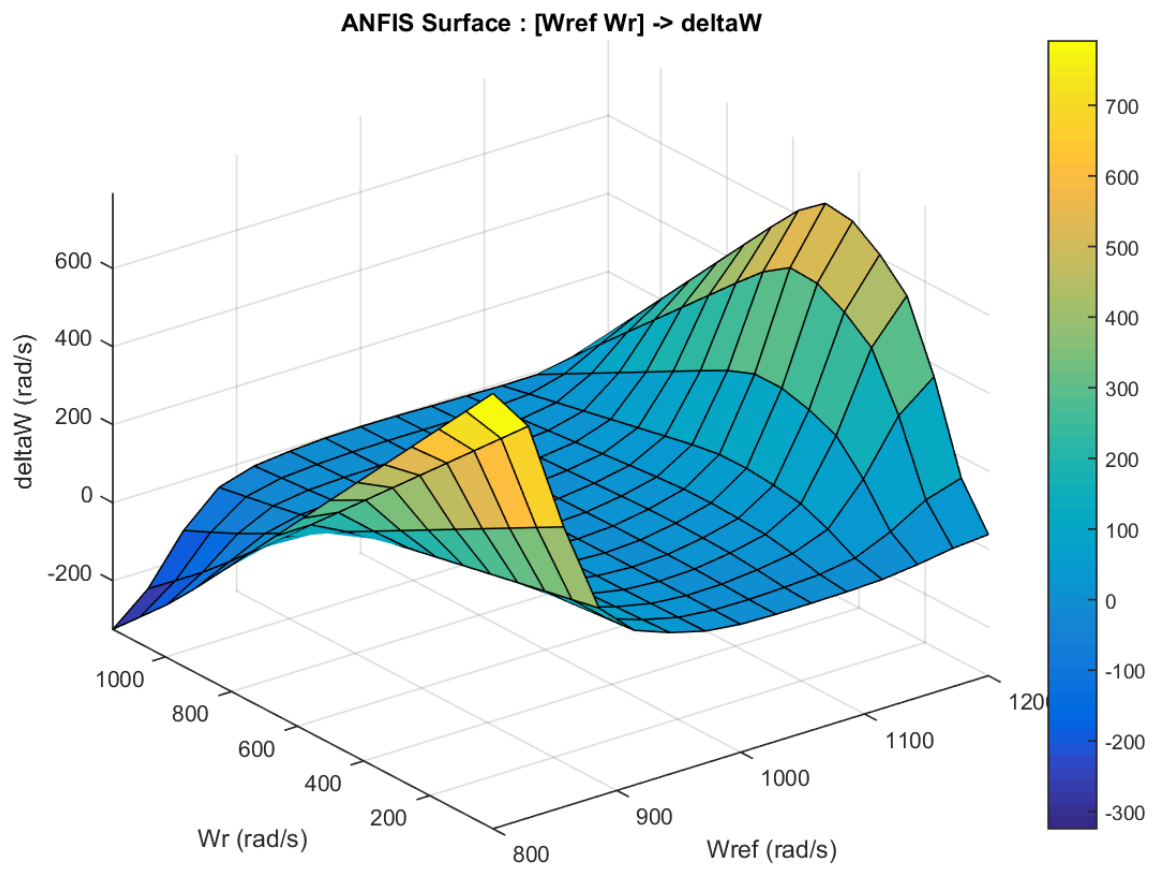


**Fig. III.8:** (ANFIS Surface)

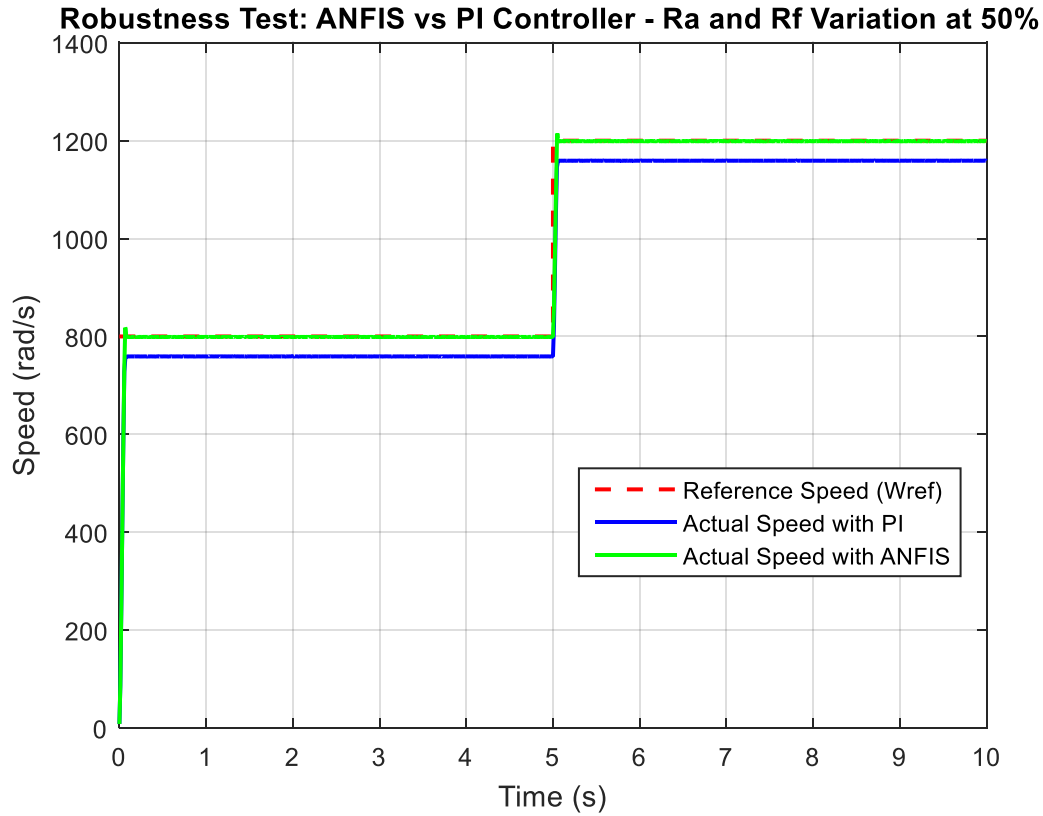


**Fig. III.9:** Speed responses – Perturbed conditions ( $R_a=1.17$  ohms,  $R_f=225$  ohms)

Fig. III.5.c illustrates the speed responses of the PI controller and the ANFIS-based controller under electrical parameter variations. The reference speed changes from 800 rad/s to 1200 rad/s at  $t=5$  s. It can be observed that the ANFIS controller tracks the reference speed almost perfectly, maintaining a negligible steady-state error and achieving a final speed very close to the desired value. In contrast, the conventional PI controller exhibits a noticeable steady-state error, resulting in a lower actual speed than the reference.

These results demonstrate that the ANFIS controller is more robust to electrical parameter variations, such as changes in stator and rotor resistances. The adaptive learning capability of ANFIS enables it to compensate for parameter uncertainties and maintain the desired dynamic performance. On the other hand, the fixed-gain PI controller is more sensitive to parameter changes, leading to degraded tracking accuracy and reduced performance.

Therefore, the ANFIS-based control strategy provides superior robustness, better reference tracking, and improved adaptability compared to the conventional PI controller when the motor electrical parameters vary.

### III.3 Speed Control of a Single-Phase AC Motor:

#### III.3.1 Operating Principle of the Single-Phase AC Motor:

A single-phase AC motor is widely used in low-power applications because it can operate from a simple single-phase supply. It is commonly found in domestic appliances, small pumps, fans, compressors, and other systems where a three-phase supply is not available. In this type of motor, the stator is supplied by a single-phase alternating voltage, which produces an alternating magnetic field in the air gap.

Unlike a three-phase induction motor, the magnetic field produced by a single-phase winding is not naturally rotating at starting. It is mainly a pulsating magnetic field. This means that the field changes its magnitude and direction with time, but it does not create a strong starting torque by itself. For this reason, a single-phase induction motor is generally not self-starting. When the rotor is initially at rest, the forward and backward magnetic field effects are balanced, and the resulting starting torque is approximately zero.

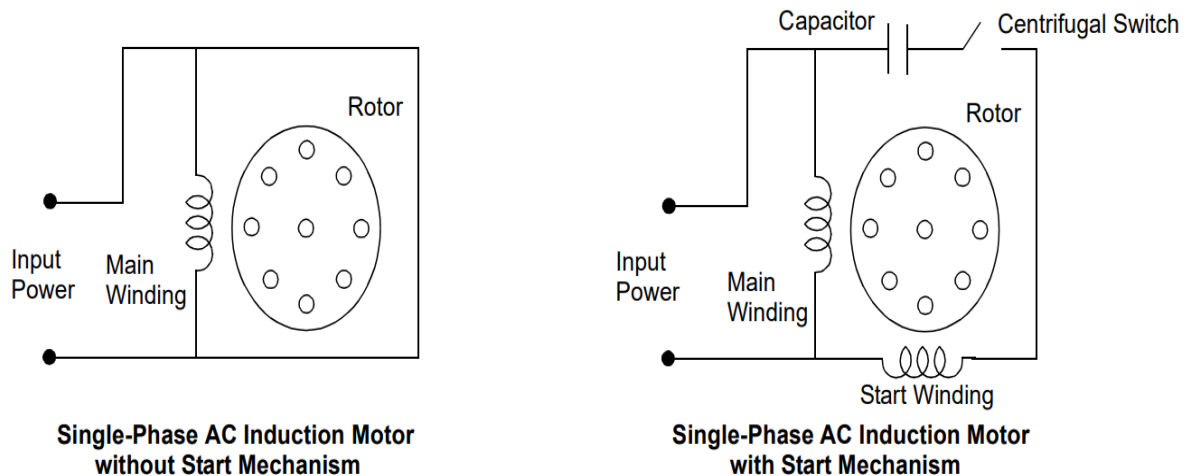


**Fig. III.10:** Single-phase AC induction motor without and with starting mechanism[8].

As shown in Fig. III.6, the motor without a starting mechanism contains only the main winding supplied by the single-phase input power. In this condition, the motor cannot produce a sufficient starting torque by itself. The second diagram shows the addition of a start winding, a capacitor, and a centrifugal switch. This auxiliary circuit helps create a phase shift between the main winding current and the start winding current, which produces a rotating magnetic field component at starting.

To start the motor, an auxiliary starting method is required. This can be done by adding an auxiliary winding with a capacitor, a shaded pole arrangement, or another starting circuit. The purpose of the starting method is to create a phase shift between two stator currents. This phase shift produces a rotating magnetic field component, which gives the rotor an initial starting torque. Once the motor starts rotating, the interaction between the stator magnetic field and the induced rotor currents produces electromagnetic torque, allowing the motor to continue running.

During operation, the rotor speed remains slightly lower than the synchronous speed. This difference is necessary because the rotor must cut the magnetic field in order to induce current and produce torque. Therefore, the motor operates with a certain slip, which depends on the load and the

supply conditions. When the mechanical load increases, the rotor speed decreases slightly, and the slip increases to produce the required torque.

In this work, the single-phase AC motor is considered from the point of view of speed control. The main control method studied is based on varying the RMS value of the supply voltage while keeping the supply frequency constant. This method affects the developed torque and the final operating speed of the motor, especially for fan-type or low-power loads.

### III.3.2 Basic Speed Equation and Slip:

The speed of the rotating magnetic field produced in the stator of an induction motor is called the synchronous speed. It depends only on the supply frequency and the number of magnetic poles of the motor. The synchronous speed is given by:

$$N_s = \frac{60 \cdot f}{P} \quad (\text{III. 24})$$

where  $N_s$  is the synchronous speed [rpm],  $f$  is the supply frequency [Hz], and  $P$  is the number of magnetic pole pairs of the motor.

For example, a four-pole motor supplied at 50 Hz rotates its magnetic field at 1500 rpm, while a two-pole motor at the same frequency gives 3000 rpm.

As explained in Section III.3.1, the rotor of a single-phase induction motor cannot reach the synchronous speed. The actual rotor speed  $N_r$  is always slightly lower. The difference between the synchronous speed and the actual rotor speed, expressed as a fraction of the synchronous speed, is called the slip and is denoted  $s$ :

$$s = \frac{N_s - N_r}{N_s} \quad (\text{III. 25})$$

where  $s$  is the slip (dimensionless),  $N_s$  is the synchronous speed [rpm], and  $N_r$  is the actual rotor speed [rpm].

The slip is often expressed as a percentage for clarity:

$$s (\%) = \frac{N_s - N_r}{N_s} \times 100 \quad (\text{III. 26})$$

From equation (III.25), the actual rotor speed can be expressed directly as a function of synchronous speed and slip:

$$N_r = (1 - s)N_s \quad (\text{III. 27})$$

where  $N_r$  is the actual rotor speed [rpm],  $s$  is the slip, and  $N_s$  is the synchronous speed [rpm].

Under normal full-load operation, the slip of a single-phase induction motor is typically between 1% and 5%, meaning the rotor runs very close to the synchronous speed. At standstill,  $s = 1$  (rotor speed is zero). At synchronous speed,  $s = 0$ , which is a theoretical condition that is never reached in practice, because zero slip means zero induced EMF, zero rotor current, and therefore zero torque.

When angular speed is preferred over rpm — for example in Simulink simulations — the synchronous speed in rad/s is obtained from:

$$\Omega_s = \frac{2\pi N_s}{60} \quad (\text{III. 28})$$

The rotor angular speed correspondingly becomes:

$$\Omega_r = (1 - s)\Omega_s \quad (\text{III. 29})$$

where  $\Omega_r$  is the actual rotor angular speed [rad/s],  $s$  is the slip, and  $\Omega_s$  is the synchronous angular speed [rad/s].

These equations form the foundation for understanding how the speed of a single-phase induction motor varies with load conditions and supply parameters. They are also essential for the voltage-based speed control method presented in the next section.

### III.3.3 Voltage-Based Speed Control Method:

The voltage-based speed control method is one of the simplest methods used to control the speed of a single-phase induction motor. In this method, the supply frequency remains constant, while the RMS value of the stator voltage is varied. Since the synchronous speed depends only on the supply frequency and the number of poles, changing the voltage does not directly change the synchronous speed. Instead, it changes the electromagnetic torque developed by the motor, which affects the slip and consequently the rotor speed.

From the previous section, the rotor speed is related to the synchronous speed and slip. By combining the synchronous speed equation with the slip relation, the motor speed can be written as follows:

$$N_r = \frac{60 \cdot f}{P} (1 - s) \quad (\text{III. 30})$$

where  $N_r$  is the motor speed [rpm],  $f$  is the supply frequency [Hz],  $P$  is the number of pole pairs, and  $s$  is the slip.

This relation is obtained by substituting the synchronous speed equation  $N_s = \frac{60 \cdot f}{P}$  into the rotor speed equation  $N_r = (1 - s)N_s$ . In voltage-based speed control, the frequency ( $f$ ) and the number of poles pairs ( $P$ ) are kept constant. Therefore, the speed variation is mainly caused by the change in slip.

The effect of the stator voltage on the electromagnetic torque is important in this method. For an induction motor supplied at constant frequency, the electromagnetic torque is approximately proportional to the square of the applied stator voltage.

$$C_{em} \propto V_s^2 \quad (\text{III. 31})$$

where  $C_{em}$  is the electromagnetic torque [N.m] and  $V_s$  is the effective RMS stator voltage [V].

This relation shows that reducing the stator voltage reduces the available electromagnetic torque strongly. For example, if the applied voltage is decreased, the torque capability decreases approximately with the square of the voltage. As a result, for the same load torque, the slip increases and the motor speed decreases. However, if the voltage is reduced too much, the motor may not produce enough torque to drive the load, which can lead to stalling.

In practice, the stator voltage of a single-phase induction motor can be adjusted using an AC voltage controller. This controller may be based on a TRIAC or two anti-parallel thyristors. By changing the

firing angle of the power switch, the RMS voltage applied to the motor is modified. In a simplified form, the effective stator voltage can be expressed as:

$$V_s = \alpha V_n \quad (\text{III. 32})$$

where  $V_s$  is the effective RMS voltage applied to the motor stator [V],  $V_n$  is the rated supply voltage [V], and  $\alpha$  is the voltage reduction factor, with  $0 < \alpha < 1$ .

This equation gives a simplified representation of the voltage control principle. In practice, the exact RMS voltage depends on the TRIAC firing angle, but the factor  $\alpha$  is used here to represent the voltage reduction level. When  $\alpha$  decreases, the effective stator voltage  $V_s$  decreases, which reduces the electromagnetic torque according to equation (III.31). As a result, the slip increases and the motor speed decreases according to equation (III.30).

The main advantage of voltage-based speed control is its simplicity, since it does not require changing the supply frequency. However, its speed range is limited because reducing the voltage also reduces the available torque. Therefore, this method is mainly suitable for low-power applications such as fans and pumps. In closed-loop operation, the measured speed is compared with the reference speed, and the controller adjusts the effective stator voltage to reduce the speed error.

#### III.3.3.1 Simulation Model in MATLAB/Simulink:

The simulation model was developed in MATLAB/Simulink to study the voltage-based speed control of a single-phase induction motor. In this model, the motor speed is controlled by changing only the effective stator voltage, while the supply frequency remains constant. The voltage is generated through a single-phase IGBT inverter controlled by PWM signals.

The reference speed is applied using a Step block, changing from 80 to 100 rad/s at  $t = 5$  s, in order to test the motor dynamic response. The measured speed is compared with the reference speed, and the error is processed by a PI controller to adjust the PWM control signal.

