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By:

Assem BENYAHIA

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Presented to the jury composed of:

Dr. Hemza MEKKI

Pr University of M'Sila

President.

Dr. Riyadh ROUABHI

MCA University of M'Sila

Supervisor.

Dr. Ali DJERIOUI

Pr University of M'Sila

A. Supervisor.

Dr. Abdelrahim ZEMMIT

MCA University of M'Sila

Examiner.

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بِسْمِ اللَّهِ الرَّحْمَنِ الرَّحِيمِ

الشكر

اشكر الله تعالى بشكل خاص على الإرادة والصحة والصبر التي قدمها لي خلال

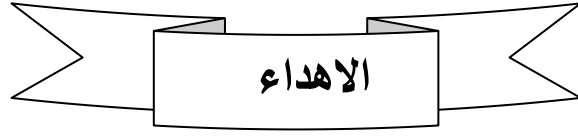
كل هذه السنوات الطويلة.

اعرب عن عميق امتناني لوالدي الأعزاء على تشجيعهم ودعمهم والتضحيات التي تحملوها.

اود أيضاً أن اعرب عن خالص شكري لمشرفي

علي جريو و رياض روابحي أولَّ وقبل كل شيء الذين لم يتوقفا عن تقديم نصائح قيمة لي.

شكرا ايضا لكل من ساهموا بشكل مباشر أو غير مباشر في تحقيق هذه الطَّروحة.



مرت قاطرة البحث بكثير من العوائق، ومع ذلك حاولت أن أتخطاها بثبات بفضل من الله ومنه

إلى أبوي وأخوتي وأصدقائي، فقد كانوا بمثابة العضد والسند في سبيل استكمال البحث

ولّ ينبغي أن أنسى أساتذتي ممن كان لهم الدور الأكبر في مساندتي ومدي بالمعلومات القيمة

أهدي لكم بحث تخرجي

داعياً المولى عز وجل أن يطيل في أعماركم ويرزقكم بالخيرات.

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Symbols

DFIG : Double Fed Induction Generator

PWM : Pulse-width modulation

f_r : Fréquence des grandeurs rotoriques

a, b, c : Indices correspondants aux trois phases

$a, b, c \max, \min$: Valeur maximale (minimale)

s, r : Indices du stator, du rotor

R_s, R_r : Résistances statorique et rotorique

L_s, L_r : Inductances propre statorique et rotorique

σ : Coefficient de dispersion

T_r, T_s : Constantes de temps rotorique et statorique

M_{sr} : Inductance mutuelle stator- rotor

M_{rs} : Inductance mutuelle rotor - stator

M_s : Inductance mutuelle entre les enroulements du stator

M_r : Inductance mutuelle entre les enroulements du rotor

Φ_s, Φ_r : Respectivement flux statorique et rotorique

Φ_{sd}, Φ_{sq} : Les flux statoriques direct et en quadrature

Φ_{rd}, Φ_{rq} : Les flux rotoriques direct et en quadrature

I_s, I_r : Les courants de phase statorique et rotorique

I_{sd}, I_{sq} : Les courants statoriques directe et en quadrature dans le repère de PARK

I_{rd}, I_{rq} : Les courants rotoriques directe et en quadrature dans le repère de PARK

V_s, V_r : Tensions simples respectivement statorique et rotorique

V_{sd}, V_{sq} : Les tensions statoriques directe et en quadrature

V_{rd}, V_{rq} : Les tensions rotoriques directe et en quadrature

θ_s, θ_r : Angles électriques statorique et rotorique

ω_s, ω_r : Pulsations statorique et rotorique

Symbols

P : Nombre de paires de poles

g : Glissement

p : Operateur de LAPLACE

cem : Couple électromagnétique

Cr : Couple résistant

Ps, Qs : Puissances active et réactive statorique

Pr, Qr : Puissances active et réactive rotorique

ρ : Masse volumique du vent

f : Le coefficient des frottements visqueux totaux de l'éolienne

V : La vitesse moyenne du vent

J : L'inertie totale de l'éolienne

G : Le rapport du multiplicateur de vitesse

λ : La vitesse spécifique de la turbine

β : L'angle de calage des pales

λ_{opt} : La vitesse spécifique optimale de la turbine

$\Omega T, \Omega g$: La vitesse de la turbine et la vitesse du générateur

CT, Cg : Le couple de la turbine et le couple électromagnétique

CT_{opt} : Le couple optimal de la turbine

Cg_{opt} : Le couple optimal du générateur

PT : La puissance convertie par la turbine

RT : Le rayon des pales de la turbine

$Cp(\lambda, \beta)$: Le coefficient de puissance

Uc : La tension du bus continu

Ic : Le courant dans le condensateur du bus continu

c : Le condensateur du bus continu

Symbols

s : Surface de glissement

v_1 : Gain La surface de glissement de puissance active

v_2 : Gain La surface de glissement de puissance réactive

∂ : Dérivée partielle

y : Variable de sortie

y_r : Trajectoire désirée

t : Variable de temps

e : Erreur de poursuite

$\dot{x}$: La dérivée de la variable x par rapport au temps x refou

x_{ref} : La référence de la variable

$\| \cdot \|$: Une norme sur $\mathbb{R}^n$

$\mathbb{R}^n$: Espace vectoriel de dimension n construit sur le corps des réels

$\mathbb{R}^+$: Ensemble des réels positifs ou nuls

$\mathbb{R}$: Ensemble des nombres réels

MPPT : Maximum Power Point Tracking

V. C : Vector Control Tit

S. M. C : Sliding Mode Control

As concerns over global warming have spread worldwide, environmental protection has emerged as a key issue. The concept of sustainable development has become integral to all industrial and economic sectors, influencing new projects. In the electricity sector, countries around the world are seeking to reduce their dependence on fossil fuels, which are characterised by high greenhouse gas emissions and volatile prices. Operators in this sector are striving to diversify their energy mix by using clean, renewable energy sources such as geothermal, biomass, solar and wind power [1, 2, 3 and 4].

Wind power has become one of the most popular forms of renewable energy in recent years. Over the last decade, it has grown in popularity thanks to its many advantages. Rather than replacing conventional resources, wind power has emerged as a complementary energy source [5, 6 and 7].

Many of today's wind turbines are equipped with a doubly-fed induction generator (DFIG), which enables them to operate over a wide range. The stator circuit connects directly to the grid, while the rotor circuit connects to it via power converters. The stator circuit connects directly to the grid, while the rotor circuit connects to it via power converters. Since the power exchanged between the rotor and the grid is low, the cost of the converters is lower than that of a variable-speed wind turbine powered by the stator. This is the main reason why this generator is used for high-power generation. A second reason is the ability to adjust the generator voltage at the point of connection [8, 9, and 10].

Traditional linear control techniques often struggle to manage the complex behaviour of WECS effectively, particularly during periods of fluctuating wind speed or when operating in transient modes. Consequently, nonlinear control methods such as sliding mode control, model predictive control and fuzzy logic have gained significant attention due to their ability to improve the efficiency, stability and reliability of these systems. However, before such advanced control algorithms can be implemented in real-world wind turbine systems, they must first be validated in an environment that realistically replicates the behaviour of the actual hardware. This is where Processor-in-the-Loop (PIL) validation comes in [11, 12, and 13].

Wind energy conversion systems (WECS) are complex, nonlinear structures susceptible to external disturbances such as wind turbulence and mechanical vibrations. To ensure efficient and reliable energy extraction, the control system of a WECS must be capable of handling these dynamic and uncertain conditions. This thesis focuses on a WECS based on a double-fed

asynchronous machine (DFIG), considering the system configuration, converter type and control technique adopted using the loop method.