Fig. III.7 shows the complete Simulink model used for this voltage-based control system.

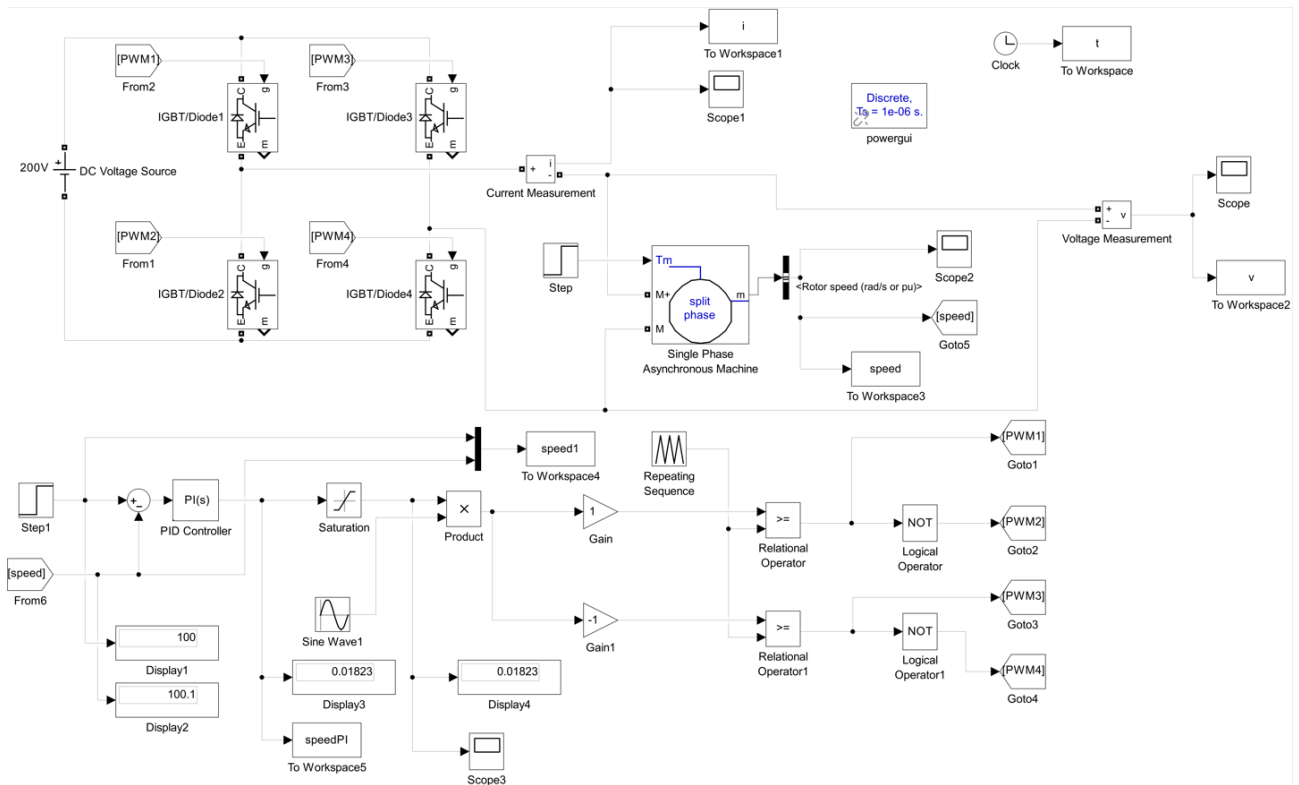


**Fig. III.11:** MATLAB/Simulink model of the single-phase AC motor speed control system.

### III.3.3.2 Simulation Results and Discussion:

Fig. III.8 shows the speed response of the single-phase AC motor using the voltage-based control method. The reference speed is first set to 80 rad/s, and the measured speed increases rapidly until it reaches the reference value with a short transient period. After that, the motor speed remains almost stable around the desired value.

At  $t = 5$  s, the reference speed changes from 80 to 100 rad/s. This sudden change produces a temporary disturbance in the measured speed, but the controller corrects the error and the motor speed quickly follows the new reference. The response shows that the voltage-based control method can regulate the motor speed, although a small transient variation appears during the change of reference speed.

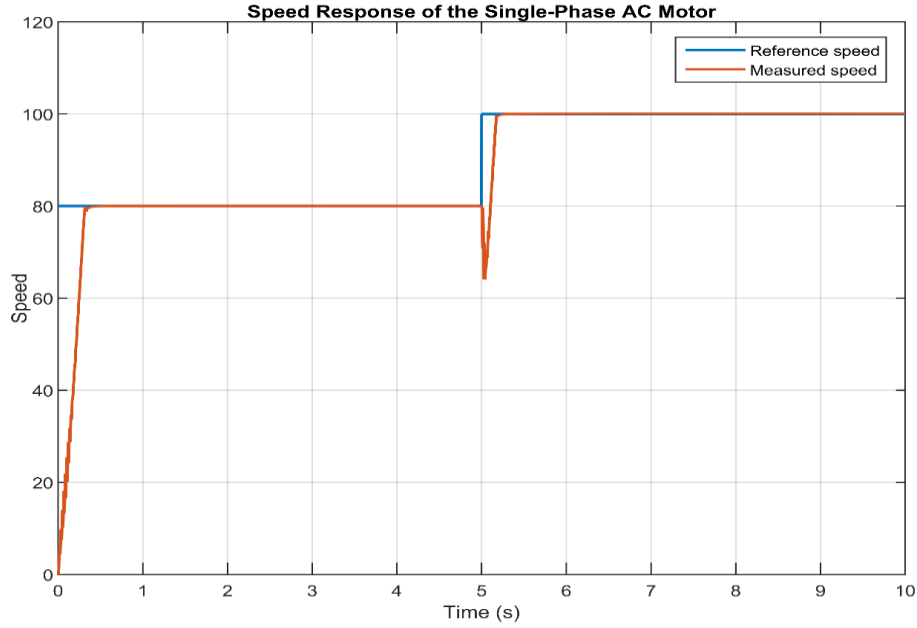


**Fig. III.12:** Speed response of the single-phase AC motor using voltage-based control.

#### III.3.4 Scalar Control Method Based on V/f Variation:

The scalar control method based on V/f variation is a more developed speed control technique compared with the simple voltage-based method. In voltage-based control, only the stator voltage is changed while the supply frequency remains constant. However, in V/f control, both the supply voltage and the supply frequency are varied at the same time. The main objective is to keep the ratio between voltage and frequency approximately constant in order to maintain a nearly constant magnetic flux in the motor.

For an induction motor, the synchronous speed is directly related to the supply frequency according to:

$$N_s = \frac{60.f}{p} \quad (\text{III. 33})$$

Therefore, changing the frequency allows the synchronous speed of the motor to be modified. However, if the frequency is reduced while the voltage remains constant, the magnetic flux may increase and cause magnetic saturation. On the other hand, if the frequency is increased without increasing the voltage, the magnetic flux decreases and the motor torque becomes weaker. For this reason, the stator voltage must be adjusted proportionally with the frequency.

The basic principle of V/f control can be expressed as:

$$\frac{V_s}{f} = \text{constant} \quad (\text{III. 34})$$

where  $V_s$  is the stator voltage and  $f$  is the supply frequency. This relation means that when the frequency decreases, the stator voltage also decreases. Similarly, when the frequency increases, the stator voltage increases up to the rated voltage. In this way, the motor flux remains approximately constant, and the motor can produce a suitable electromagnetic torque over a wider speed range.

In the Simulink model used in this work, the scalar control subsystem receives the reference speed and the measured motor speed. The measured speed is compared with the reference speed, and the resulting error is processed by a PI controller. The output of the controller is used to generate the angular frequency command. This frequency command is then converted into the corresponding voltage amplitude according to the V/f law. The generated voltage reference is finally used to control the single-phase inverter supplying the motor.

The voltage reference can be written in simplified form as:  $V_s^* = K_v f \Omega_s$

where  $V_s^*$  is the stator voltage reference,  $\Omega_s$  is the angular supply frequency, and  $K_v$  is the V/f proportionality constant. In the implemented model, this proportionality is represented by a gain based on the rated voltage and rated angular frequency. For a rated voltage of 220 V and a rated frequency of 50 Hz, the V/f gain is expressed as:

$$K_v f = \frac{220}{2\pi \cdot 50} \quad (\text{III. 35})$$

After calculating the voltage amplitude and the angular frequency, sinusoidal reference signals are generated for the inverter. These signals define the desired AC voltage waveform applied to the motor. The use of sinusoidal references allows the inverter to supply the motor with a variable-frequency and variable-voltage source.

Compared with voltage-based control, V/f scalar control provides better speed regulation because the speed is mainly controlled through frequency variation. In addition, maintaining an approximately constant V/f ratio helps preserve the motor flux and improves torque production. This makes scalar control more suitable for variable-speed operation of induction motors.

#### III.3.4.1 Simulation Model in MATLAB/Simulink:

The simulation model was implemented in MATLAB/Simulink to test the scalar V/f control of the single-phase asynchronous motor. The model consists of a speed reference, a feedback signal from the motor speed, a PI controller, a scalar control subsystem, a single-phase inverter, and the asynchronous motor.

The reference speed is compared with the measured speed, and the resulting error is applied to the PI controller. The controller output is sent to the scalar control subsystem, which generates the voltage and frequency references required for the inverter. These references are then used to supply the motor with a variable-voltage and variable-frequency AC signal.

The inverter is controlled by sinusoidal reference signals generated inside the scalar control block. This allows the motor speed to follow the imposed reference while maintaining stable operation. The complete Simulink model used for this control method is shown in Fig. III.8.

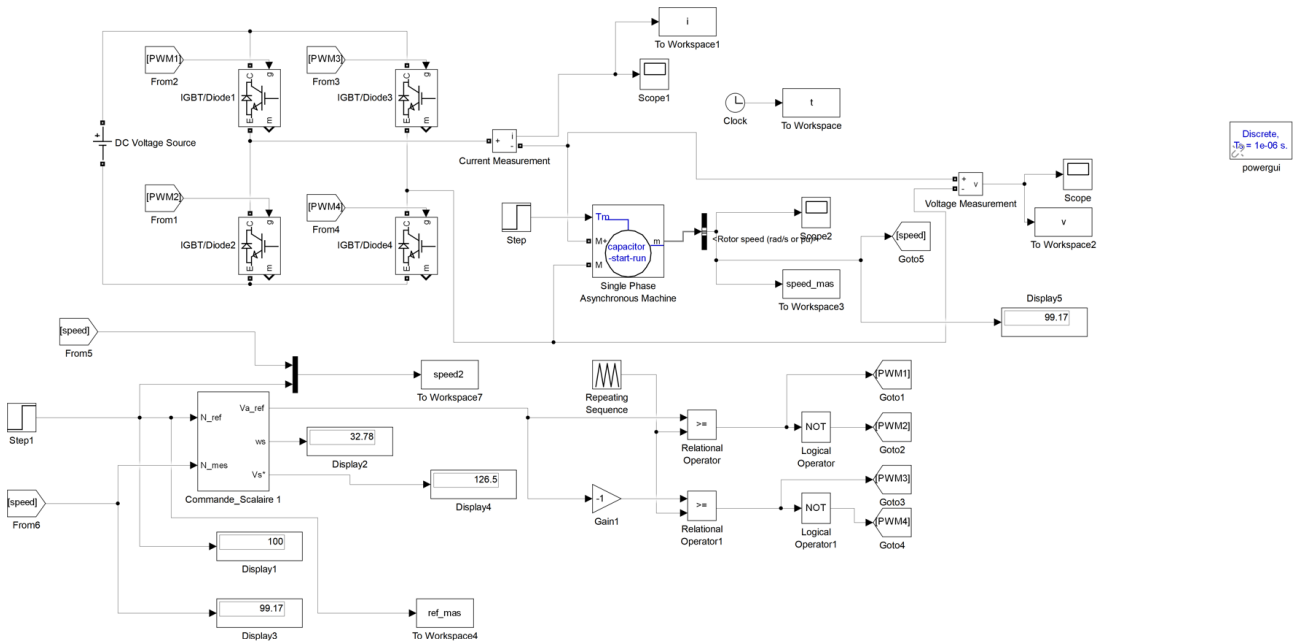


**Fig. III.13:** MATLAB/Simulink model of the single-phase AC motor using scalar V/f control.

### III.3.4.2 Simulation Results and Discussion:

Fig. III.10 shows the speed response of the single-phase AC motor using the scalar V/f control method. The reference speed is initially set to 80 rad/s. At the beginning of the simulation, the measured speed increases from zero and reaches the reference value after a short transient period. Then, the motor speed remains stable and close to the desired value.

At  $t = 5$  s, the reference speed is increased from 80 rad/s to 100 rad/s. The measured speed follows this change smoothly and reaches the new reference with a small transient response. After this period, almost no steady-state error is observed.

These results show that the scalar V/f control method provides good speed tracking and stable operation. By varying both the voltage and the frequency, the motor can follow the reference speed while maintaining an approximately constant V/f ratio.

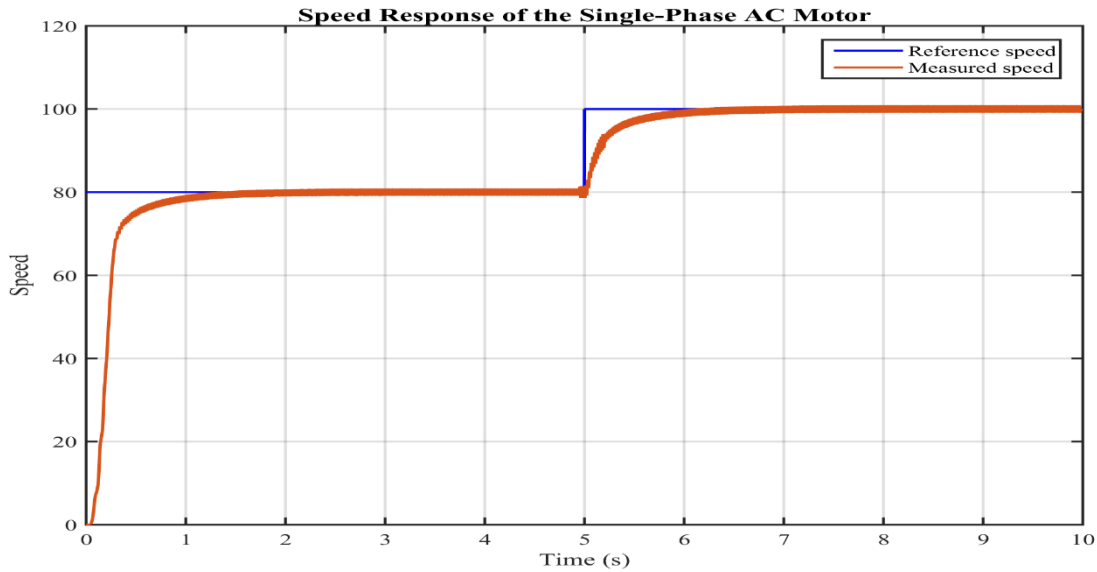


**Fig. III.14:** Speed response of the single-phase AC motor using scalar V/f control.

### III.3.4.3 Effect of Motor Resistance Variation on the Fixed PI Response

In this simulation, the motor resistance values were increased by 250% in order to study the effect of parameter variation on the fixed PI controller. The new resistance values were  $R_s = 5.05 \, \Omega$ ,  $R_r = 10.30 \, \Omega$ , and  $R_{aux} = 17.85 \, \Omega$ .

Fig. III.11 shows the speed response after this resistance variation. At the beginning of the simulation, the motor speed reaches the reference value, but a noticeable overshoot appears before the speed becomes stable. At  $t = 5 \, \text{s}$ , the reference speed changes from 80 rad/s to 100 rad/s. The motor follows the new reference, but another transient overshoot appears before the speed stabilizes.

This result shows that the fixed PI controller is affected by the variation of motor parameters. Although the speed reaches the reference value in steady state, the transient response becomes less smooth. Therefore, an adaptive correction method such as ANFIS can be added to improve the dynamic response under parameter variations.

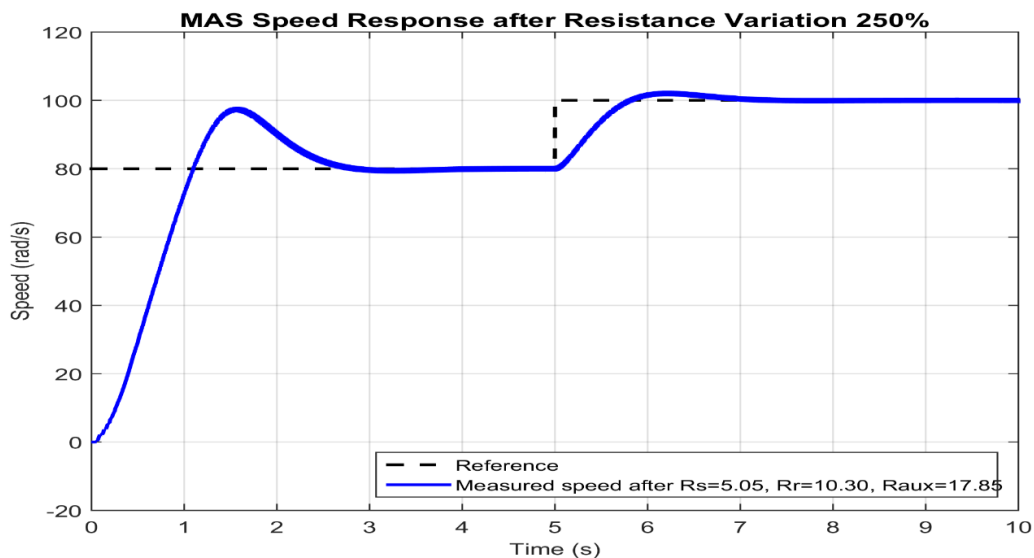


**Fig. III.15:** Speed response of the single-phase AC motor after 250% resistance variation.

### III.3.5 Adaptive Neuro-Fuzzy Inference System (ANFIS)

#### III.3.5.1 Neural networks :

Neural networks form a family of nonlinear functions that make it possible to build, through learning, a very wide class of models and controllers. A neural network is a system of interconnected nonlinear operators that receives signals from the outside through its inputs and delivers output signals, which are in fact the activities of certain neurons.

##### A. formal neuron:

The formal neuron is a simplified mathematical model of the biological neuron. It has a number of inputs, the dendrites, a body that processes the inputs according to the all-or-nothing principle, and an axon that carries the neuron's response.

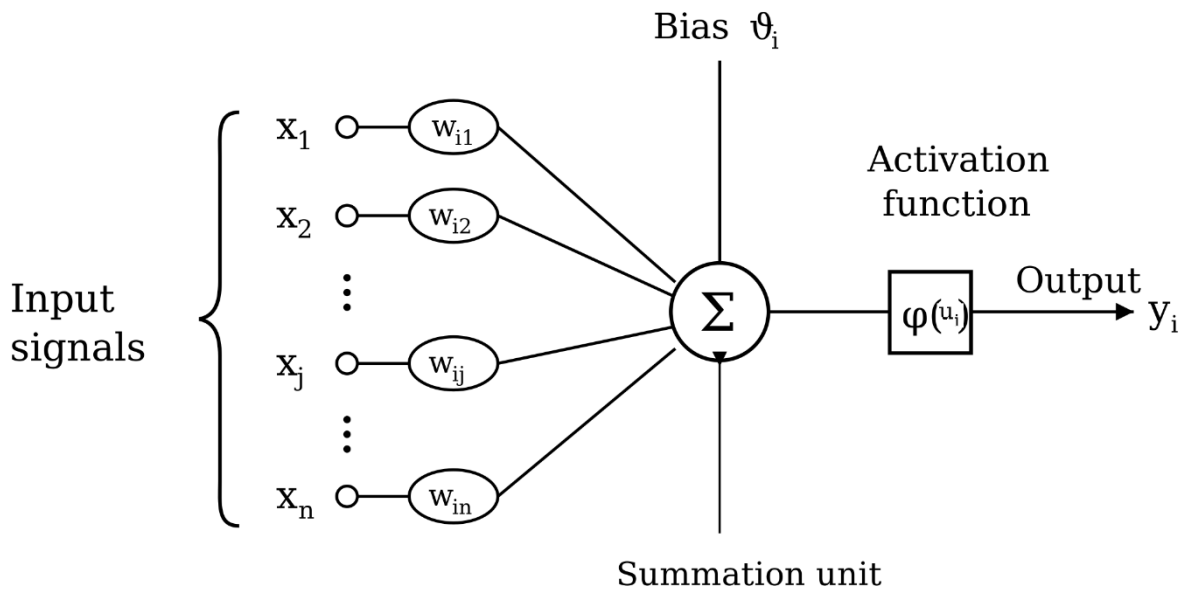


**Fig. III.16:** Basic model of a formal neuron

The output  $u_i$  of the formal neuron is given by the relation:

$$u_i = \sum_{j=0}^n w_{ji} x_j \quad (\text{III. 36})$$

From this value, an activation function computes the output value  $y$  of the neuron. It is this value that will be transmitted to the downstream neurons:

$$y_i = \varphi(u_j) \quad (\text{III. 37})$$

There are many possible forms for the activation function. The most common ones are shown in the following figure. It will be noted that, unlike biological neurons whose state is binary, most activation functions are continuous, offering an infinite number of possible values within the interval  $[0, +1]$  (or  $[-1, +1]$ ).

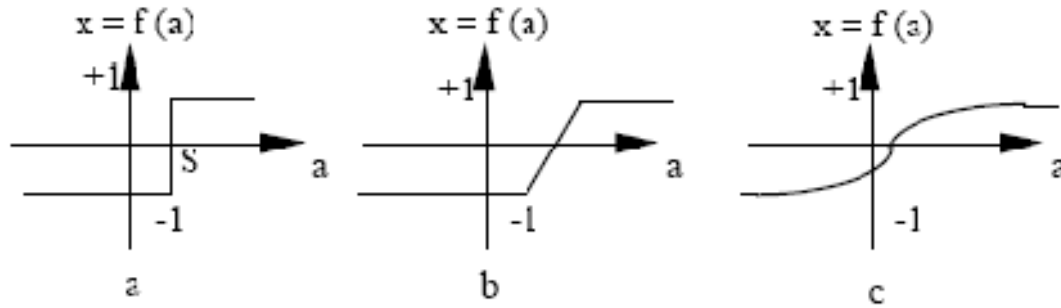


**Fig. III.17:** Différents types de fonctions d'activation pour le neurone formel

a: threshold function (S: threshold value), b: piecewise linear, c: sigmoid.

**B. Learning methods:** In expert systems, the expert's knowledge has an enumerated form: it is expressed as rules. In the case of neural networks, knowledge has a distributed form: it is encoded in the connection weights, the network topology, the transfer functions of each neuron, the threshold of these functions, and the learning method used. There are a number of learning methods.

**B.1 Hebb's rule:** This is the oldest learning method, and it is inspired by biology. It reflects the strengthening of the connections linking two activated neurons. If at least one of the two neurons is not activated, the weight of the connection is not modified.

Here is Hebb's rule:

$$w_{ji}(t + 1) = w_{ji}(t) + \Delta w_{ji} \quad (\text{III.38})$$

with  $\Delta w_{ji} = \eta \cdot x_j \cdot x_i$

where:

- $\Delta w_{ji}$  is the weight of the connection between neuron i and neuron j,
- $\eta$  is the learning rate (a positive constant),
- $x_i$  is the activation (output) of neuron i.
- $x_j$  is the activation (output) of neuron j.

## B.2 Error gradient backpropagation:

Backpropagation is a learning algorithm for feedforward (layered) neural networks, which have an input layer, one or more hidden layers, and an output layer, with no loops or same-layer connections. An input vector is fed forward through the layers to produce an output, which is then compared to the desired output to compute an error. The gradient of this error is propagated backward (from output to input) to adjust the network's weights. The process is repeated for each input vector until a stopping criterion is met.

## C. The perceptron:

Following the results of the work of McCulloch and Pitts (1943), which led to the definition of the formal neuron, as well as those of Hebb, explaining the learning effects of memory and conditioning based on groups of cells. To explain these learning effects, Hebb proposed that cells learn to modify the strength of the connections linking them according to their simultaneous activity.

The idea of some researchers was to use models of neurons and of the evolution of synapses to simulate neural networks. The first solid model was presented in 1959 by F. Rosenblatt . This is the perceptron, in other words a network reduced to a single formal neuron.

#### D. Process identification using neural networks:

The principle of identification by neural networks consists in replacing classical parametric models with neural models, that is, proposing a model establishing a relationship between its input and its output, and determining, from the pair of input-output signals, the behavior of the model. Two important reasons motivate us:

- to predict the behavior of a system for different operating conditions;
- to develop a control law to be applied to the process so that it achieves the assigned objective.

We will mention two identification techniques based on multilayer neural networks: the direct identification method and the inverse identification method.

**D.1 Direct identification:** The figure shows the general scheme of direct identification of a process. In this figure, the neural network identifier (NNI) is used in parallel with a black-box-type process. The output of the process,  $y$ , is compared with the output of the neural network  $\hat{y}$ , and then the error  $e = y - \hat{y}$  is used to fine-tune the parameters of the neural system.

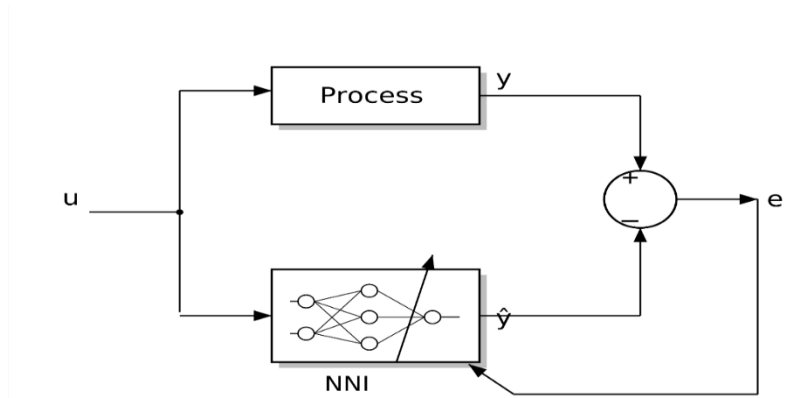


**Fig. III.18:** Scheme of direct identification of a process using a neural network.

The differential equation of the output  $y$  at  $t+1$  can be written as follows:

$$y(t+1) = f(y(t), \dots, y(t-n+1), \dots, u(t), \dots, u(t-m+1)) \quad (\text{III. 39})$$

where  $y(t+1)$  is the output of the process at time  $t+1$ , and  $f$  is the nonlinear function governing the operation of the process.

The output of the neural network at time  $t+1$  is described as follows:

$$\hat{y}(t+1) = \hat{f}(y(t), \dots, y(t-n+1), \dots, u(t), \dots, u(t-m+1)) \quad (\text{III. 40})$$

where  $\hat{f}$  it represents the nonlinear approximation function of the process function  $f$ . The output of the neural network  $\hat{y}(t+1)$  depends on the previous outputs and inputs of the process, up to orders  $n$  and  $m$  respectively.

### D.2 Inverse identification:

In this method, the input of the process is compared with the output of the neural identifier (NNI), and the output of the process is fed as the input of the neural network.

After offline learning of the inverse model, the NNI can be configured to provide direct control of the process.

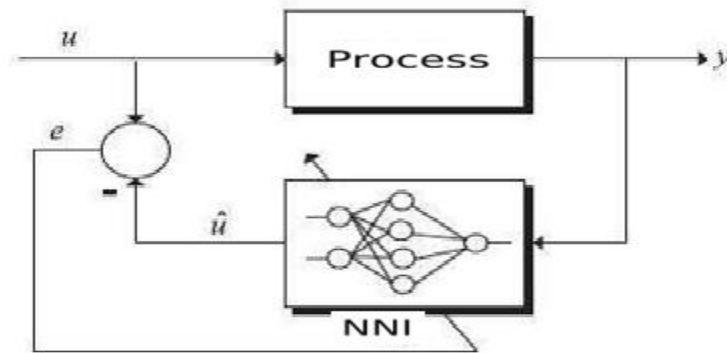


**Fig. III.19:** Scheme of inverse identification of a process using a neural network.

### III.3.5.2 Fuzzy logic:

Fuzzy logic replaces binary (true/false) logic with a logic in which variables can take intermediate values to a certain degree, reflecting human reasoning based on imprecise or incomplete data. Its theoretical foundations were established in 1965 by Professor Lotfi A. Zadeh, but were not taken seriously at first. Initially applied in non-technical fields (commerce, law, medicine, etc.) to enhance expert systems with decision-making ability, it was only later, mainly in Japan from 1985, that it saw significant growth in control and regulation systems, being used for industrial problems in energy, transport, machine tools, and robotics.

#### A. Structure of a fuzzy logic control:

Unlike classical control techniques, fuzzy logic control does not use well-defined or precise mathematical formulas or relations. Instead, it handles inferences with several fuzzy rules based on the fuzzy operators AND, OR, THEN, etc., applied to linguistic variables.

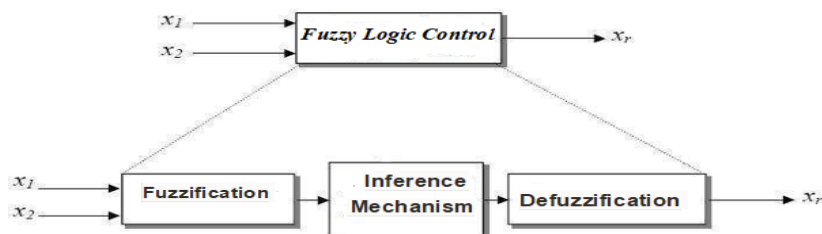


**Fig. III.20:** Internal structure of a fuzzy controller.

The most commonly used defuzzification method, that of determining the center of gravity of the resulting membership function, is given by the following general relation:

$$x_0 = \frac{\sum_{i=1}^n x_i \mu_{\text{res}}(x_i)}{\sum_{i=1}^n \mu_{\text{res}}(x_i)} \quad (\text{III. 41})$$

With:

$x_i$ , element  $i$  of the base of  $\mu_{\text{res}}(x_i)$  ( $i = 1, \dots, n$ )

### III.3.5.3 Neuro-Fuzzy control structure:

The joint use of neural networks and fuzzy logic makes it possible to combine the advantages of both methods: the learning capabilities of the former and the readability and flexibility of the latter. Various combinations of these two methods have been developed since 1988. They have given rise to neuro-fuzzy systems, which are most often oriented toward the control of complex systems and classification problems.

#### A. Methods of Neuro-Fuzzy combinations

- There are four main categories of combinations of neural networks with fuzzy logic:

Neural fuzzy network: fuzzy techniques enhance the learning capabilities of neural networks.

- Simultaneous neural/fuzzy system: the neural network and the fuzzy system work together on the same task without influencing each other.
- Cooperative neuro-fuzzy models: the neural network determines the parameters (rules and fuzzy sets) of a fuzzy system, which then operates on its own.
- Hybrid neuro-fuzzy models: the neural network and the fuzzy system are combined in a homogeneous architecture; this is the most widespread form in practice.

Various combinations of these two methods have been developed since 1988. They have given rise to neuro-fuzzy systems, which are most often oriented toward the control of complex systems and classification problems.

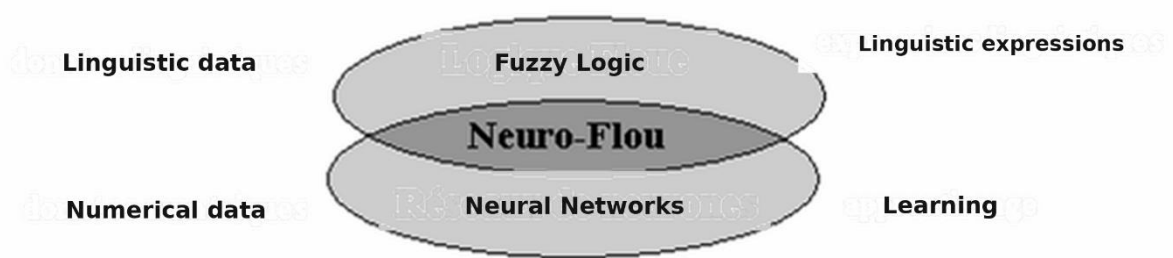


**Fig. III.21:** Principle of the Neuro-fuzzy system

#### B. Integrated Hybrid Neuro-Fuzzy Systems

Integrated hybrid neuro-fuzzy systems constitute the most elaborate form of the neural network–fuzzy logic association. Three main types can be distinguished.

##### Mamdani-type Integrated Neuro-Fuzzy Systems

The Mamdani-type neuro-fuzzy system uses backpropagation to learn the parameters of the membership functions. It contains five layers:

Layer 1 Input: transmits the input values directly to the next layer.

Layer 2 Fuzzification: each node computes the degree of membership of an input to a linguistic label. A classification algorithm determines the initial number and type of the membership functions.

Layer 3 Antecedent rules: one node per rule computes the firing strength via a Min or product operator.

Layer 4 Consequent rules: combines the antecedents and determines the degree of membership to the output linguistic label.

Layer 5 Defuzzification: provides the crisp output using the center of gravity method.

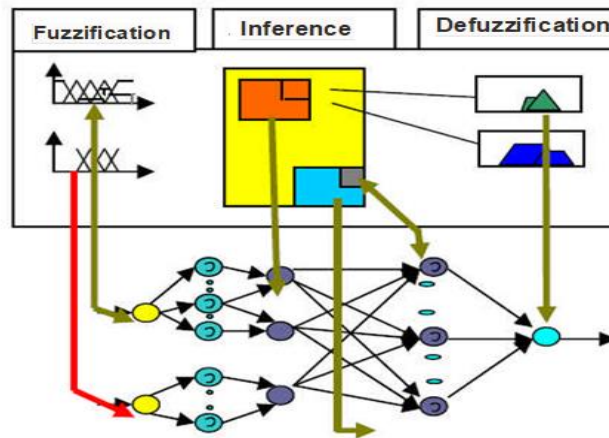


**Fig. III.22:** Operating principle of a Mamdani-type Neuro-fuzzy network

#### Takagi-Sugeno-type Integrated Neuro-Fuzzy Systems

Takagi-Sugeno-type neuro-fuzzy systems combine backpropagation for the membership functions and the least squares method for the linear coefficients of the rule conclusions. This system contains six layers:

Layers 1, 2, 3: identical to the Mamdani-type system.

Layer 4 Normalization: computes the ratio of the firing strength of rule  $i$  to the sum of the firing strengths of all rules.

Layer 5 Rule consequent layer: parameters determined by the least mean squares algorithm.

Layer 6 Inference: the single node computes the overall output by summation.