Our work aims to develop and implement a control algorithm that ensures good performance and the production of high-quality electrical energy by a WECS. This control algorithm is known as sliding mode control based on estimated gains. This approach involves selecting an appropriate sliding surface and determining a convergence condition to formulate a control law that can independently regulate active and reactive power generation.

The first chapter aims to develop a comprehensive mathematical and simulation model of the WECS. This involves modelling the system's key components, such as the wind turbine, gearbox (if applicable), generator, and power electronics interface. The dynamic behaviour and interrelations of these components under varying wind conditions will then be determined. This will lay the groundwork for performance analysis, control design and system optimisation in subsequent chapters.

The second chapter outlines the design and implementation of a sliding mode control strategy for regulating active and reactive power in the WECS. This ensures that the system performs robustly under a variety of operating conditions. The third chapter provides an overview of the digital signal processing (DSP) card and explains how it works. We examine its mechanism and processing structure before introducing the Processor-in-the-Loop (PIL) approach. We then load the code onto the DSP card to run simulations. Finally, we analyse the outcomes in detail.

I.1 Introduction

In order to simulate the behavior of our wind energy conversion system under different operating conditions and understand its mechanisms, we must model the system individually.

- We selected a dual-fed asynchronous machine to serve as the generator for our wind energy system.
- A two-level inverter serves as the machine-side converter, controlling the flow of generated energy.
- The grid-side converter, a two-level rectifier, will improve the grid-side power factor and regulate the DC bus.
- The wind turbine and its control system maximize power collection and limit it during low and high wind speeds.

I.2 Modelling the double-fed asynchronous machine

In order to simplify and linearise the DFIG model, we will create a mathematical model of it in the Park reference frame, which is linked to the rotating field. This will enable us to control the DFIG under different operating conditions and accurately depict its behaviour [14–18].

I.2.1 Simplifying assumptions

Due to its complex structure and geometry, DFIG is difficult to model. Therefore, certain assumptions must be made to create a simple model that closely resembles reality. The assumptions are as follows:

- ✚ Firstly, it is assumed that the machine is symmetrical. Secondly, it is assumed that the air gap thickness is uniform.
- ✚ It is also assumed that the magnetic circuit is not saturated, meaning the fluxes can be expressed as linear functions of the currents.
- ✚ Hysteresis losses and eddy currents are also negligible.
- ✚ The current density is uniform across the cross-section of the elementary conductors and the skin and notch effects are negligible.
- ✚ The resistances and self-inductance coefficients are constant and the windings are temperature-independent. The mutual inductance coefficients depend on the position of their magnetic axes.
- ✚ The magnetomotive forces created by the stator or rotor are sinusoidal.

I.2.2 Mathematical equations for DFIG in the (abc) frame of reference

As previously mentioned, the DFIG is a conventional machine with an accessible rotor identical to the stator. Therefore, its model is the same as an asynchronous cage machine's, except for the non-zero rotor voltages [19].

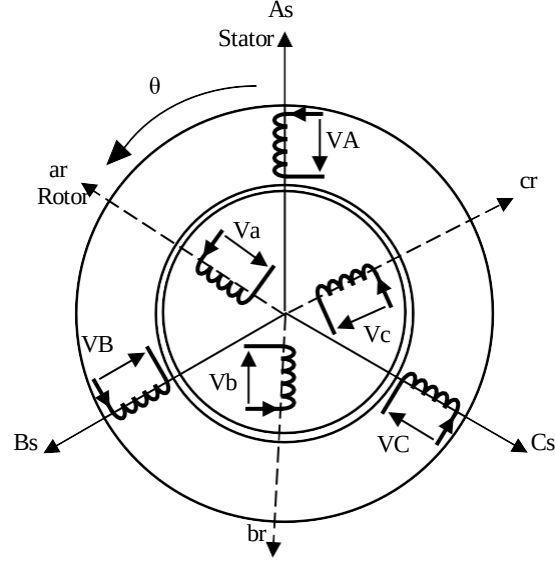


Figure I-1: This is an electrical diagram of the DFIG.

❖ Electrical Equations :

The equations that govern the DFIG electrical voltages are:

$$\begin{cases} [V_s] = [R_s] \cdot [I_s] + \frac{d}{dt} [\varphi_s] \\ [V_r] = [R_r] \cdot [I_r] + \frac{d}{dt} [\varphi_r] \end{cases} \quad (I.1)$$

with :

$$\begin{aligned} [V_s] &= \begin{bmatrix} V_{as} \\ V_{bs} \\ V_{cs} \end{bmatrix}; & [I_s] &= \begin{bmatrix} I_{as} \\ I_{bs} \\ I_{cs} \end{bmatrix}; & [\varphi_s] &= \begin{bmatrix} \varphi_{as} \\ \varphi_{bs} \\ \varphi_{cs} \end{bmatrix}; & [R_s] &= \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix} \\ [V_r] &= \begin{bmatrix} V_{ar} \\ V_{br} \\ V_{cr} \end{bmatrix}; & [I_r] &= \begin{bmatrix} I_{ar} \\ I_{br} \\ I_{cr} \end{bmatrix}; & [\varphi_r] &= \begin{bmatrix} \varphi_{ar} \\ \varphi_{br} \\ \varphi_{cr} \end{bmatrix}; & [R_r] &= \begin{bmatrix} R_r & 0 & 0 \\ 0 & R_r & 0 \\ 0 & 0 & R_r \end{bmatrix} \end{aligned}$$

Where: R_s and R_r : These are the resistances of the stator and rotor windings.

V_s, V_r : The stator and the rotor are simple voltage sources, respectively.

I_s, I_r : The phase currents of the stator and rotor are, respectively.

φ_s, φ_r : These are the stator and rotor fluxes, respectively.

❖ Magnetic equations

The magnetic flux equations that govern the DFIG are as follows:

$$\begin{cases} [\varphi_s] = [L_{ss}] \cdot [I_s] + [M_{sr}] \cdot [I_r] \\ [\varphi_r] = [L_{rr}] \cdot [I_r] + [M_{rs}] \cdot [I_s] \end{cases} \quad (I.2)$$

$$\text{With:} \quad [L_{ss}] = \begin{bmatrix} l_s & M_s & M_s \\ M_s & l_s & M_s \\ M_s & M_s & l_s \end{bmatrix} \quad ; \quad [L_{rr}] = \begin{bmatrix} l_r & M_r & M_r \\ M_r & l_r & M_r \\ M_r & M_r & l_r \end{bmatrix}$$

Where:

L_{ss}, L_{rr} : These are the self-inductances of the stator and rotor phases, respectively.

M_s, M_r : These are the mutual inductances between the stator and rotor phases.

$[M_{sr}] = [M_{rs}]^T$: The mutual inductance matrix, also known as the stator-rotor coupling matrix, is represented by the following formula (I.3):

$$[M_{sr}] = [M_{rs}]^T = M_{max} \begin{bmatrix} \cos(\theta) & \cos(\theta + \frac{2\pi}{3}) & \cos(\theta - \frac{2\pi}{3}) \\ \cos(\theta - \frac{2\pi}{3}) & \cos(\theta) & \cos(\theta + \frac{2\pi}{3}) \\ \cos(\theta + \frac{2\pi}{3}) & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta) \end{bmatrix} \quad (I.3)$$

M_{max} : is the maximum mutual inductance between a stator phase and a rotor phase.

Substituting (I.2) into (I.1) yields the following system of equations:

$$\begin{cases} [V_s] = [R_s] \cdot [I_s] + \frac{d}{dt} [[L_{ss}] \cdot [I_s] + [M_{sr}] \cdot [I_r]] \\ [V_r] = [R_r] \cdot [I_r] + \frac{d}{dt} [[L_{rr}] \cdot [I_r] + [M_{rs}] \cdot [I_s]] \end{cases} \quad (I.4)$$

❖ Mechanic Equation

The mechanical equation of the machine is given as follows:

$$C_{em} = C_r + f_r \cdot \Omega + J \cdot \frac{d\Omega}{d\theta} \quad (I.5)$$

With:

- C_{em} - The machine's electromagnetic torque.
- C_r - resistive torque;
- f_r - The coefficient of viscous friction of the DFIG.
- Ω - is the rotational speed of the DFIG axis.
- J - is the inertia of the rotating parts.
- The expression for the electromagnetic torque is given by the following relationship:

$$C_{em} = p \cdot [I]_s^t \frac{d}{d\theta} [[M]_{sr} \cdot [I]_r] \quad (I.6)$$

Where p - The number of DFIG pole pairs.

The DFIG model is represented by the following differential equations in the (a, b, c) reference frame: (I.4); (I.5); (I.6).

I.2.3 Park transformation

This transformation is a mathematical tool that simplifies three-phase electrical machine equations (frame (a, b, c)) into a two-phase system (frame (d, q)). The Park transformation matrix ($P(\theta)$) conserves power during this transition and is defined as follows:

$$[P(\theta)] = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos(\theta) & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta - \frac{4\pi}{3}) \\ -\sin(\theta) & -\sin(\theta - \frac{2\pi}{3}) & -\sin(\theta - \frac{4\pi}{3}) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (I.7)$$

θ : This is the angle between the reference frames of the three-phase system undergoing transformation and the Park reference frame.

I.2.4 DFIG uses mathematical equations in the (dq) reference frame.

We use the Park transformation equation to convert three-phase equations into two-phase equations, and vice versa. The equation is as follows:

$$[x_{dq0}] = [P(\theta)] \cdot [x_{abc}] \Rightarrow [x_{abc}] = [P(\theta)]^{-1} \cdot [x_{dq0}]$$

With :

$$[P(\theta_s)] \cdot \frac{d}{dt} [[P(\theta_s)]^{-1}] = \frac{d\theta_s}{dt} \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}; [P(\theta_r)] \cdot \frac{d}{dt} [[P(\theta_r)]^{-1}] = \frac{d\theta_r}{dt} \begin{bmatrix} 0 & -1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix}$$

These equations were obtained through the transformation process.

❖ Electrical equations

$$\begin{cases}
 V_{sd} = R_s I_{sd} + \frac{d\varphi_{sd}}{dt} - w_s \varphi_{sq} \\
 V_{sq} = R_s I_{sq} + \frac{d\varphi_{sq}}{dt} + w_s \varphi_{sd} \\
 V_{rd} = R_r I_{rd} + \frac{d\varphi_{rd}}{dt} - w_r \varphi_{rq} \\
 V_{rq} = R_r I_{rq} + \frac{d\varphi_{rq}}{dt} + w_r \varphi_{rd}
 \end{cases} \quad (I.8)$$

V_{sd} , V_{sq} , V_{rd} And V_{rq} : represent the forward and quadrature stator and rotor voltages, respectively.

I_{sd} , I_{sq} , I_{rd} And I_{rq} : represent the direct and quadrature stator and rotor currents, respectively.

φ_{sd} , φ_{sq} , φ_{rd} And φ_{rq} : represent the direct and quadrature stator and rotor fluxes of the system, respectively.

w_s , w_r : represent stator and rotor pulsations, respectively.

θ_s, θ_r : The electrical angles formed by the stator and rotor axes with the direct axis of the two-phase Park

❖ Magnetic equations

$$\begin{cases}
 \varphi_{sd} = L_s I_{sd} + M I_{rd} \\
 \varphi_{sq} = L_s I_{sq} + M I_{rq} \\
 \varphi_{rd} = L_r I_{rd} + M I_{sd} \\
 \varphi_{rq} = L_r I_{rq} + M I_{sq}
 \end{cases} \quad (I.9)$$

With :

L_r , L_s : The machine's inductances are the stator and rotor inductances.

M : is the mutual inductance.

❖ Mechanical equation

$$C_{em} = p \frac{M}{L_s} (I_{rd} \varphi_{sq} - I_{rq} \varphi_{sd}) \quad (I.10)$$

$$\begin{cases}
 P = V_{sd} I_{sd} + V_{sq} I_{sq} \\
 Q = V_{sq} I_{sd} - V_{sd} I_{sq}
 \end{cases} \quad (I.11)$$

With p - The number of DFIG pole pairs.

I.2.5 Choice of reference frame

In our study, we opted for a rotating reference frame fixed in relation to the rotating field due to the advantages this offers. This frame of reference is used at synchronous pulse speed. In this frame of reference, the relative components of the machine remain constant, which is why it is widely used for controlling asynchronous machines:

$$\frac{d\theta_s}{dt} = \frac{d\theta_r}{dt} + \frac{d\theta}{dt} = w_s = w_r + w.$$

I.2.6 The state representation of our machine model is as follows:

The model can be expressed as a matrix [20–24]:

$$[X] = [A].[X] + [B].[U] \quad (I.12)$$

$$\text{With :} \quad [X] = [\varphi_{sd} \quad \varphi_{sq} \quad i_{rd} \quad i_{rq}]^t ; \quad [U] = [v_{sd} \quad v_{sq} \quad v_{rd} \quad v_{rq}]^t ;$$

$$[A] = \begin{bmatrix} \frac{-1}{T_s} & w_s & \frac{M}{T_s} & 0 \\ -w_s & \frac{-1}{T_s} & 0 & \frac{M}{T_s} \\ \alpha & -\beta(w_s - w_r) & -\delta & w_r \\ \beta(w_s - w_r) & \alpha & -w_r & -\delta \end{bmatrix} ; \quad [B] = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ \frac{-M}{\sigma L_r L_s} & 0 & \frac{1}{\sigma L_r} & 0 \\ 0 & \frac{-M}{\sigma L_r L_s} & 0 & \frac{1}{\sigma L_r} \end{bmatrix}$$

α , β And δ are constants defined as follows :

$$\alpha = \frac{M}{\sigma T_s L_s L_r} ; \quad \beta = \frac{M}{\sigma L_s L_r} ; \quad \delta = \frac{1}{\sigma} \left(\frac{1}{T_r} + \frac{M^2}{T_s L_r L_s} \right).$$

To facilitate simulation, the matrix [A] is decomposed as follows:

$$[A] = [A_1] + w_s.[A_2] + w_r.[A_3] \quad (I.13)$$

$$[A_1] = \begin{bmatrix} \frac{-1}{T_s} & 0 & \frac{M}{T_s} & 0 \\ 0 & \frac{-1}{T_s} & 0 & \frac{M}{T_s} \\ \alpha & 0 & -\delta & 0 \\ 0 & \alpha & 0 & -\delta \end{bmatrix}, [A_2] = \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & -\beta & 0 & 0 \\ \beta & 0 & 0 & 0 \end{bmatrix}, [A_3] = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & \beta & 0 & 1 \\ -\beta & 0 & -1 & 0 \end{bmatrix}$$

I.3 Modelling and control of the machine-side converter

Variable-speed wind energy conversion systems based on a DFIG require the use of a static converter (inverter) which supplies the rotor of the machine. The main objectives of this converter are: to undulate the DC bus voltage in order to supply it to the rotor winding, to enable

the application of commands to control the power generated by the stator of this machine and to inject it into the grid.