#### C. ANFIS System (Adaptive Network Fuzzy Inference System)

The ANFIS system is a class of adaptive network proposed by Jang. It is the most widely used model in practice. Applications in signal processing, adaptive filtering, and motor control have been carried out with this architecture. This architecture refines the fuzzy rules obtained from human experts to describe the input-output behavior of a complex system. It is based on the first-order Takagi-Sugeno inference model.

#### D. Presentation of the ANFIS Structure (Adaptive Network Fuzzy Inference System)

The ANFIS structure proposed by J-S.R. Jang is a five-layer structure. Consider a first-order Sugeno-type system with two input variables  $x_1$ ,  $x_2$  and one output  $y$ , whose rule base contains:

Rule 1: If  $x_1$  is  $A_1$  AND  $x_2$  is  $B_2$ , Then  $y_1 = p_1 x_1 + q_1 x_2 + r_1$

Rule 2: If  $x_1$  is  $A_2$  AND  $x_2$  is  $B_1$ , Then  $y_2 = p_2 x_1 + q_2 x_2 + r_2$

where  $p_i, q_i, r_i$  are the consequent parameters. The ANFIS network is a multilayer network with unit weights (adaptive square nodes / fixed circular nodes):

$$(O_i^k = f(O_i^{k-1}, \dots, O_{n-1}^{k-1}; p, q, r, \dots)) \quad (\text{III.42})$$

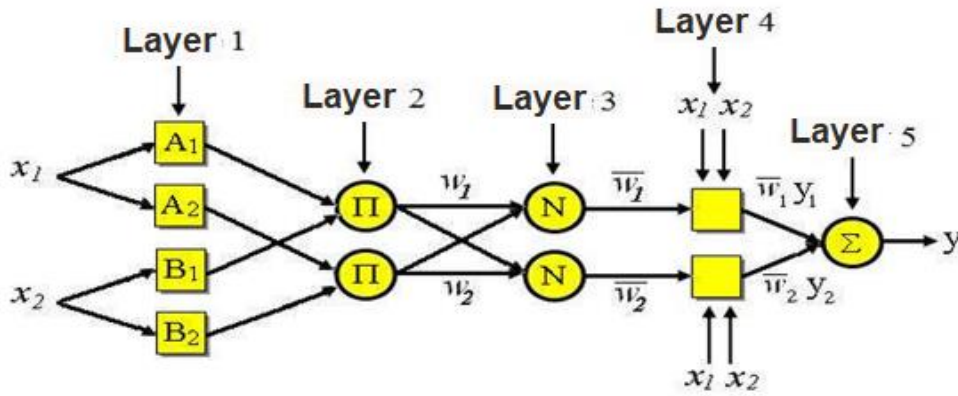


**Fig. III.23:** Architecture of the 5-layer ANFIS network

#### E. Conclusion:

Hybrid Neuro-fuzzy systems are mainly used when the mathematical model of the physical system is difficult to develop; they exploit fuzzy rules derived from human expertise to model the dynamic behavior of the system. Neuro-fuzzy-based control makes it possible to define a digital control law based on the learning capabilities of the neural network and the readability and flexibility of fuzzy logic. In this work, we used ANFIS because it overcomes the main limitations of fuzzy logic and neural networks taken separately. Indeed, a purely fuzzy system cannot learn or automatically adjust its membership functions and rules, which must be set manually by an expert, while a purely neural network behaves like a black box that lacks interpretability. ANFIS combines the two approaches within a single architecture: it keeps the readability and the linguistic interpretation of fuzzy logic, while using the learning ability of neural networks to automatically optimize the parameters of the membership functions and the rule consequents. This results in a more accurate, faster-converging, and more reliable model, which justifies our choice of ANFIS.

#### III.3.5.4 ANFIS-Based Adaptive Correction:

To improve the speed response under motor parameter variation, an ANFIS-based adaptive correction was added to the fixed PI controller. In this control structure, the reference speed and the measured speed are compared to generate the speed error. The PI controller provides the main control action, while the ANFIS block generates an adaptive correction signal to compensate the effect of resistance variation.

The corrected control signal is then used inside the scalar control subsystem to generate the frequency and voltage references applied to the inverter. In this way, the PI + ANFIS structure improves the robustness of the control system without changing the fixed PI gains.

Fig. III.20 shows the MATLAB/Simulink implementation of the PI + ANFIS adaptive control structure used in this work.

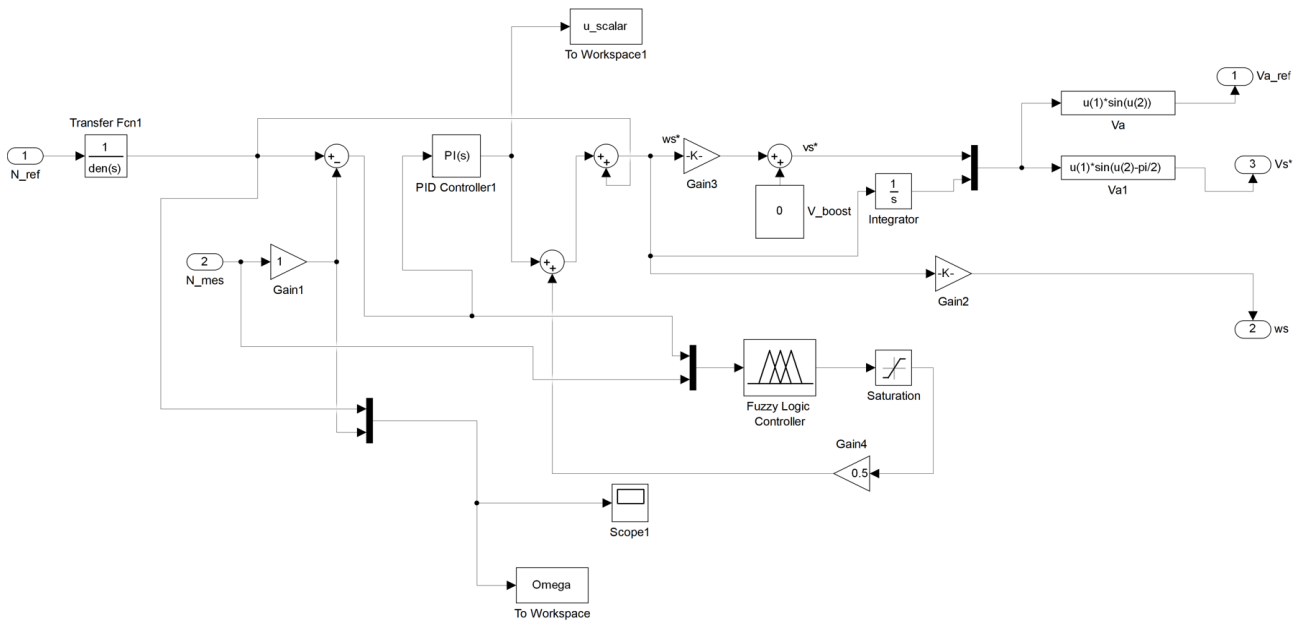


**Fig. III.24:** MATLAB/Simulink implementation of the PI + ANFIS adaptive control structure.

The speed response obtained with the PI + ANFIS controller after resistance variation is shown in Fig. III.21. Compared with the fixed PI response, the overshoot is clearly reduced and the speed follows the reference more smoothly. At  $t = 5$  s, when the reference speed changes from 80 rad/s to 100 rad/s, the adaptive controller provides a faster and more stable response with almost no steady-state error.

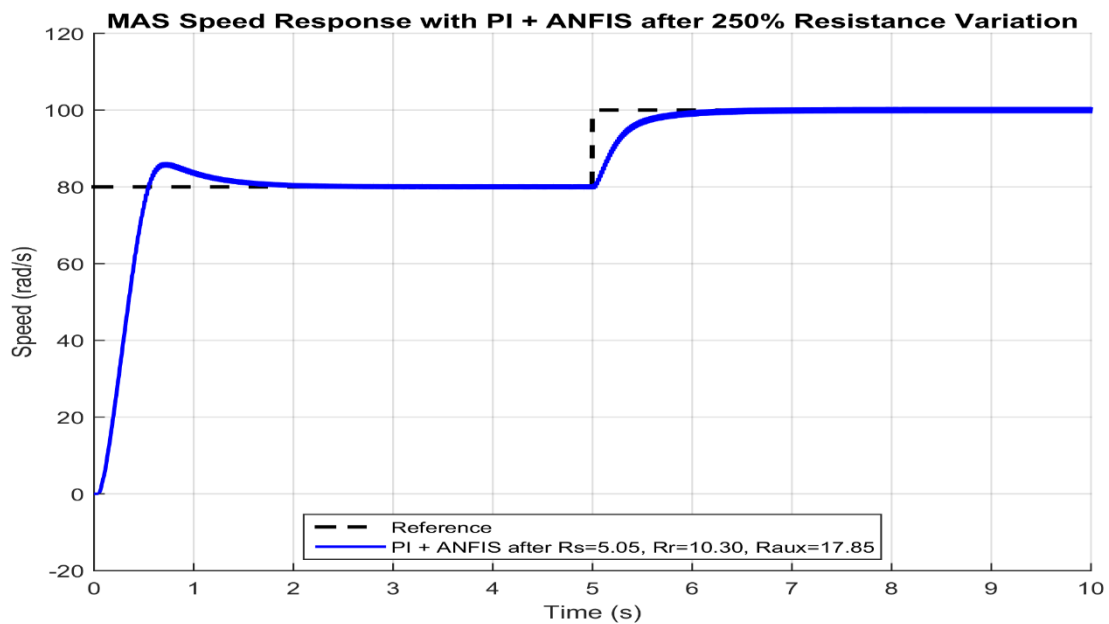


**Fig. III.25:** Speed response of the single-phase AC motor with PI + ANFIS after resistance variation.

These results confirm that the ANFIS-based correction improves the dynamic performance of the control system and makes it more robust against motor resistance variation.

The ANFIS controller used in this work is a Sugeno-type fuzzy inference system. It has two inputs and one output. The first input is the speed tracking error, defined as the difference between the reference speed and the measured motor speed. The second input is the variation of the speed error. The output of the ANFIS block is an adaptive correction signal added to the fixed PI controller output.

The adopted control law can be written as:

$$u_{\text{(final)}} = u_{\text{(PI)}} + \Delta u_{\text{(ANFIS)}}$$

where  $u_{\text{(PI)}}$  is the output of the fixed PI controller,  $\Delta u_{\text{(ANFIS)}}$  is the adaptive correction generated by the ANFIS controller, and  $u_{\text{(final)}}$  is the final control signal applied to the scalar control stage.

The ANFIS structure contains three generalized bell membership functions for each input, which gives a total of nine fuzzy rules. The output membership functions are linear, as commonly used in Sugeno-type ANFIS systems. The training data were obtained from MATLAB/Simulink simulations of the motor drive system and arranged in the form [input 1, input 2, output]. A total of 30000 training samples were used.

The neural structure of the proposed ANFIS controller is shown in Fig. III.22. This structure illustrates the two input variables, the membership function layer, the rule layer, the normalization layer, the consequent layer, and the final adaptive correction output.

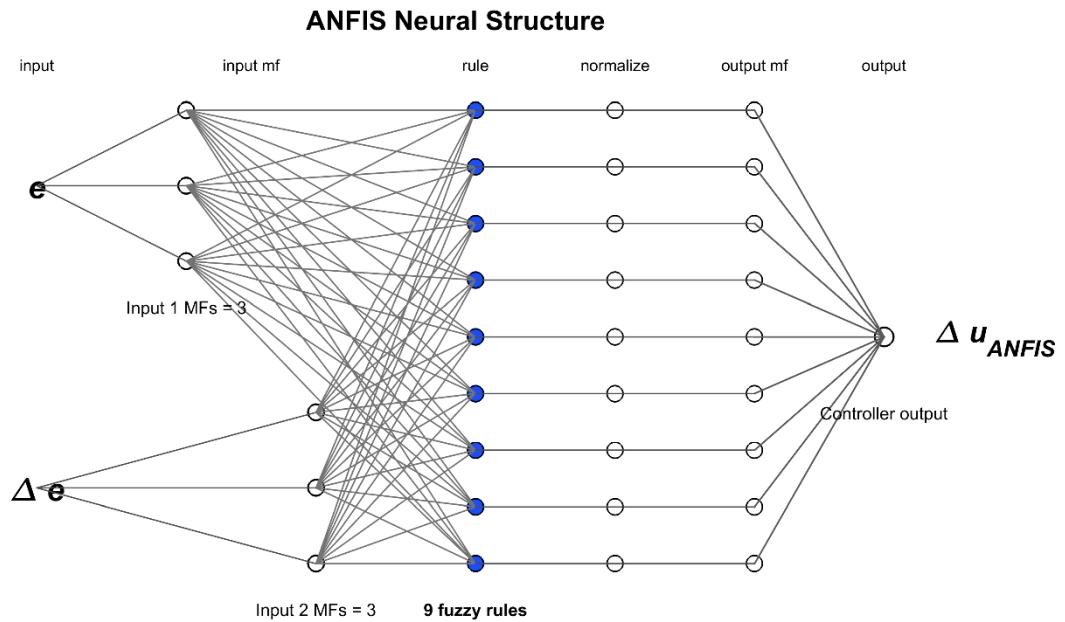


**Fig. III.26:** Neural structure of the proposed ANFIS controller.

The ANFIS was trained using the hybrid learning algorithm for 25 epochs. The final training error was 0.0000845613, which indicates that the trained ANFIS was able to learn the required correction behavior with good accuracy. After training, the obtained FIS file was inserted into the Simulink model using the Fuzzy Logic Controller block.

In this work, the PI gains were kept constant. Therefore, the improvement obtained with the PI + ANFIS structure is due to the adaptive correction generated by the ANFIS block, not to a manual retuning of the PI controller.

### III.3.5.5 Comparative Results between Fixed PI and ANFIS:

Fig. III.21 presents the comparison between the fixed PI controller and the PI + ANFIS controller after motor resistance variation. With the fixed PI controller, the speed response shows a noticeable overshoot during the starting period and after the reference speed change at  $t = 5$  s. This means that the fixed controller is affected by the change in motor parameters.

After adding the ANFIS adaptive correction, the speed response becomes smoother and closer to the reference. The overshoot is reduced, and the motor reaches the desired speed with better stability. This shows that the ANFIS block improves the dynamic behavior of the system without manually retuning the PI gains.

From this comparison, it can be concluded that the PI + ANFIS controller gives better performance than the fixed PI controller under resistance variation. Therefore, the ANFIS-based correction is more suitable for improving the robustness of the scalar V/f control system.

## III.4 Speed Control of a Three-Phase Induction Motor:

### III.4.1 Introduction:

The three-phase induction motor is widely used in industry due to its robustness, reliability, and low maintenance. For variable-speed operation, scalar V/f control varies the stator voltage and frequency together to keep the magnetic flux nearly constant and maintain torque capability. This method is simpler than vector control but has lower dynamic performance, especially at low speeds or during fast transients. The section presents the motor principle, models, slip, torque, V/f control with PI controller, PWM inverter, and MATLAB/Simulink simulation results.

### III.4.2 Operating Principle of the Three-Phase Induction Motor:

A three-phase induction motor works through the interaction between the rotating magnetic field produced by the stator and the currents induced in the rotor. The rotor speed is always slightly lower than the synchronous speed, creating slip, which is necessary to induce rotor currents and produce electromagnetic torque. Since motor speed mainly depends on stator frequency and torque depends on flux and slip, speed control is achieved by varying frequency while adjusting voltage to maintain the magnetic flux and torque capability.

### III.4.3 Parameter Identification of the Asynchronous Machine

To establish an accurate model of the three-phase induction motor, the electrical parameters of the machine must be identified experimentally. This section presents the three classical tests used for this purpose: the DC test for the stator resistance, the no-load test for the magnetizing inductance, and the blocked-rotor test for the leakage inductances and the rotor resistance. The procedure used to determine the moment of inertia and the friction coefficient is also presented.

#### III.4.3.1 Stator Resistance Measurement (DC Test)

The stator resistance  $R_s$  is obtained by applying a DC voltage between two stator terminals and measuring the corresponding current, in the same manner as for the DC machine armature resistance.

### III.4.3.2 No-Load Test (Determination of the Magnetizing Inductance)

The motor is run unloaded at its nominal voltage and frequency ( $V_n, f_n$ ). Under this condition, the no-load current  $I_0$  and the total active power  $P_0$  are measured. The following quantities are then calculated:

$$Q_0 = \sqrt{(3V_{\text{phase}}I_0)^2 - P_0^2}$$

$$L_s = \frac{Q_0}{3\omega_s I_0^2} \quad \text{with} \quad \omega_s = 2\pi f_n$$

$$L_s \approx M + L_{fs}$$

where  $Q_0$  is the no-load reactive power,  $L_s$  is the cyclic stator inductance,  $M$  is the magnetizing (mutual) inductance, and  $L_{fs}$  is the stator leakage inductance.

### III.4.3.3 Blocked-Rotor Test (Determination of $R_r$ and $L_f$ )

The rotor shaft is mechanically locked, and a reduced voltage  $V_{cc}$  is applied to the stator so as to reach the nominal current  $I_n$ . The voltage  $U_{cc}$ , the current  $I_n$ , and the active power  $P_{cc}$  are then measured.