I.3.1 Modelling the machine-side converter

This converter is a two-level, bidirectional power inverter equipped with power switches that control the opening and closing processes. These switches can be MOSFET transistors associated with head-to-tail diodes [25–27].

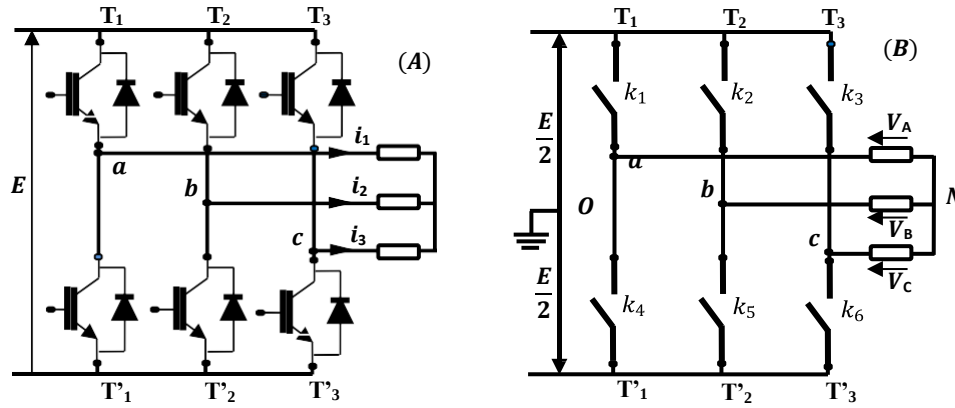


Figure I-2: Shows the converter's structure on the machine side.

The following equations represent the mathematical model of the three-phase two-level PWM inverter [25–27].

$$\begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = \frac{E}{6} \cdot \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_1 \\ S_2 \\ S_3 \end{bmatrix} \quad (\text{I. 14})$$

I.3.2 Pulse width modulation control

Pulse width modulation (PWM) control involves comparing the three reference sinusoidal voltages required by the inverter with a triangular signal known as the 'carrier signal'. Control pulses (i.e. the opening and closing of the switches) are then generated based on this comparison. This control is characterised by two parameters:

- The modulation index (m) is defined as follows : $m = \frac{f_p}{f}$
- f_p : The carrier frequency ratio.
- f : The frequency of the reference voltage.

The modulation rate (r) is defined as follows: $r = \frac{V_{ref}}{U_p} \quad (\text{I. 15})$

V_{ref} : The amplitude of the reference voltage.

U_P : The amplitude of the carrier voltage.

I.4 Modelling and control of the grid-side converter

This converter is currently used to produce stable and variable DC voltages in response to load demands from an AC network, thereby ensuring bidirectional power flow. It also reduces harmonic distortion, providing an input current that is almost sinusoidal in form with a power factor of unity.

I.4.1 Modelling the converter on the grid side

In order to control and regulate the converter in its various operating modes, it must be modelled. The following expressions represent the converter's mathematical model [28,29]:

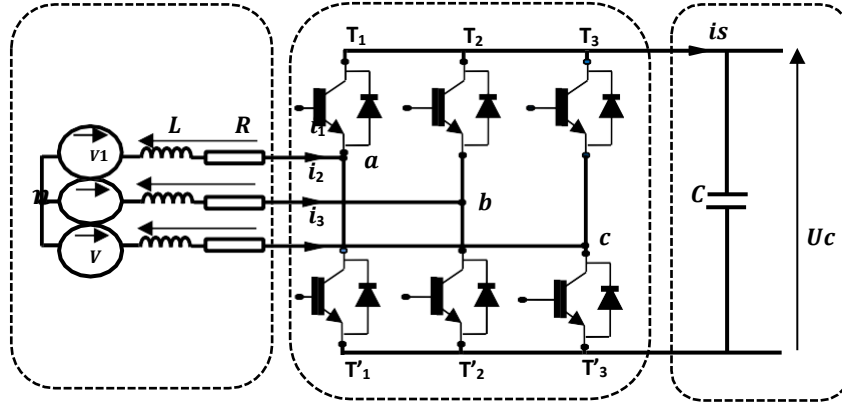


Figure I-3: Network-side converter structure.

The line voltage equations are given by :

$$\begin{cases} V_1 = R * i_1 + L \frac{di_1}{dt} + V_{an} \\ V_2 = R * i_2 + L \frac{di_2}{dt} + V_{bn} \\ V_3 = R * i_3 + L \frac{di_3}{dt} + V_{cn} \end{cases} \quad (I.16)$$

The line currents are given by the following formula :

$$\frac{d}{dt} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} = \begin{bmatrix} -\frac{R}{L} & 0 & 0 \\ 0 & -\frac{R}{L} & 0 \\ 0 & 0 & -\frac{R}{L} \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} + \frac{1}{L} \begin{bmatrix} V_1 - V_{an} \\ V_2 - V_{bn} \\ V_3 - V_{cn} \end{bmatrix} \quad (I.17)$$

The network-side converter has the same structure as the machine-side converter modelled above.

$$\begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = \frac{U_c}{3} \cdot \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_1 \\ S_2 \\ S_3 \end{bmatrix} \quad (\text{I. 18})$$

$$i_s = \begin{bmatrix} S_1 & S_2 & S_3 \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} \quad (\text{I. 19})$$

$$\frac{dU_c}{dt} = \frac{1}{C} (i_s - i_L) \quad (\text{I. 20})$$

We can use equations I.17, I.18, I.19 and I.20 to create a block diagram of our converter [28,29].

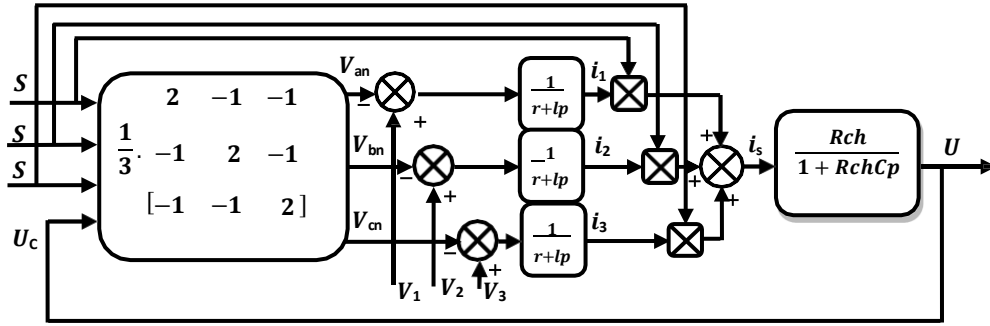


Figure I-4: This block diagram shows how the grid-side converter model operates.

I.4.2 Voltage orientation control of the mains-side converter

Voltage orientation control is one of the most effective solutions used in recent years to address harmonic issues and improve the power factor. This method uses two internal regulators in parallel to control phase currents synchronised with phase voltages (the fastest method), thereby ensuring that reactive power remains at zero while maintaining a unity power factor. In series with an external regulator, these internal regulators control the voltage conversion to the desired value. The equations governing the line voltages in the (d, q) reference frame are as follows:

$$\begin{cases} V_{pd} = V_d - R i_d - L \frac{di_d}{dt} + L \omega i_q \\ V_{pq} = V_q - R i_q - L \frac{di_q}{dt} - L \omega i_d \end{cases} \quad (\text{I. 21})$$

V_{pd} Et V_{pq} : are the voltage components at the converter input.

V_d Et V_q : are the components of the mains voltages.

i_d Et i_q : are the components of the mains currents.

ω : is the network pulsation.

$$\begin{cases} P = \frac{3}{2} \cdot [V_d I_d + V_q I_q] \\ Q = \frac{3}{2} \cdot [V_q I_d - V_d I_q] \end{cases} \quad (I.22)$$

The following matrix form can be used to represent this system:

$$\begin{bmatrix} P \\ Q \end{bmatrix} = \frac{3}{2} \cdot \begin{bmatrix} V_d & V_q \\ V_q & -V_d \end{bmatrix} \begin{bmatrix} I_d \\ I_q \end{bmatrix} \quad (I.23)$$

We pose :

$$P_{ref} = U_{cmes} I_{red_ref} \quad \text{And} \quad Q_{ref} = 0.$$

- The reference voltage (U_{Cref}) is then compared with the voltage (U_{Cmes}) measured across the capacitor.
- The PI corrector maintains the output voltage at the desired constant value and generates the reference current (I_{Cref}).
- To obtain the active power and charge the capacitor, the rectified current (I_{red_ref}) is multiplied by the measured voltage (U_{Cmes}).
- The reference reactive power must be kept at zero.
- The reference currents are obtained from the measured voltages and reference powers.
- We then compare the reference currents with the measured currents.
- Current comparison errors are regulated by PI regulators that generate reference voltages, which are then compared with the measured voltages.
- Voltage comparison errors generate voltages that are compared with the filter voltages.
- Current comparison errors generate control voltages that switch the six rectifier switches on and off [28,29]

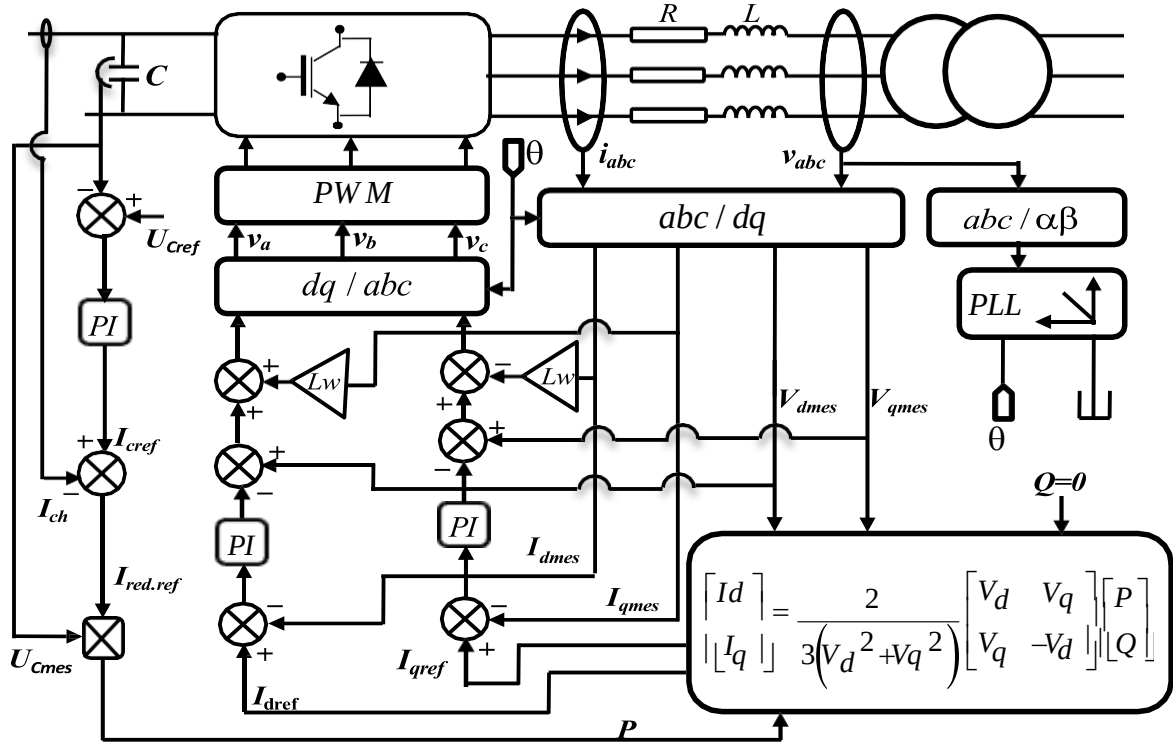


Figure I-5: Block diagram showing voltage orientation control of the grid-side converter.

I.4.3 Numerical simulation results

The parameters used in the rectifier simulation can be found in the appendix.

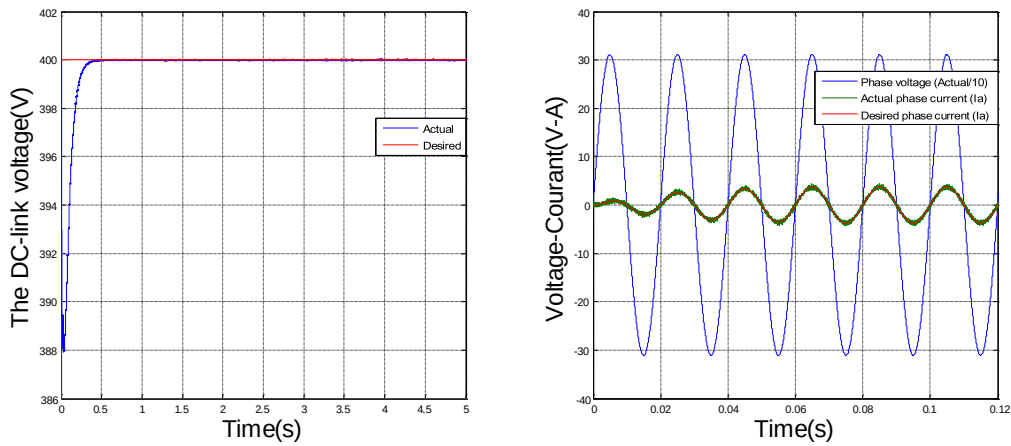


Figure I-6: (a): DC bus voltage, (b): line current and voltage.

I.4.4 Interpretation of results

Clearly, the DC voltage follows the shape of the imposed reference voltage see Figure I-6 (a). Additionally, the line currents perfectly follow the sinusoidal shapes of the reference currents, which are in phase with the line voltage see Figure I-6 (b). These results demonstrate the effectiveness and robustness of grid-side converter voltage orientation control in reducing harmonics and improving the power factor.

I.5 Turbine modelling and control

All variable-speed wind energy conversion systems are driven by a turbine with variable blade pitch to control energy capture from the wind during low- and high-speed periods. Accurate modelling and robust controls for different wind speeds are required to ensure these turbines operate correctly. [30,31]

I.5.1 Wind turbine modelling

Our variable-speed wind energy conversion system uses a horizontal-axis turbine with three steerable blades, each of which measures R_T in length. The blades are connected to a drive shaft, which is linked to a speed multiplier with a transformation ratio of G . This multiplier then drives the electric generator shaft. The model of this turbine is given by the following equations [31,32]:

The turbine power is P_T :

$$P_T = C_p(\lambda, \beta) \cdot P_{mt} = \frac{1}{2} \cdot \rho \cdot \pi \cdot R_T^2 \cdot V^3 \cdot C_p(\lambda, \beta) \quad (I.24)$$

ρ : The density of air is 1.22 kg/m³ at 15 °C and under normal atmospheric pressure.

R_T : is the radius of the turbine blades.