The total resistance seen from the stator side is calculated as:

$$R_{cc} = \frac{P_{cc}}{3I_n^2}$$

The rotor resistance is then deduced as:

$$R_r = R_{cc} - R_s$$

The total leakage reactance and leakage inductance are obtained from:

$$X_{cc} = \frac{\sqrt{(3V_{cc}I_n)^2 - P_{cc}^2}}{3I_n^2}$$

$$L_f = \frac{X_{cc}}{\omega_e}$$

Assuming the stator and rotor leakage inductances are equal, each is taken as half of the total leakage inductance:

$$L_{fs} = L_{fr} = \frac{L_f}{2}$$

### III.4.3.4 Moment of Inertia and Friction Coefficient

The mechanical power dissipated by friction is obtained by subtracting the stator copper losses from the total no-load active power:

$$P_{mec} = P_0 - 3R_s I_0^2$$

The corresponding total loss torque is then expressed as:

$$C_{\text{pertes}} \approx \frac{P_{\text{vide}} - R_s I_0^2}{\Omega_s}$$

The moment of inertia  $J$  of the rotating assembly can be obtained either from the deceleration test, using the angular deceleration  $\alpha = \frac{\Delta\Omega}{\Delta t}$ , or from the run-down time  $t_{\text{stop}}$ :

$$J = \frac{C_{\text{pertes}}}{\alpha}$$

$$J \approx \frac{2P_{\text{mec}} t_{\text{stop}}}{\Omega_n^2}$$

The friction coefficient  $f$  and the friction torque  $C_{\text{frott}}$  are then deduced as:

$$f = \frac{C_{\text{frott}}}{\Omega_s} = \frac{P_{\text{mec}}}{\Omega_s^2}$$

$$C_{\text{frott}} = \frac{P_{\text{mec}}}{\Omega_s}$$

#### III.4.3.5 Power Measurement by the Two-Wattmeter Method

The active power absorbed by the induction motor (MAS) is measured using the two-wattmeter method, in which two wattmeters are connected to two of the three supply lines, as shown in the measurement set-up. The total active power is the sum of the two wattmeter readings:

$$P_0 = P_T = P_1 + P_2$$

The reactive power is obtained from the difference between the two readings:

$$Q = (P_1 - P_2)$$

#### III.4.4 Basic Electrical and Mechanical Model:

The modeling of a three-phase induction motor is generally based on a balanced three-phase machine supplied by sinusoidal voltages or by a PWM inverter producing quasi-sinusoidal voltages. For scalar control, the model is usually simplified in order to obtain practical relations between the stator voltage, the supply frequency, the magnetic flux, and the rotor speed.

For one stator phase, the electrical equation can be written as:

$$V_s = R_s i_s + \frac{d\Psi_s}{dt} \quad (\text{III. 43})$$

where  $v_s$  is the stator phase voltage,  $R_s$  is the stator resistance,  $i_s$  is the stator current, and  $\Psi_s$  is the stator flux linkage. This equation shows that the applied stator voltage is used to overcome the resistive voltage drop and to establish the magnetic flux in the machine.

In sinusoidal steady-state operation, this relation can be expressed approximately as:

$$V_s = R_s I_s + j\Omega_s \Psi_s \quad (\text{III. 44})$$

where  $\Omega_s$  is the stator angular frequency. At medium and high frequencies, the resistive voltage drop  $R_s I_s$  is relatively small compared with the flux term. Therefore, the stator voltage can be approximated by:

$$V_s \approx \Omega_s \Psi_s \quad (\text{III. 45})$$

From this expression, the stator flux can be written as:

$$\Psi_s \approx \frac{V_s}{\Omega_s}$$

This relation is very important in scalar control. It shows that the magnetic flux can be maintained approximately constant if the ratio between stator voltage and stator frequency is kept constant. This is the basic idea of the V/f control method.

The mechanical behavior of the induction motor is described by the dynamic equation:

$$J \frac{d\Omega_r}{dt} + f\Omega_r = C_{em} - C_r \quad (\text{III. 46})$$

where  $J$  is the moment of inertia,  $f$  is the viscous friction coefficient,  $\omega_m$  is the mechanical rotor speed,  $C_{em}$  is the electromagnetic torque, and  $C_r$  is the load torque. This equation shows that the motor accelerates when the electromagnetic torque is greater than the load torque and decelerates when the load torque becomes dominant.

In the Laplace domain, and by considering the net torque as the input and the rotor speed as the output, the simplified mechanical transfer function becomes:

$$\frac{\Omega_r(s)}{C_{net}(s)} = \frac{1}{Js+f} \quad (\text{III. 47})$$

This first-order mechanical model is useful for the design of the speed controller. In this work, it is mainly used to justify the use of a PI controller in the scalar speed control structure.

#### III.4.5 Speed, Slip, and Electromagnetic Torque:

The difference between these two speeds is represented by the slip. The slip is defined as:

$$s = \frac{N_s - N_r}{N_s}$$

where  $s$  is the slip,  $N_s$  is the synchronous speed, and  $N_r$  is the rotor speed. The slip can also be written using angular speeds:

$$s = \frac{\Omega_s - \Omega_r}{\Omega_s}$$

Slip is necessary for torque production. If the rotor rotates at exactly the synchronous speed, there is no relative motion between the rotating magnetic field and the rotor conductors. Therefore, no voltage is induced in the rotor, no rotor current flows, and no electromagnetic torque is produced.

The electromagnetic torque of the induction motor depends mainly on the magnetic flux and the slip. In a simplified form, this relationship can be expressed as:

$$C_{em} \propto \Psi_s^2 s \quad (\text{III. 48})$$

The torque capability of an induction motor depends strongly on the magnetic flux, so scalar control keeps the flux nearly constant by maintaining a constant V/f ratio. For low slip values, torque increases almost linearly with slip, but after reaching a maximum, it decreases. Therefore, scalar control is suitable for standard industrial speed control, but less effective when a very fast torque response is required.

#### III.4.6 Scalar Control Strategy Based on V/f Law:

The scalar control strategy of a three-phase induction motor is mainly based on the relationship between the stator voltage, the stator frequency, and the magnetic flux. The objective of this method is to vary the motor speed by changing the supply frequency while keeping the magnetic flux approximately constant.

From the simplified stator voltage equation, and by neglecting the stator resistance voltage drop at medium and high frequencies, the stator voltage can be written as:

$$V_s \approx \omega_s \Psi_s$$

Since the stator angular frequency is related to the supply frequency by:

$$\omega_s = 2\pi f_s$$

the stator flux can be expressed as:

$$\Psi_s \approx \frac{V_s}{2\pi f_s} \quad (\text{III. 49})$$

This equation shows that the stator flux depends on the ratio between the stator voltage and the stator frequency. Therefore, in order to keep the flux nearly constant, the voltage must be varied proportionally with the frequency. This gives the classical V/f law:

$$\frac{V_s}{f_s} = \text{constant}$$

In practice, when the reference speed increases, the controller increases the supply frequency. At the same time, the stator voltage is increased according to the V/f ratio. This allows the motor to operate with an almost constant magnetic flux and to maintain a suitable torque capability in the nominal operating range.

At low frequencies, the stator resistance voltage drop becomes more important and can no longer be neglected. If a simple proportional V/f law is used, the flux may decrease, which reduces the available electromagnetic torque. To compensate for this effect, a low-frequency voltage boost can be added. The voltage reference can then be written as:

$$V_s = V_0 + K_f f_s \quad (\text{III. 50})$$

where  $V_0$  is the voltage boost used at low speed, and  $K_f$  is the proportional coefficient of the V/f law. This compensation improves the starting torque and helps the motor operate more correctly at low speed.

The scalar control method is simple and easy to implement because it does not require a detailed decoupling between flux and torque. However, its dynamic performance is limited compared with vector control. For this reason, it is mainly used in applications where the speed variation is required but a very fast torque response is not necessary, such as pumps, fans, and conveyor systems.

In this work, the V/f control principle is used as the basic strategy for controlling the speed of the three-phase induction motor. The supply frequency is adjusted according to the desired speed, and the stator voltage is modified proportionally in order to keep the motor flux approximately constant.

#### III.4.7 PI Speed Controller Design:

In scalar speed control, the PI controller is commonly used to reduce the error between the reference speed and the measured motor speed. The controller acts on the control variable of the drive, usually the reference frequency or the slip frequency, in order to adjust the motor speed toward the desired value.

The speed error is defined as:

$$e(t) = \Omega_{\text{ref}}(t) - \Omega_r(t) \quad (\text{III. 51})$$

where  $\Omega_{\text{ref}}$  is the reference speed and  $\Omega_r$  is the measured rotor speed. The PI controller generates a control signal based on the proportional and integral actions. Its time-domain expression is:

$$u(t) = K_p e(t) + K_i \int e(t) dt \quad (\text{III. 52})$$

where  $K_p$  is the proportional gain and  $K_i$  is the integral gain. The proportional action improves the response speed by reacting directly to the instantaneous error. The integral action eliminates the steady-state error by accumulating the error over time.

In the Laplace domain, the PI controller can be written as:

$$C(s) = K_p + \frac{K_i}{s}$$

or equivalently:

$$C(s) = K_p \left( 1 + \frac{1}{T_i s} \right)$$

where  $T_i$  is the integral time constant, defined by:

$$T_i = \frac{K_p}{K_i}$$

The design of the PI controller can be related to the simplified mechanical model of the motor. From the mechanical equation, the speed dynamics can be approximated by a first-order system:

$$G_m(s) = \frac{1}{Js + f}$$

where  $J$  is the moment of inertia and  $f$  is the viscous friction coefficient. This simplified model shows that the mechanical behavior of the motor is mainly determined by inertia and friction.

A classical tuning method consists of choosing the PI zero in order to compensate the mechanical pole. The mechanical time constant is:

$$T_m = \frac{J}{f}$$

Therefore, the integral time constant of the PI controller can be selected as:

$$T_i = T_m = \frac{J}{f}$$

After selecting the integral time constant, the proportional gain can be adjusted according to the desired speed response. A higher value of  $K_p$  generally gives a faster response, but it may also increase overshoot and oscillations. The integral gain  $K_i$  is then calculated from:

$$K_i = \frac{K_p}{T_i}$$

In this work, the PI controller is used to improve the speed tracking of the three-phase induction motor. The controller compares the reference speed with the measured rotor speed and generates the control signal required by the scalar V/f control block. This structure allows the motor speed to follow the reference while keeping the control method simple and suitable for MATLAB/Simulink simulation.

### III.4.8 Vector Control

Vector control, also known as field-oriented control, consists of decoupling the flux and the electromagnetic torque of the induction motor, so that it can be controlled in a manner similar to a separately excited DC motor. In this method, known as rotor-flux-oriented control, the d-axis of the rotating reference frame is aligned with the rotor flux vector. This orientation forces the quadrature component of the rotor flux to zero and allows independent control of the flux through the direct stator current component and of the torque through the quadrature stator current component.

#### III.4.8.1 State Variables

The state variables used to describe the induction motor model in the rotor-flux-oriented reference frame are the direct and quadrature stator currents and the direct and quadrature rotor flux components:

$i_{sd}, i_{sq}, \Psi_{rd}, \Psi_{rq}$

#### III.4.8.2 Flux Orientation

The rotor-flux orientation condition is expressed as:

$$\Psi_{rd} = \Psi_r, \Psi_{rq} = 0 \quad (\text{III.53})$$

Under this condition, the rotor flux is entirely aligned with the d-axis, which simplifies the electromagnetic torque expression and allows flux and torque to be controlled separately.

The resulting state-space model can be written in the general form:

$$\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \quad (\text{III.54})$$

### III.4.8.3 Electrical Equations (Stator Currents)

The stator current dynamics in the rotor-flux-oriented frame are given by:

$$\frac{di_{sd}}{dt} = -\gamma \cdot i_{sd} + \omega_s \cdot i_{sq} + \left(\frac{K}{T_r}\right) \cdot \Psi_{rd} + \left(\frac{1}{\sigma L_s}\right) \cdot v_{sd} \quad (\text{III. 55})$$

$$\frac{di_{sq}}{dt} = -\omega_s \cdot i_{sd} - \gamma \cdot i_{sq} + K \cdot \omega \cdot \Psi_{rd} + \left(\frac{1}{\sigma L_s}\right) \cdot v_{sq} \quad (\text{III. 56})$$

### III.4.8.4 Magnetic Equations (Rotor Flux)

The rotor flux dynamics and the slip relation are given by:

$$\frac{d\Psi_{rd}}{dt} = \left(\frac{M}{T_r}\right) \cdot i_{sd} - \left(\frac{1}{T_r}\right) \cdot \Psi_{rd} \quad (\text{III. 57})$$

$$0 = \left(\frac{M}{T_r}\right) \cdot i_{sq} - (\omega_s - \omega) \cdot \Psi_{rd} \quad (\text{III. 58})$$

$$\omega_g = \omega_s - \omega = \frac{M}{T_r \cdot \Psi_{rd}} \cdot i_{sq} \quad (\text{III. 59})$$

with:

$$T_r = \frac{L_r}{R_r}, \quad \sigma = 1 - \frac{M^2}{L_s \cdot L_r}, \quad K = \frac{M}{\sigma \cdot L_s \cdot L_r}, \quad \gamma = \frac{R_s}{\sigma L_s} + \frac{R_r \cdot M^2}{\sigma L_s \cdot L_r^2} \quad (\text{III. 60})$$

The electromagnetic torque is expressed as:

$$C_{em} = \frac{P \cdot M}{L_r} \cdot (\Psi_{rd} \cdot i_{sq} - \Psi_{rq} \cdot i_{sd}) \quad (\text{III. 61})$$

Since the flux orientation imposes  $\Psi_{rq} = 0$ , the torque expression reduces to:

$$C_{em} = \frac{P \cdot M}{L_r} \cdot \Psi_{rd} \cdot i_{sq} \quad (\text{III. 62})$$

This relation shows that, once the rotor flux is fixed, the electromagnetic torque becomes directly proportional to the quadrature stator current  $i_{sq}$ , which is the main principle exploited by rotor-flux-oriented vector control: the rotor flux is regulated through  $i_{sd}$ , while the torque is regulated through  $i_{sq}$ , independently of each other.

### III.4.8.5 Transformation from abc to dq

The three-phase stator currents are first transformed into the two-phase stationary reference frame ( $\alpha$ - $\beta$ ) using the Clarke transformation:

$$\begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \end{bmatrix} = \frac{2}{3} \cdot \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \cdot \begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} \quad (\text{III. 63})$$

The stationary-frame currents are then transformed into the rotating reference frame (d-q), aligned with the rotor flux, using the Park transformation:

$$\begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} = \begin{bmatrix} \cos\theta_s & \sin\theta_s \\ -\sin\theta_s & \cos\theta_s \end{bmatrix} \cdot \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \end{bmatrix} \quad (\text{III. 64})$$

where  $\theta_s$  is the angular position of the rotor flux vector, obtained from the flux estimator.

#### III.4.8.6 Decoupling Technique

Because the d-axis and q-axis stator voltages are mutually coupled through the cross terms  $\omega_s \cdot \sigma L_s \cdot i_{sq}$  and  $\omega_s \cdot \sigma L_s \cdot i_{sd}$ , a decoupling technique is applied to compensate for these coupling terms and to allow the current loops to be designed independently. The final voltage references applied to the machine are obtained as:

$$V_{sd,final} = V_{sd,PI} - \omega_s \cdot \sigma L_s \cdot i_{sq} \quad (\text{III. 65})$$

$$V_{sq,final} = V_{sq,PI} + \omega_s \cdot \sigma L_s \cdot i_{sd} + \omega_s \cdot \Psi_r \cdot \left( \frac{M}{L_r} \right) \quad (\text{III. 66})$$

where  $v_{sd,PI}$  and  $v_{sq,PI}$  are the outputs of the direct and quadrature current PI controllers, respectively. This compensation removes the coupling effect between the two axes, so that each current loop behaves as a simple first-order system.

#### III.4.8.7 Current Loop

The transfer function of the stator current loop, after decoupling, is given by:

$$G_i(s) = \frac{1}{R_\sigma + \sigma L_s \cdot s} \quad (\text{III. 67})$$

with:

$$R_\sigma = R_s + \frac{R_r \cdot M^2}{L_r^2} \quad (\text{III. 68})$$

The current-loop PI controller is designed using pole-zero cancellation, with the following structure:

$$C_i(s) = K_{pc} \cdot \left( 1 + \frac{1}{T_{ic} \cdot s} \right), \quad T_{ic} = \frac{\sigma L_s}{R_\sigma} \quad (\text{III. 69})$$

The proportional and integral gains are calculated as:

$$K_{pc} = \frac{\sigma L_s}{\tau_{ic}} , \quad K_{ic} = \frac{K_{pc}}{T_{ic}} \quad (\text{III. 70})$$

where  $\tau_{ic}$  is the desired current-loop time constant, chosen as  $\tau_{ic} = 1$  ms.

#### III.4.8.8 Speed Loop

The mechanical equation of the motor is given by:

$$J \cdot \frac{d\Omega}{dt} = C_{em} - C_r - f \cdot \Omega \quad (\text{III. 71})$$

which leads to the simplified speed-loop transfer function:

$$G_{\Omega}(s) = \frac{K_t}{J \cdot s} \quad (\text{III. 72})$$

with  $K_t = (P \cdot M / L_r) \cdot \Psi_{rd}$ .

$$T_{i\Omega} = 10 \cdot T_{ic} , \quad K_{p\Omega} = \frac{2 \cdot \xi \cdot \omega_n \cdot J}{K_t} , \quad K_{i\Omega} = \frac{\omega_n^2 \cdot J}{K_t} \quad (\text{III. 73})$$

with  $\xi = 0.7$  and  $\omega_n = 3/tr$ , where  $tr = 10 \cdot \tau_{ic}$  is the desired speed-loop response time.