$C_p(\lambda, \beta)$: is the turbine's power coefficient. To maximise power and efficiency $C_p = 0,5$ the values of λ and β must be kept at their optimal levels. $\lambda = 9,2$ et $\beta = 2$.

$$C_p(\lambda, \beta) = (0,5 - 0,0167 \cdot (\beta - 2)) \cdot \sin \left[\frac{\pi \cdot (\lambda + 0,1)}{18,5 - 0,3 \cdot (\beta - 2)} \right] - 0,00184 \cdot (\lambda - 3) \cdot (\beta - 2) \quad (I.25)$$

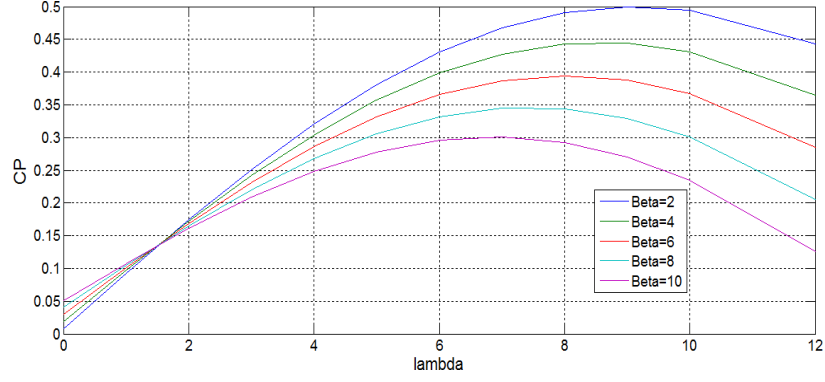


Figure I-7: The power coefficient varies according to the blades' orientation and the specific speed.

λ : The specific speed is represented by

$$\lambda = \frac{\Omega_T \cdot R_T}{V} \quad (\text{I. 26})$$

The speed and torque of the turbine are therefore determined as follows:

$$C_T = \frac{P_T}{\Omega_T} = \left(\frac{1}{2} \cdot \rho \cdot \pi \cdot R_T^3 \cdot V \cdot C_p(\lambda, \beta) \right) \cdot \left(\frac{R_T}{V \cdot \lambda} \right) \quad (\text{I. 27})$$

$$C_T = \frac{1}{2} \cdot \rho \cdot \pi \cdot \frac{R_T^3 \cdot V^2}{\lambda} \cdot C_p(\lambda, \beta) \quad (\text{I. 28})$$

The multiplier gains are presented by :

$$G = \frac{C_T}{C_{Mec}} \quad (\text{I. 29}), \quad \text{And} \quad G = \frac{\Omega_g}{\Omega_T} \quad (\text{I. 30})$$

The multiplier gains are presented by :

$$\frac{C_T}{G} - C_g = J \cdot \frac{d\Omega_g}{dt} + f \cdot \Omega_g \quad (\text{I. 31})$$

C_T, C_{Mec}, C_g : represent the turbine torque, the mechanical torque from the gearbox and the electromagnetic torque generated by the generator, respectively.

G : is the speed multiplier ratio.

Ω_g, Ω_T : represent the rotation speeds of the generator and turbine, respectively.

J : represents the turbine's inertia.

f : is the turbine's coefficient of viscous friction.

I.5.2 Wind turbine control for low and high wind speeds

To optimise the conversion of kinetic energy into mechanical energy, two control mechanisms are employed.

Indirect maximum power point tracking (MPPT) maximises the power sensor during low wind speeds.

Pitch control stabilises and regulates the converted power during high wind speeds.

I.5.2.1 Wind turbine control for low wind speeds

For low wind speeds: The available power in this zone is less than the turbine's rated power. The control system's objective is to extract the maximum possible power from the wind using a technique known as indirect MPPT [33–34].

I.5.2.1.1 Indirect maximum power extraction technique (MPPT)

To implement indirect MPPT control, the values of two variables β_{opt} et λ_{opt} must be kept at their optimum levels to ensure a maximum power coefficient C_{pmax} . The algorithm for this control is determined by the following equations:

Wind speed is given by:

$$V = \frac{\Omega_T \cdot R_T}{\lambda_{opt}} \quad (I.32)$$

λ_{opt} : The optimal specific speed.

The optimum torque is:

$$C_{T_{opt}} = \frac{1}{2} \cdot \rho \cdot \pi \cdot R_T^3 \cdot V^2 \cdot \frac{C_p(\lambda_{opt})}{\lambda_{opt}} \quad (I.33)$$

Substituting the wind speed into equation (I.33) gives the following expression for the optimal torque:

$$C_{T_{opt}} = \frac{1}{2} \cdot \rho \cdot \pi \cdot R_T^5 \cdot \frac{C_p(\lambda_{opt})}{\lambda_{opt}^3} \cdot \Omega_T^2 \quad (I.34)$$

The optimum aerodynamic torque is proportional to the rotor speed squared:

$$C_{T_{opt}} = k_{opt} \cdot \Omega_T^2 \quad (I.35)$$

Such as :

$$k_{opt} = \frac{1}{2} \cdot \rho \cdot \pi \cdot R_T^5 \cdot \frac{C_p(\lambda_{opt})}{\lambda_{opt}^3} \quad (I.36)$$

In steady state, the mechanical equation can be expressed as follows:

$$\frac{C_T}{G} - C_g - f \cdot \Omega_g = 0 \quad (I.37)$$

Substituting (I.35) into (I.37) gives us the following mechanical equation:

$$\frac{k_{opt}}{G} \cdot \Omega_T^2 - f \cdot \Omega_g - C_g = 0 \quad (I.38)$$

With : $\Omega_g = G \cdot \Omega_T$

If the electromagnetic torque C_g is equal to the optimum torque, the wind turbine will operate close to its optimum efficiency curve. This optimum torque is denoted by C_{gopt} .

$$C_{gopt} = \frac{k_{opt}}{G^3} \cdot \Omega_g^2 - f \cdot \Omega_g \quad (I.39)$$

I.5.2.2 Wind turbine control for high wind speeds

For high wind speeds. The aim of this control system is to limit and maintain the turbine's speed to ensure it produces the rated power and avoids overloads. This is achieved by adjusting the blade pitch angle (pitch control).

I.5.2.2.1 Pitch control system technique

In order to apply the second pitch control command and maintain the active power produced at its nominal value, the power coefficient can be reduced by adjusting the blade pitch angle. This limits the converted power. This is achieved using a PI-type corrector, which adjusts the generated power $P_{Gén}$ to match the reference power P_{nom} . The rate of change of the reference angle β_{ref} , is obtained by correcting the wedge angle β using the reference value β_{ref} already obtained. The reference speed β_{ref} is then limited $\frac{1}{p}$ integrator is used to calculate the wedge angle β .

The model block diagram of the turbine and its controls is shown below.

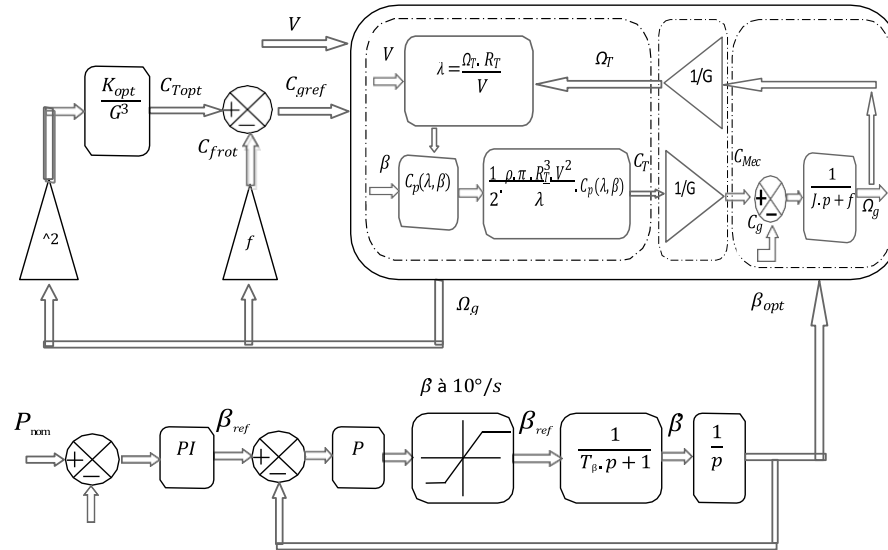


Figure I-8: Block diagrams showing the turbine model and its controls for low and high wind speeds (MPPT and pitch control).

I.5.3 Numerical simulation results

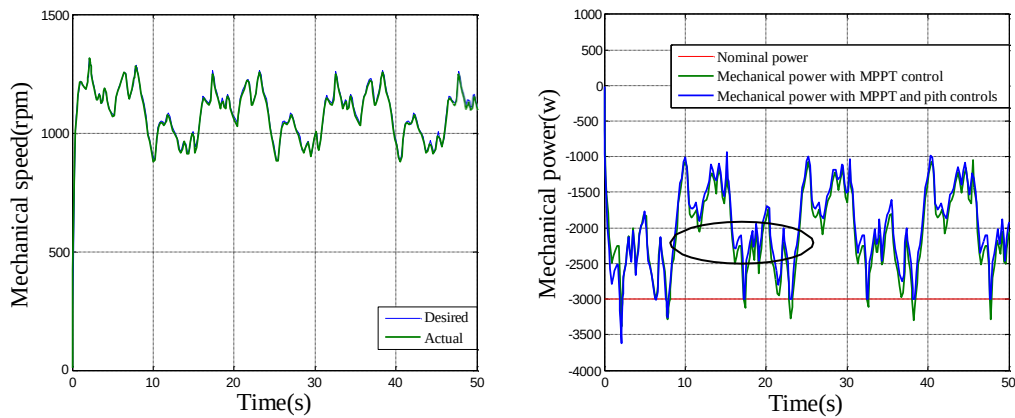


Figure I-9: Mechanical speed and power output.

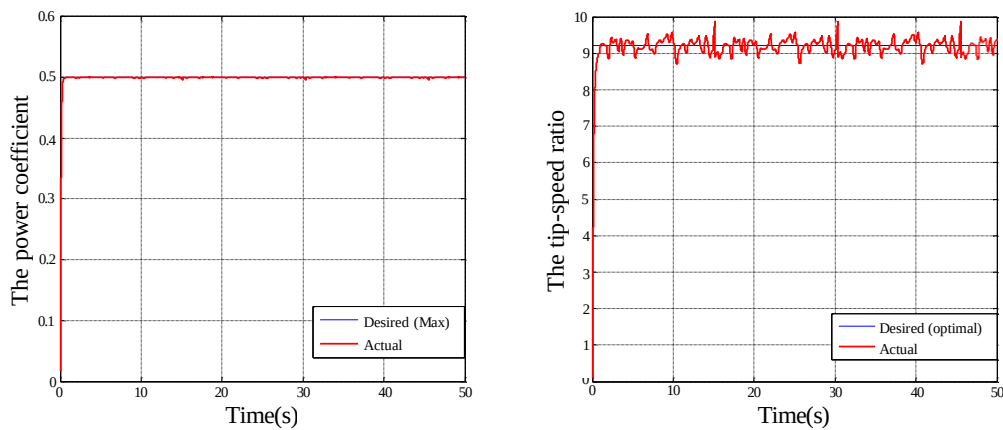


Figure I-10: Power coefficient and specific speed.

I.5.4 Interpretation of results

The simulation enabled us to verify the effectiveness of the two control systems.

For low wind speeds, we employed indirect maximum power point tracking (MPPT), with the speed adhering to the optimum reference value and the power coefficient aligning with the desired maximum. In pitch control for high wind speeds, the mechanical speed varies in line with the wind speed. However, the pitch control system strictly limits the converted power to its nominal value in order to protect the wind turbine from possible overloads.

I.6 Conclusion

In this chapter, we have presented the modelling of the variable speed wind energy conversion chain separately. This chain consists of a double-fed asynchronous machine driven by a variable-speed turbine and controlled by the rotor via two bi-directional power converters operating in PWM and connected to the grid.

First, we modelled the double-fed asynchronous machine in the Park reference frame (a two-phase (d, q) reference frame linked to the rotating field), which simplified the modelling process. Next, we modelled the converter on the machine side that feeds the rotor of the DFIG. This allows commands to be applied to control the power generated by the machine's stator.

Finally, we modelled the grid-side converter and its control system, which maintains a constant DC bus voltage. This improves the grid-side power factor.

Lastly, we modelled the wind turbine and its two controls: indirect maximum power point tracking (MPPT) to maximise power collection during low wind speeds and blade orientation to limit and normalise converted power during high wind speeds.

We simulated the conversion chain models using MATLAB software. The simulation results obtained are similar to those already published.

II.1 Introduction

Efficient control of active and reactive power is essential for the stability and performance of modern electrical energy systems, particularly when integrating renewable energy sources. This chapter examines how sliding mode control (SMC) can effectively and robustly manage power flow. SMC is widely recognised for its ability to handle nonlinearities and external disturbances, making it ideal for dynamic and uncertain environments. It begins by outlining the fundamental principles of SMC and its application to power control, before exploring the advantages it offers over traditional control techniques. The aim is to utilise SMC to achieve precise and reliable power regulation, thereby enhancing system performance and resilience [35].

II.2 Sliding mode control of active and reactive power

In what follows, we will present a theoretical study of this control system to demonstrate its effectiveness. We will adapt it to control the active and reactive power generated by an DFIG in a wind energy conversion system. Subsequently, we will present simulation results to demonstrate and evaluate the performance of this control technique. The basic idea behind this control is to manipulate the system's dynamics or state. This is achieved by selecting a suitable region, known as the sliding surface. The next step is to design a control law to ensure that the system remains within that region at all times. Sliding control can be divided into two parts. Sliding control can be divided into two parts [36] :

- Synthesising a surface such that all the system's trajectories obey a desired tracking, regulation and stability behaviour;
- The second part determines a switching control law that can maintain all the trajectories on this surface.

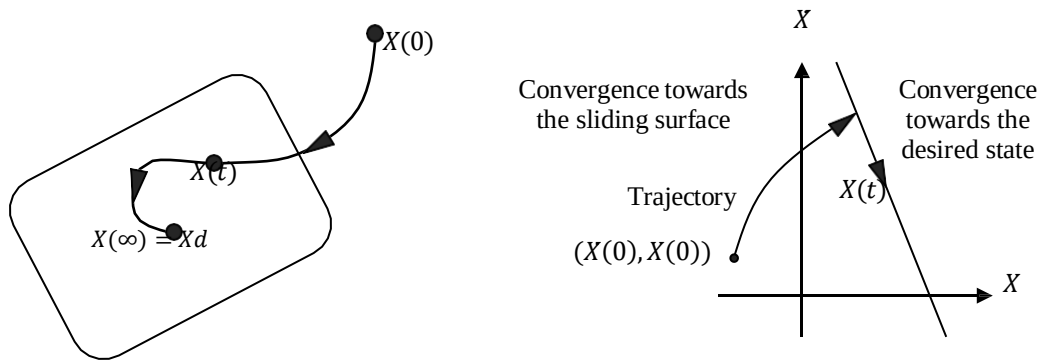


Figure II-1: The vergence of the sliding system.