This cascaded structure, combining an inner decoupled current loop and an outer speed loop, constitutes the rotor-flux-oriented vector control strategy implemented for the three-phase induction motor in this thesis. The validity of this control structure is confirmed through the MATLAB/Simulink simulation results presented in section III.5, where the vector-controlled drive is tested under nominal conditions and under parameter variation, using both a classical PI controller and the proposed ANFIS controller.

### III.5 Speed Control of Three-Phase Induction Motor with Vector Control based on ( $\alpha$ - $\beta$ ) Axes:

Three-phase induction motors are the workhorses of industry due to their robustness, low cost, and minimal maintenance requirements. However, their speed control is more complex than DC motors because the rotor speed depends on both the supply frequency and the slip. Vector control (also known as field-oriented control) provides a solution by decoupling the torque and flux components, allowing independent control similar to a separately excited DC motor.

The vector control strategy based on the  $\alpha$ - $\beta$  reference frame uses the Clarke transformation to convert three-phase quantities (a-b-c) into a two-phase stationary orthogonal reference frame ( $\alpha$ - $\beta$ ). This transformation is the first step in the vector control implementation and allows the representation of the three-phase system in a simpler two-dimensional plane.

### III.5.1 Block Simulink and Integration ANFIS Speed Controller

The ANFIS speed controller provides the torque reference  $C_{em}$  which is used to calculate the  $\alpha$ - $\beta$  current references. The relationship between electromagnetic torque and stator currents in the  $\alpha$ - $\beta$  frame is given by:

$$C_{em} = p \cdot (\Psi_{ds} \cdot I_{qs} - \Psi_{qs} \cdot I_{ds}) \quad (III.74)$$

where:

- $p$ : number of pole pairs.
- $\Psi_{ds}, \Psi_{qs}$ : rotor flux components in  $\alpha$ - $\beta$  frame.

The flux estimator computes the rotor flux components from the measured stator currents and voltages using the machine model. The ANFIS controller adapts its output based on the speed error and the estimated flux position to achieve optimal torque production.

In this section, an Adaptive Neuro-Fuzzy Inference System (ANFIS) is implemented for the vector control of a three-phase induction motor. The ANFIS controller combines the learning capability of neural networks with the reasoning capability of fuzzy logic, providing improved performance compared to classical PI controllers, especially under parameter variations.

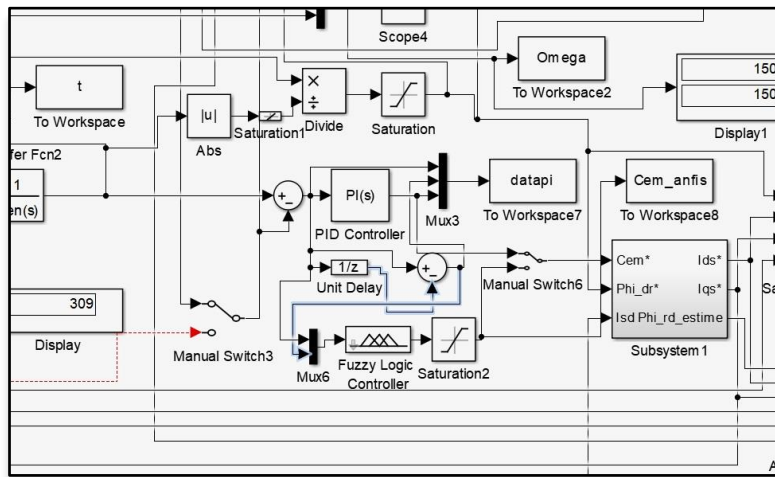


**Fig. III.27:** Simulink Block Diagram – ANFIS and PI Controller Integration

#### III.5.1.1 Interpretation of the Simulink Block Diagram:

This figure shows the complete Simulink implementation of the vector control system with dual controller architecture:

- **Speed Error Calculation:** The difference between reference speed and measured speed ( $\Omega$ ) is computed. This error feeds both the PID and ANFIS controllers.
- **ANFIS Controller:** The Fuzzy Logic Controller block implements the trained ANFIS system. It receives the speed error and its derivative (via Unit Delay) and produces an adaptive torque reference.
- **Manual Switch:** Allows selection between PID and ANFIS control modes for comparative testing. When switched to ANFIS, the system uses the neuro-fuzzy controller output.

**-Current reference calculation** (subsystem 1): the torque reference ( $C_{em}$ ) is converted to  $\alpha$ - $\beta$  current reference ( $I_{qs}$ ,  $I_{ds}$ ) using the rotor flux estimate ( $\Psi_{dr}$ ).

**-PWM Generation:** The current references are processed through a PWM modulator (Mux6  $\rightarrow$  Fuzzy Logic Controller  $\rightarrow$  Saturation) to generate switching signals for the inverter.

**-Data Logging:** Multiple "To Workspace" blocks (data\_pi, Cem\_anfis, etc.) record simulation data for post-processing and comparison between PID and ANFIS performance.

-The flux estimator ( $I_{ds}$ ,  $\Psi_{dr}$ ) estimate continuously computes the rotor flux position and magnitude from measured stator currents and voltages, providing the necessary information for coordinate transformation and torque calculation.

### III.5.2 ANFIS Structure and Membership Functions:

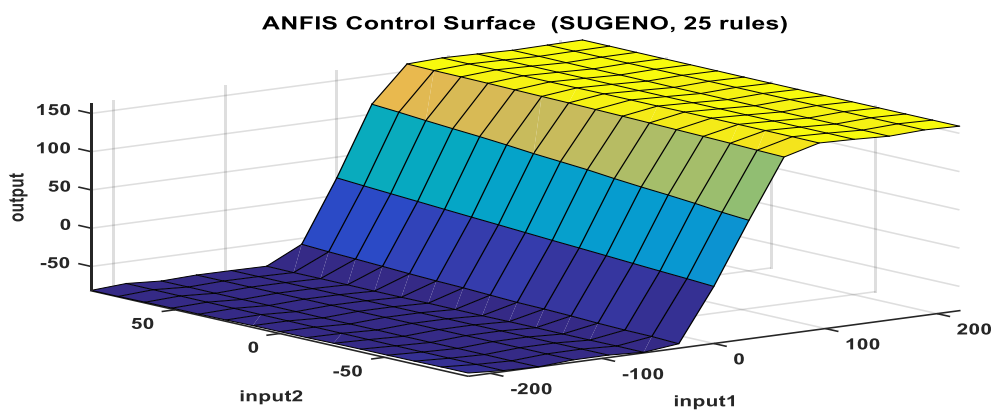


Fig. III.28:( ANFIS Surface)

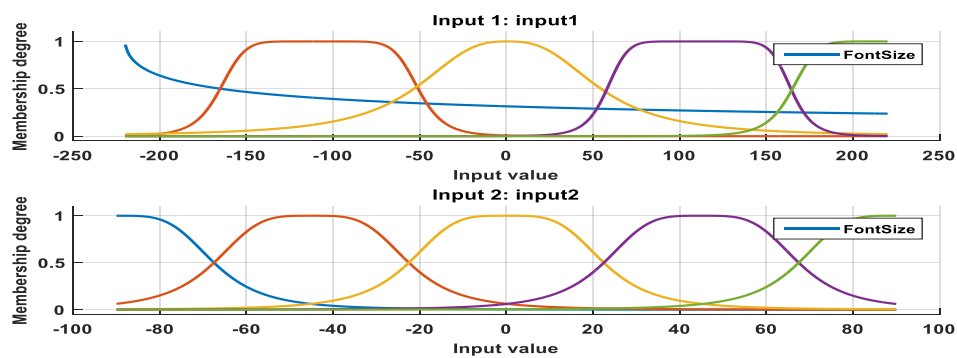


Fig. III.29:( ANFIS input membership functions)

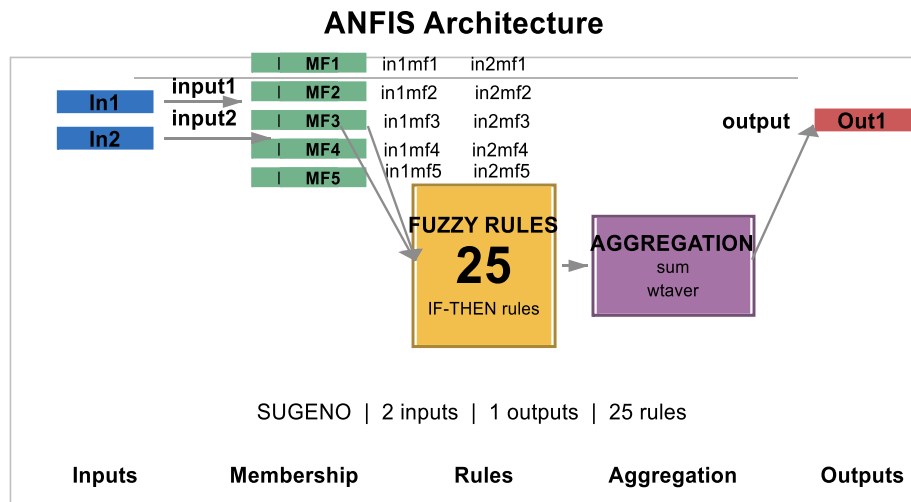


Fig. III.30:( ANFIS structure)

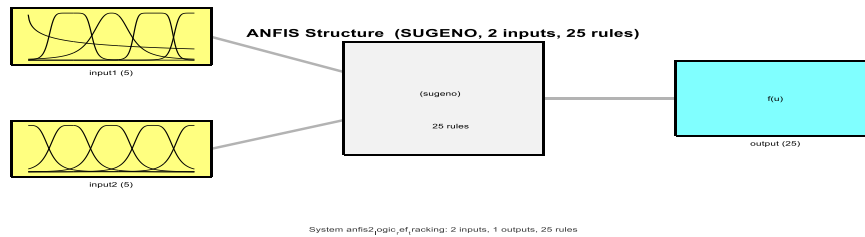


Fig. III.31:( ANFIS structure)

### III.5.2.1 Interpretation of the ANFIS Properties:

This composite figure displays four essential views of the ANFIS controller:

- **Membership Function Editor:** Shows five Gaussian membership functions (in1mf1 to in1mf5) covering the input range [-220, 220]. These functions represent the fuzzy partitioning of the speed error input. The smooth overlapping transitions ensure continuous control action without abrupt changes.
- **Fuzzy Logic Designer:** Presents the overall Sugeno-type FIS architecture with 2 inputs, 25 rules, and 1 output. The configuration uses product implication, max aggregation, and weighted average defuzzification (wtaver), which is optimal for generating smooth control surfaces suitable for gradient-based learning.
- **ANFIS Model Structure:** Illustrates the five-layer neural network architecture. Layer 1 contains membership functions, Layer 2 computes rule firing strengths (25 "AND" nodes shown in blue), Layer 3 normalizes activations, Layer 4 calculates linear consequences, and Layer 5 performs defuzzification. This hybrid structure enables simultaneous optimization of membership function parameters and rule consequents.
- **Surface Viewer:** Displays the 3D input-output mapping. The characteristic shape—a flat region near zero error for steady-state precision and steep slopes for large errors provides both fine control near the setpoint and aggressive response during transients.

### III.5.3 Simulation Results and Interpretation :

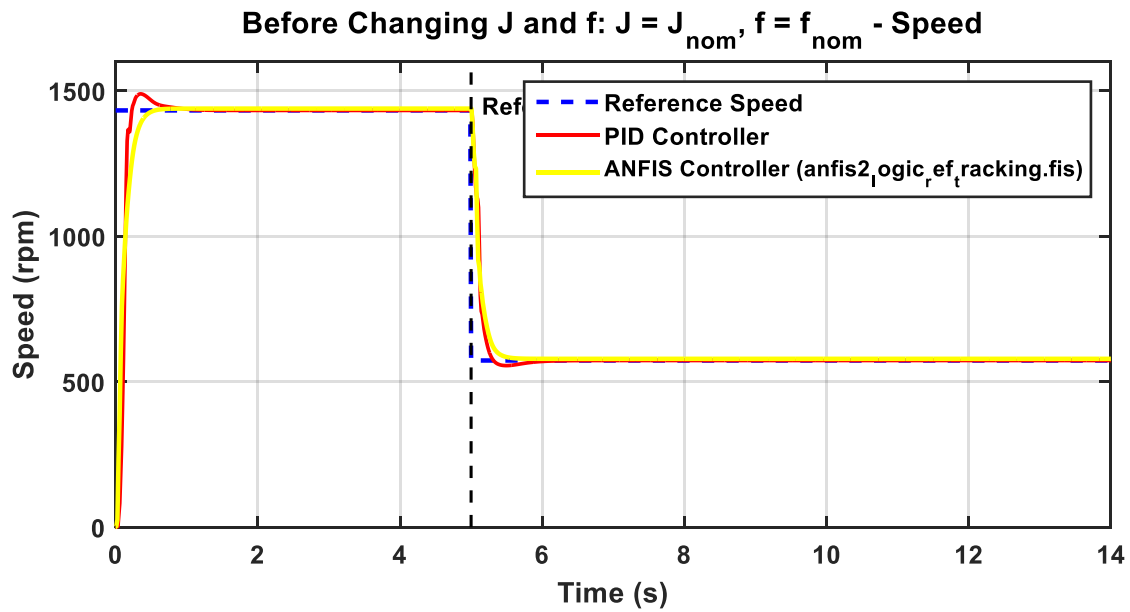


**Fig. III.32:** Speed responses – Nominal conditions ( $J = J_{nom}$ ,  $f = f_{nom}$ )

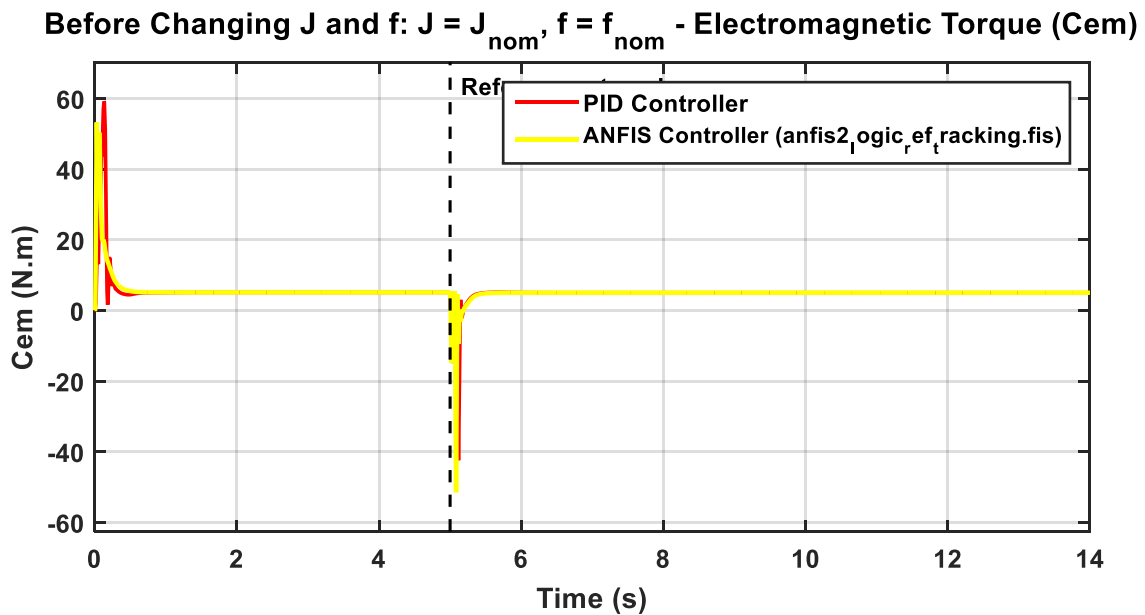


**Fig. III.33:** Torque responses – Nominal conditions ( $J = J_{nom}$ ,  $f = f_{nom}$ )

#### III.5.3.1 Interpretation of Speed and Electromagnetic Torque Responses for Variable Speed Drive with Vector Control and ANFIS Controller:

1. The two figures present the performance of an electric motor supplied by a variable speed drive using the vector control technique under nominal conditions, where  $J = J_{nom}$  and  $f = f_{nom}$ .
2. The variable speed drive is responsible for controlling the motor speed by adjusting the voltage and frequency applied to the machine.
3. With vector control, the drive separates the control of flux and electromagnetic torque, which gives a fast and accurate dynamic response.

4. In the speed response, the reference speed is initially about 1450 rpm and then decreases suddenly to about 580 rpm at  $t = 5$  s.
5. Both PID and ANFIS controllers allow the variable speed drive to track the reference speed correctly.
6. However, the PID controller produces a noticeable overshoot at start-up, where the speed reaches nearly 1500 rpm.
7. This overshoot means that the variable speed drive applies a strong control action, which can increase electrical and mechanical stress.
8. The ANFIS controller improves the behavior of the variable speed drive by providing a smoother speed response with reduced overshoot.
9. At  $t = 5$  s, when the speed reference decreases, ANFIS helps the drive reach the new speed more stably and with fewer oscillations.
10. In the torque response, the variable speed drive generates a high electromagnetic torque at start-up to accelerate the motor.
11. With PID control, the torque peak is sharper, which may cause torque ripples and stress on the motor shaft.
12. With ANFIS control, the electromagnetic torque becomes smoother during both acceleration and braking phases.
13. This smoother torque response reduces vibrations, current peaks, and mechanical constraints in the motor-drive system.
14. Therefore, ANFIS has a positive effect on the vector-controlled variable speed drive by improving stability, robustness, speed tracking, and overall dynamic performance.

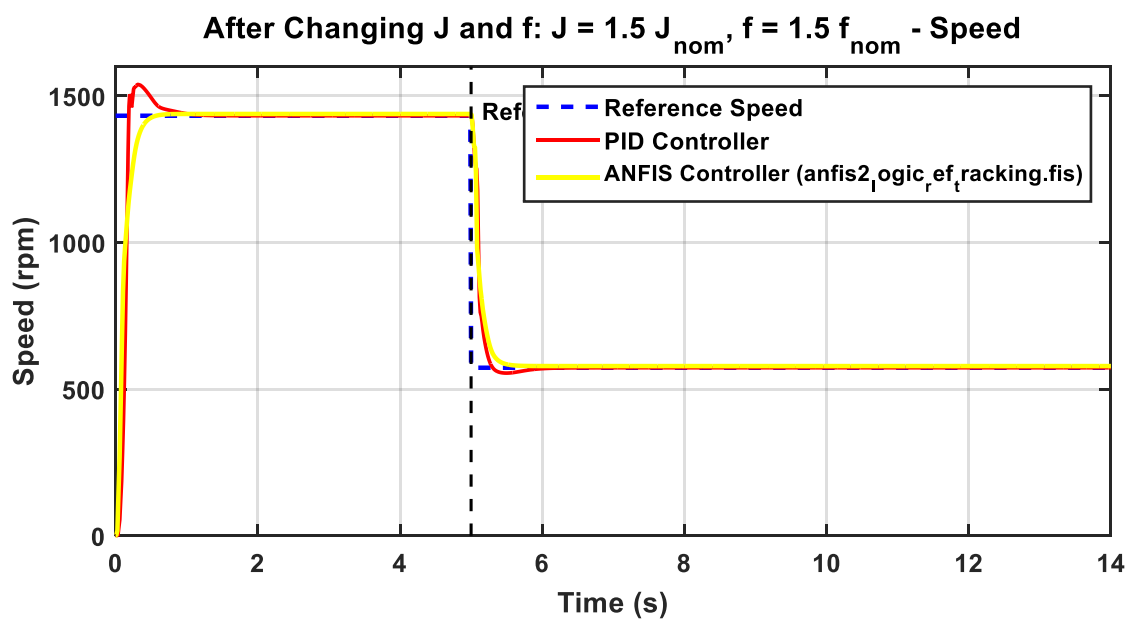


**Fig. III.34:** Speed responses – Perturbed conditions ( $J = 1.5 J_{nom}$ ,  $f = 1.5 f_{nom}$ )

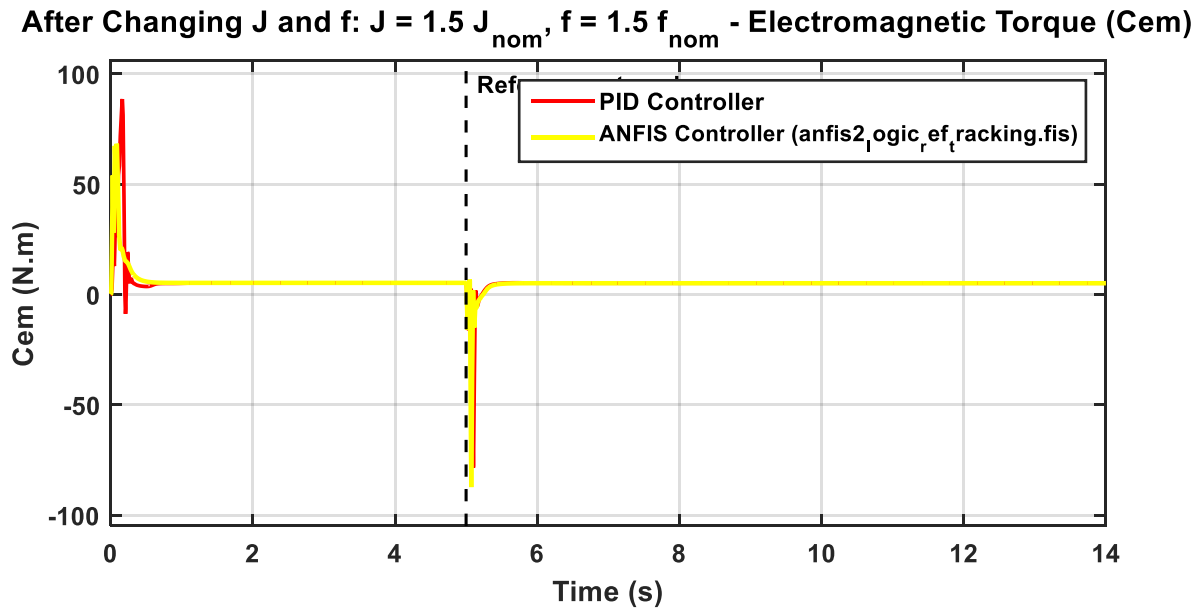


Fig. III.35: torque responses –Perturbed conditions ( $J = 1.5 J_{nom}$ ,  $f = 1.5 f_{nom}$ )

### III.5.3.2 Interpretation After Changing J and f for Variable Speed Drive with Vector Control and ANFIS Controller:

1. The two figures show the behavior of an electric motor supplied by a variable speed drive using vector control after changing the mechanical parameters, where  $J = 1.5 J_{nom}$  and  $f = 1.5 f_{nom}$ .
2. This increase in inertia and friction makes the motor-drive system more demanding because the motor needs more torque to accelerate and decelerate.
3. The variable speed drive controls the motor speed by adjusting the voltage and frequency applied to the machine through the inverter.
4. With the vector control technique, the drive separately controls the flux and torque components, which improves the dynamic response even under parameter variations.
5. In the speed response, the reference speed is initially about 1450 rpm and then decreases suddenly to about 580 rpm at  $t = 5$  s.
6. Both PID and ANFIS controllers allow the variable speed drive to follow the reference speed despite the increase of J and f.
7. However, the PID controller presents a visible overshoot at start-up, where the speed exceeds the reference and reaches nearly 1550 rpm.
8. This overshoot shows that the variable speed drive with PID applies a strong control action, which may increase current peaks and mechanical stress.
9. The ANFIS controller improves the behavior of the drive by giving a smoother speed response with reduced overshoot and better damping.
10. At  $t = 5$  s, when the reference speed decreases, ANFIS helps the variable speed drive reach the new speed more smoothly and with fewer oscillations.

11. In the electromagnetic torque response, a high starting torque is produced to overcome the higher inertia and accelerate the motor.
12. With PID control, the torque peak is sharper and reaches a high value, close to 90 N.m, which can produce torque ripples and stress on the motor shaft.
13. With ANFIS control, the torque variation is smoother during acceleration and braking, which reduces vibrations, current peaks, and mechanical constraints.
14. Therefore, ANFIS has a positive effect on the vector-controlled variable speed drive by improving robustness, stability, speed tracking, and torque smoothness under increased J and f conditions.

### **III.5.3.3 Effect of Motor Resistances and J/f Variations**

#### **➤ Comparative analysis of both scenarios:**

In the studied motor-drive system, the electric motor is supplied by a variable speed drive using vector control. The variation of J and f indirectly reflects the influence of parameter changes, including the effects related to stator and rotor resistances, on the dynamic behavior of the drive. Increasing the inertia J slows down the mechanical response of the variable speed drive system, while increasing the friction f adds additional damping and modifies the closed-loop system poles. Therefore, the drive must generate a more suitable torque and current response to maintain good speed tracking under these variations.

#### **➤ ANFIS Controller Robustness:**

The ANFIS controller improves the robustness of the vector-controlled variable speed drive because the neuro-fuzzy system was trained on a dataset covering different operating conditions. This allows it to learn the nonlinearities and uncertainties of the motor-drive system. Thanks to this adaptation capability, ANFIS can maintain stable speed regulation and smoother electromagnetic torque even when J, f, or motor resistance effects vary, without requiring gain retuning.

#### **➤ PID Controller Limitations:**

The classical PID controller is generally designed for a nominal operating point of the variable speed drive. When the system parameters deviate from their nominal values, the PID performance can degrade because the fixed gains  $K_p$  and  $K_i$  cannot fully compensate for changes in the mechanical time constant and motor-drive dynamics. As a result, the drive may present higher overshoot, sharper torque peaks, and more oscillations compared with the ANFIS-based control.

#### **➤ Positive effect on the variable speed drive:**

By integrating ANFIS into the speed control loop, the variable speed drive becomes more adaptive and more efficient under parameter variations. The drive can better regulate the voltage, frequency, and torque-producing current required by the motor. Consequently, the ANFIS-based variable speed drive provides improved stability, reduced mechanical stress, smoother torque response, and better tracking of the reference speed than the conventional PID controller.

### **III.5.4 Conclusion:**

The study of d-q vector control applied to an induction machine supplied by a Variable Frequency Drive (VFD) and controlled by an ANFIS controller has highlighted the significant advantages of this

intelligent approach compared to classical PI control. In this drive system, the VFD plays a central role because it adjusts the voltage and frequency applied to the induction motor through the inverter, allowing accurate speed regulation under different operating conditions.

The electromagnetic torque relation as a function of rotor flux and stator currents in the  $\alpha$ - $\beta$  reference frame:

$$C_{em} = p \cdot (\Psi_{ds} \cdot I_{qs} - \Psi_{qs} \cdot I_{ds})$$

constitutes the theoretical foundation of vector decoupling, enabling independent control of torque and flux. This decoupling allows the VFD to control the induction machine with high dynamic performance, similar to a separately excited DC motor.

The ANFIS architecture, combining the learning capabilities of neural networks and fuzzy reasoning, enables automatic adaptation to parameter variations and operating condition changes. This adaptive capability improves the behavior of the VFD by generating more appropriate control actions without requiring manual retuning of controller parameters. Such a characteristic is particularly valuable in industrial applications, where machine parameters and load conditions may vary over time.

In conclusion, ANFIS-based  $\alpha$ - $\beta$  vector control integrated into a Variable Frequency Drive represents a high-performance and robust solution for speed control of induction machines. It offers an excellent compromise between implementation complexity and superior dynamic and static performance compared to the classical PI regulator. By improving speed tracking, reducing overshoot, limiting torque oscillations, and enhancing robustness against parameter variations, ANFIS allows the VFD to operate more efficiently, smoothly, and reliably.

The comparative simulation between the classical PI controller and the ANFIS controller applied to the indirect vector control of an induction machine supplied by a VFD shows that ANFIS achieves the best overall performance. Indeed, the settling time is reduced by 28%, from 0.703 s with the PI controller to 0.506 s with the ANFIS controller, which represents a major advantage for VFD-based systems requiring a fast dynamic response. Furthermore, ANFIS exhibits superior robustness against variations in mechanical parameters, especially with a 50% increase in the inertia ( $J$ ) and viscous friction coefficient ( $f$ ), while maintaining accurate speed tracking and stable operation with virtually no performance degradation. The residual steady-state error of ANFIS, approximately 1% of the reference value, remains acceptable and can be further reduced by enriching the training data of the neuro-fuzzy system.

Furthermore, the robustness of the proposed ANFIS controller was evaluated under significant electrical parameter variations in the DC motor drive. The armature resistance ( $R_a$ ) and the field resistance ( $R_f$ ) were intentionally increased to assess the controller performance under parameter uncertainties. The simulation results showed that the conventional PI controller suffered from a noticeable degradation in speed tracking accuracy and exhibited a significant steady-state error. In contrast, the ANFIS controller maintained accurate reference tracking and preserved the desired speed with negligible steady-state error. These results demonstrate the superior capability of ANFIS to

compensate for electrical parameter variations and maintain stable operation without requiring controller retuning.

Therefore, the proposed ANFIS control strategy demonstrates strong robustness against both electrical parameter variations in the DC motor ( $(R_a, R_f)$ ) and mechanical parameter variations in the induction motor ( $(J, f)$ ). The intelligent learning capability of ANFIS enables the control system to maintain high performance, stability, and tracking accuracy despite significant changes in machine parameters and operating conditions.

Therefore, the integration of ANFIS into the control loop of a vector-controlled VFD significantly improves the global performance of the induction motor drive. The obtained results demonstrate that ANFIS not only provides faster dynamic response and better reference tracking than the conventional PI controller, but also maintains high performance under significant electrical and mechanical parameter variations. This confirms that the proposed ANFIS-based VFD is a promising solution for industrial variable-speed applications requiring fast response, robustness, stability, adaptability, and high control accuracy.

### III.5.5 Hardware Architecture:

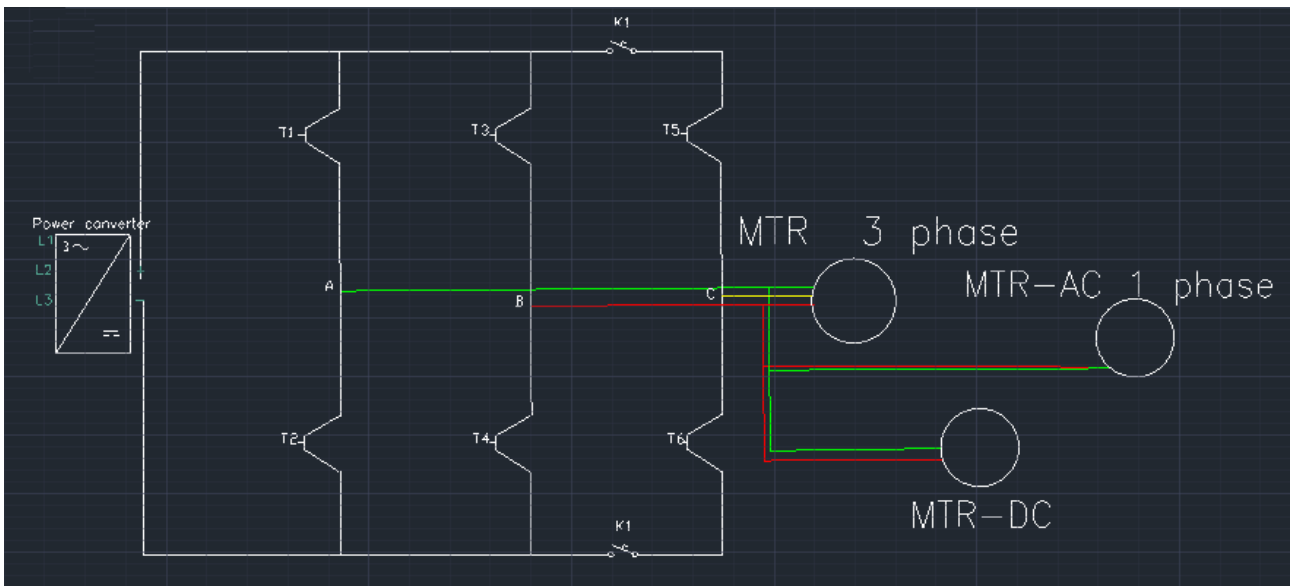


**Fig III.36:** Hardware Architecture: Inverter Bridge for Three-Phase and Single-Phase Motor Control.

**Interpretation:** Using a single inverter bridge to drive both types of motors (three-phase and single-phase) demonstrates the flexibility of VFD control and its ability to adapt to different mechanical loads according to the requirements of the industrial application.

### III.5.6 Practical Implementation Results :

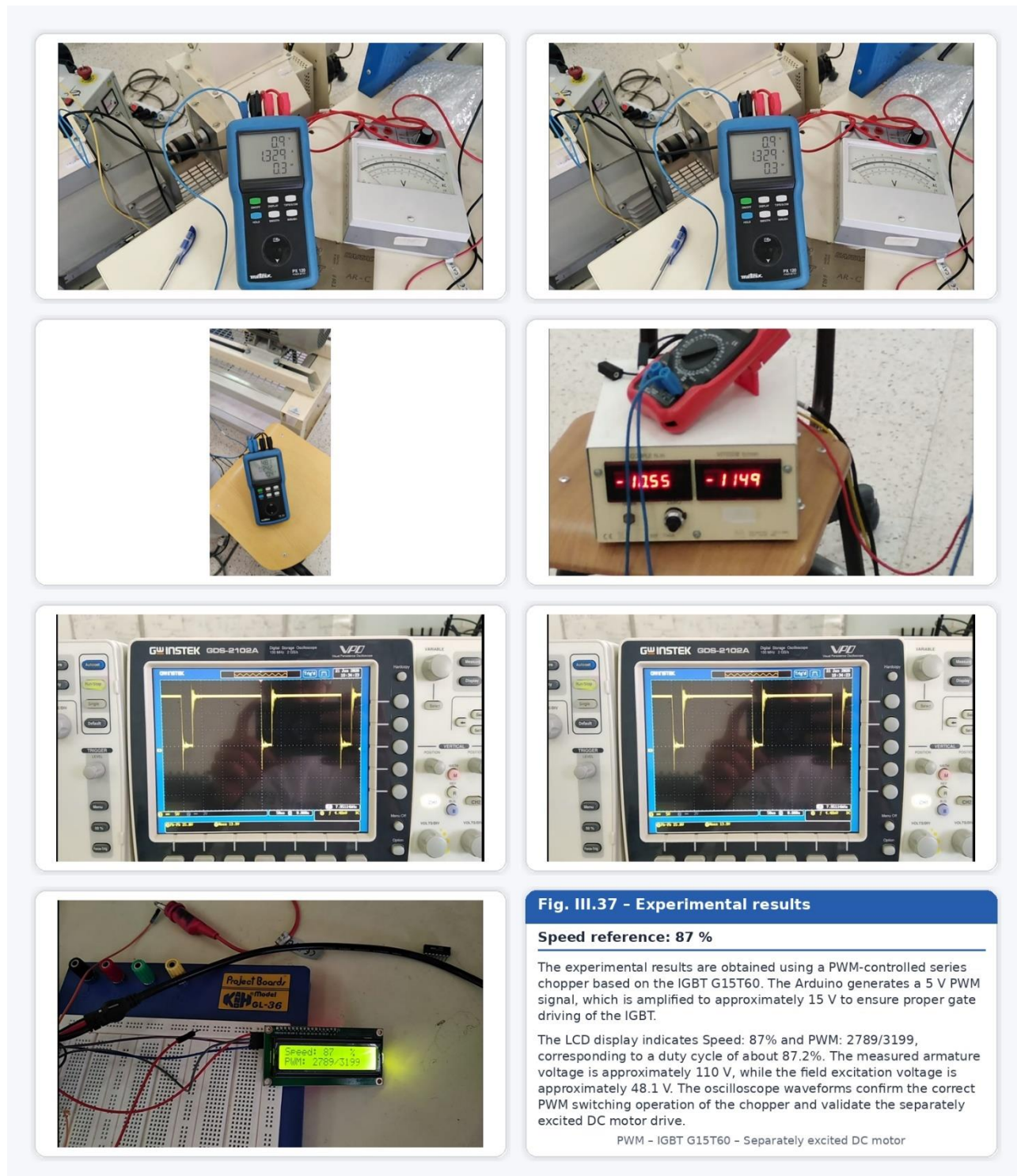


**Fig. III.37** results for a speed reference of 87% using a PWM-controlled

Fig. III.37 For a speed reference of 87%, the motor speed reaches approximately 1149 rpm, with an armature current of about 1.32 A. The PWM value is 2789/3199, corresponding to a duty cycle of about 87.2%. These results show that a higher PWM duty cycle produces a higher motor speed, confirming the correct operation of the speed control system.

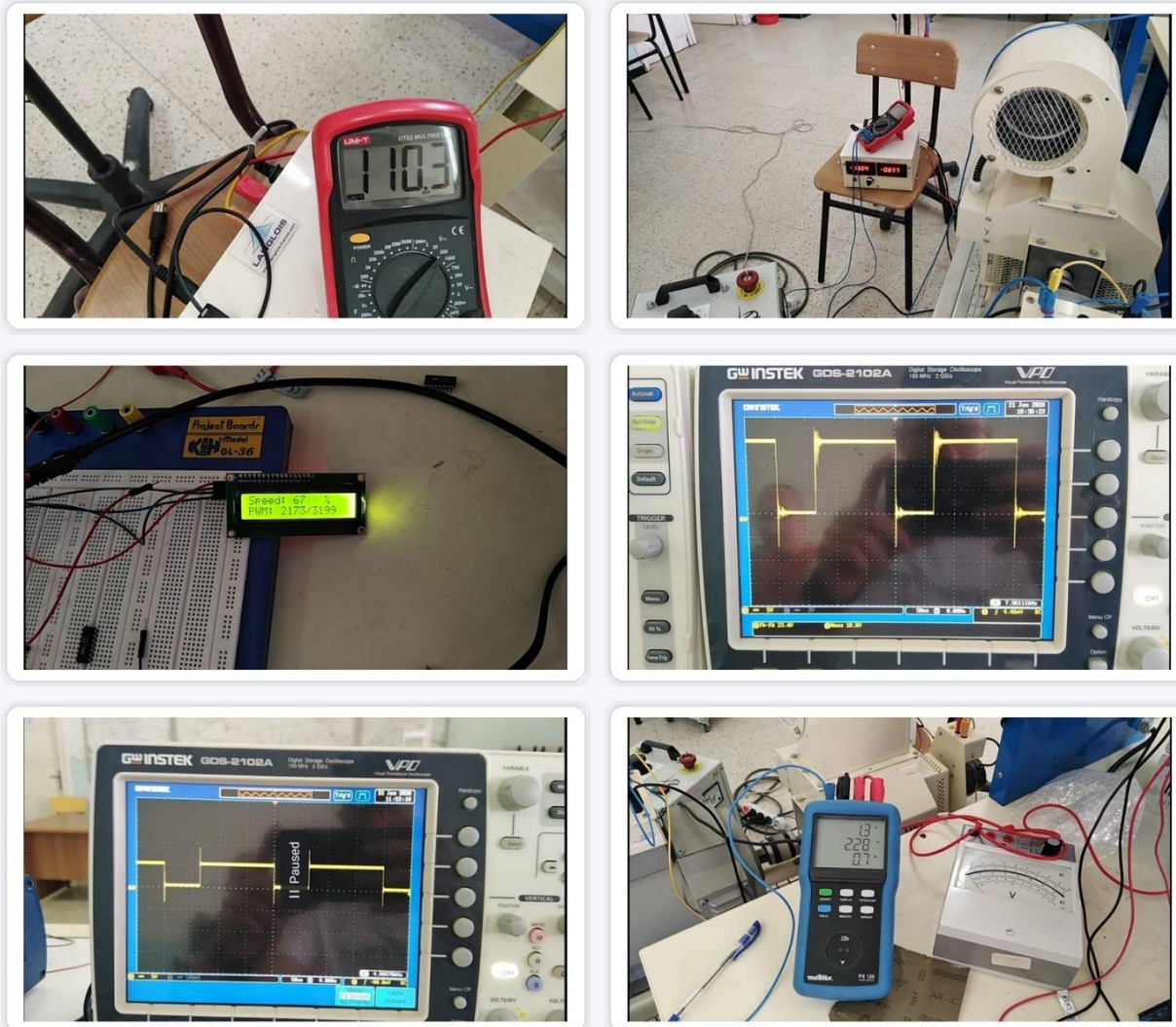


**Fig. III.38 - Experimental results**

**Speed reference: 67 %**

The experimental results are obtained using a PWM-controlled series chopper based on the IGBT G15T60. The Arduino generates a 5 V PWM signal, which is amplified to approximately 15 V to ensure proper gate driving of the IGBT.

The LCD display indicates Speed: 67% and PWM: 2173/3199, corresponding to a duty cycle of about 67.9%. The measured armature voltage is approximately 110.3 V. The oscilloscope waveforms confirm the PWM switching operation of the chopper and validate the correct operation of the speed drive for the separately excited DC motor.

PWM - IGBT G15T60 - Separately excited DC motor

**Fig. III.38** Interpretation of Experimental Results (Duty Cycle = 67%, Armature Voltage = 110 V)

Fig. III.38 For a speed reference of 67%, the motor speed decreases to approximately 837 rpm, with an armature current of about 2.28 A. The PWM value is 2173/3199, corresponding to a duty cycle of about 67.9%. The decrease in speed compared with the 87% test confirms the effect of reducing the PWM duty cycle on the motor drive.

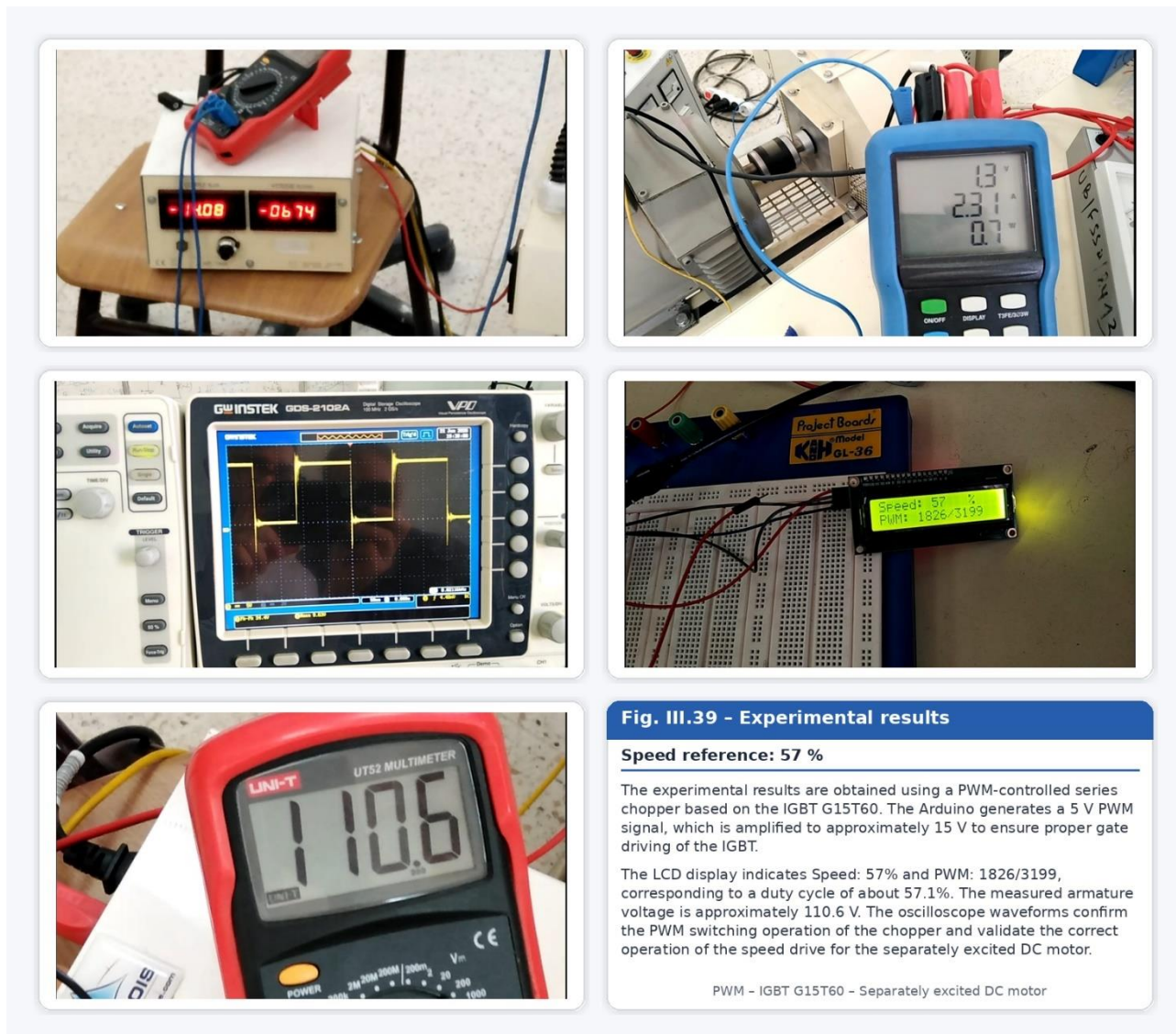


**Fig. III.39** Interpretation of Experimental Results (Duty Cycle = 57%, Armature Voltage = 110 V)

Fig. III.39 For a speed reference of 57%, the motor speed is approximately 674 rpm, with an armature current of about 2.31 A. The PWM value is 1826/3199, corresponding to a duty cycle of about 57.1%. These results confirm that lowering the PWM duty cycle reduces the motor speed, while the current remains sufficient to maintain motor operation under load.

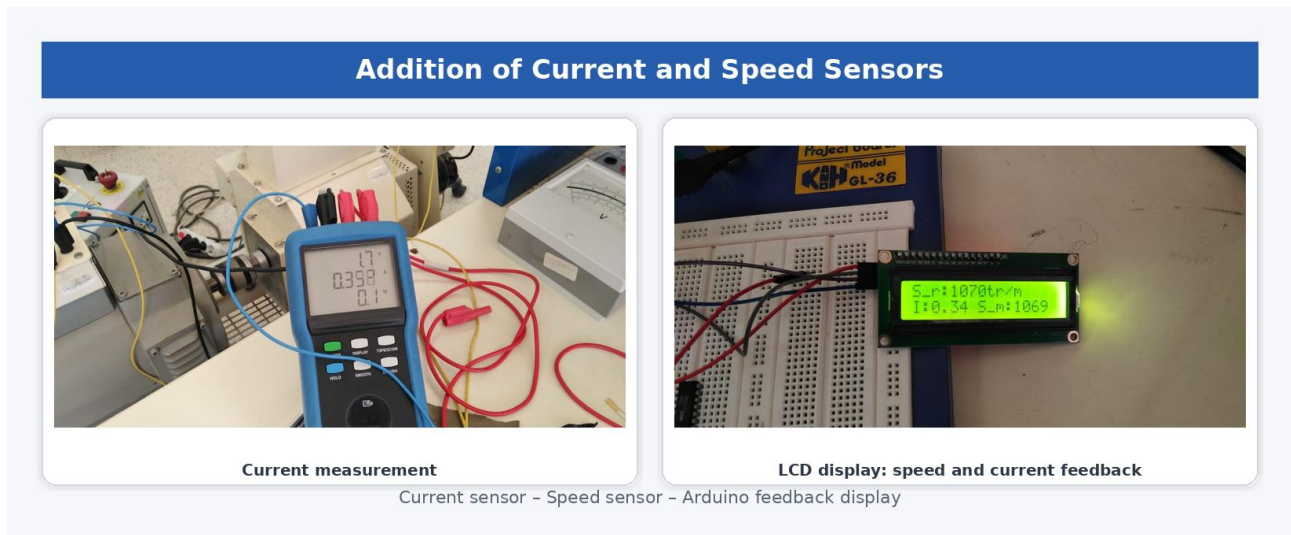


**Fig. III.40** sensors allow real-time feedback of the DC motor operating conditions

Interpretation: The two added sensors allow real-time feedback of the DC motor operating conditions. The current sensor measures the motor armature current, while the speed sensor provides the rotational speed in rpm. The LCD display shows a speed of approximately 1070 rpm, a measured current of about 0.34 A, and a reference speed of 1069 rpm.

The close agreement between the measured speed and the reference speed confirms the correct operation of the speed sensor and the control system. The measured current indicates the electrical load of the motor during operation. These results show that the integration of the current and speed sensors improves monitoring and allows better supervision of the separately excited DC motor drive.

# General Conclusion

This thesis addressed the study and design of an intelligent speed controller for electric motors, with the aim of improving the dynamic response and robustness of motor drives compared with classical PI/PID regulation. To achieve this objective, three representative machines were considered: the separately excited DC motor, the single-phase induction motor, and the three-phase induction motor. Each machine was mathematically modeled from its electrical, electromechanical, and mechanical equations, and its associated power converter and control structure were developed and simulated in MATLAB/Simulink under different operating conditions. This progressive approach, from the fundamentals of electrical machines and drive architecture to detailed modeling and simulation, allowed the proposed control strategy to be validated on machines of increasing complexity.

For each machine, a conventional PI or PID controller was first implemented as a reference, after which an ANFIS-based intelligent controller was designed and integrated into the same control loop, either as a direct speed regulator or as an adaptive correction added to the PI structure. In every case, the neuro-fuzzy controller was able to learn the nonlinear behavior of the motor-converter system and to generate more appropriate control actions without requiring manual retuning of the controller gains. This confirms that the ANFIS approach can be applied consistently across DC and AC drives, regardless of the differences in their mathematical models and control strategies.

The simulation results obtained throughout the thesis show a clear and consistent improvement brought by the ANFIS controller over the classical approach. For the three-phase induction motor under indirect vector control, the settling time was reduced by about 28%, from 0.703 s with the PI controller to 0.506 s with the ANFIS controller, while the residual steady-state error remained close to 1% of the reference value. When the mechanical parameters were increased by 50% (inertia  $J$  and viscous friction coefficient  $f$ ), the ANFIS controller maintained accurate speed tracking with almost no performance degradation, whereas the PID controller exhibited a larger overshoot and a sharper, more oscillatory torque response. Similar robustness was observed for the DC motor drive, where increasing the armature and field resistances caused the PI controller to develop a noticeable steady-state error, while the ANFIS controller preserved the desired speed with negligible deviation. For the single-phase induction motor operating under scalar V/f control, adding an ANFIS-based correction to the fixed PI controller reduced the overshoot caused by motor resistance variation and produced a smoother, more stable speed response. Taken together, these results demonstrate that the proposed intelligent controller consistently improves speed tracking

accuracy, reduces overshoot and torque ripple, and increases robustness against both electrical and mechanical parameter variations, without sacrificing the simplicity of the underlying control structure.

Beyond the simulation study, the practical relevance of the proposed approach was also illustrated through a hardware implementation. A common inverter bridge architecture was shown to be adaptable to both single-phase and three-phase motor control, reflecting the flexibility required in real industrial drives. In addition, a DC motor prototype was built and tested at different PWM duty cycles, and the measured speed and current values obtained through the integrated sensors closely matched the reference values used in the simulations, which supports the validity of the modeling and control approach developed earlier in the thesis. This link between simulation and hardware testing strengthens the practical significance of the work, since intelligent controllers are often evaluated only in simulation, without physical confirmation of their behavior.

From an industrial point of view, the ability of the ANFIS controller to adapt automatically to parameter uncertainties and load variations is particularly valuable, as machine parameters and operating conditions rarely remain constant in real installations. Reducing the need for manual gain retuning, while at the same time improving speed regulation and reducing mechanical stress on the motor shaft, directly addresses common limitations of classical PI/PID-based variable speed drives used in conveyor systems, pumps, fans, and other variable-load industrial applications.

Several directions can be considered to extend this work. The experimental validation, currently limited to the DC motor, could be extended to the single-phase and three-phase induction motors using a suitable microcontroller or DSP platform capable of running the ANFIS algorithm in real time. The training data used for the neuro-fuzzy system could also be enriched with additional operating scenarios in order to further reduce the residual steady-state error observed in the vector-controlled induction motor drive. Finally, it would be of interest to compare the ANFIS controller with other intelligent or adaptive strategies, such as sliding mode control or reinforcement learning-based controllers, and to evaluate the proposed approach under real industrial load profiles, in order to confirm its performance beyond the simulated and laboratory conditions considered in this thesis.

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