II.3 Design of the sliding mode control algorithm

Sliding mode control provides a systematic approach to addressing stability and performance issues. This approach involves three complementary stages, as defined in :

- The choice of sliding surfaces;
- definition of the conditions of existence and convergence of the sliding regime;
- determination of the control law.

II.3.1 Choice of sliding surfaces

For stabilisation purposes and to define the desired dynamics of the system in sliding mode, the sliding surface $s(x)$ can generally be chosen as a hyperplane passing through the origin of space.

The following state model is considered:

$$[X] = [A]. [X] + [B]. [U] \quad (\text{II. 1})$$

where $[X]$: represents the state vector;

$[U]$: the control vector.

In general, the number of sliding surfaces selected is equal to the dimension of the control vector U for a state variable X to converge to its reference value X^* , a sliding surface must be selected that is a scalar function along which the controlled variable moves. In this work, we are interested in the following nonlinear surface:

$$s(x) = \left(\frac{d}{dt} - \lambda \right)^{r-1} . e(x) \quad (\text{II. 2})$$

with :

λ : The desired control bandwidth is determined by the positive gain.

$e(x) = X - X^*$: The deviation of the variable to be controlled.

r : the relative degree. This is the smallest positive integer representing how many times the control must be applied for it to appear.

$$\text{for } r = 1 \quad s(x) = e(x).$$

$$\text{for } r = 2 \quad s(x) = \lambda_x e(x) + \dot{e}(x).$$

$$\text{for } r = 3 \quad s(x) = \lambda_x^2 . e(x) + 2 * \lambda_x \dot{e}(x) + \ddot{e}(x).$$

$s(x)$: This is an autonomous linear differential equation whose response ($e(x)$) approaches zero when the gain (λ) is set correctly. This is the objective of the control.

In other words, the challenge is to track the trajectory while keeping $s(x)$ at zero. This is equivalent to the exact linearisation of the deviation $e(x)$, provided that the convergence condition is met. The purpose of exact linearisation is to ensure that the dynamics of the deviation (reference minus output) are the same as those of an autonomous linear system of order r .

In summary, sliding speed control can be divided into two types:

II.3.1.1 Ideal sliding regime

In theory, the switching device is insensitive to noise, and the sliding trajectory describes $S(x) = 0$ perfectly. This involves infinite-frequency, zero-amplitude oscillations, with the point representing the system's evolution sliding perfectly on the switching hypersurface S .

II.3.1.2 Real sliding regime

In practice, the switching device is made from an imperfect relay, which introduces switching delays. In this case, the phase trajectory of the sliding regime remains close to the switching surface. This gives rise to undesirable oscillations that eliminate the system's precision and stability.

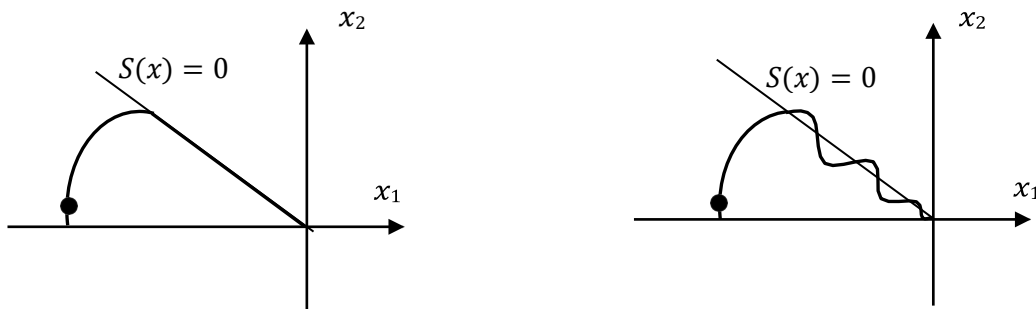


Figure II-2: Ideal and real slippage.

II.3.2 Existence and convergence conditions for the sliding regime

These conditions determine whether the system's dynamics will converge towards, and remain on, the sliding surface despite disturbances. Two types of condition are presented:

II.3.2.1 Direct approach

This is the oldest approach, proposed and studied by Emilyanov and Utkin. The principle is to ensure that the product of the surface function $s(x)$ and its derivative remains less than zero: $s'(x)s(x) < 0$.

II.3.2.2 Lyapunov approach

This approach involves selecting a Lyapunov candidate function, $V(x)$, which is a positive scalar function for the system state variables, and selecting a control law that causes this function to decrease.

For example, this can be achieved by defining a Lyapunov candidate function for the system as follows:

$$V(x) = \frac{1}{2}s(x)^2 \quad (\text{II. 3})$$

Deriving the latter, we obtain:

$$\dot{V}(x) = \dot{s}(x) \cdot s(x) \quad (\text{II. 4})$$

In order for the Lyapunov candidate function to decrease, it is sufficient to ensure that:

$$\dot{s}(x) \cdot s(x) < 0 \quad (\text{II. 5})$$

This approach is used to estimate control performance and study the robustness and stability of nonlinear systems.

II.3.3 Determining the control law

The control law required to attract and maintain the state trajectory towards the surface, and then towards its equilibrium point, has been determined. This ensures that the conditions for the existence of the sliding mode are met.

The control law $U(t)$ comprises two controls: U_{eq} , relating to the exact linearisation of the system, and ΔU , relating to the stabilisation of the system. The latter is important in sliding mode control techniques because it eliminates the effects of model inaccuracy and rejects external disturbances.

$$U(t) = \Delta U + U_{eq} \quad (\text{II. 6})$$

The nominal control, U_{eq} , is determined by the system model. It can be considered the average value of the control during rapid switching between the two values, U_{max} and U_{min} . It is used to keep the controlled variable on the sliding surface, $\dot{s}(x) = 0$. The equivalent control is derived by setting the derivative of the surface to zero.

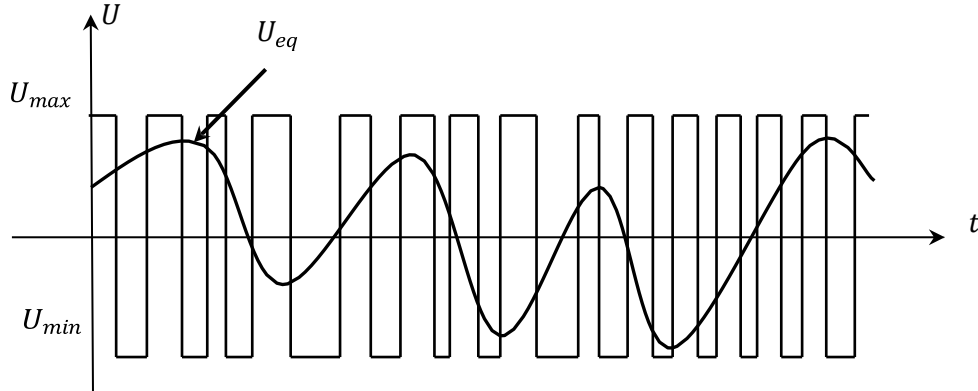


Figure II-3: The continuous value U_{eq} taken by the controller when switching between U_{max} et U_{min}

ΔU : Slip or attractive control can be used to compensate for model uncertainties. This consists of the sign function of the sliding surface ($s(x)$) multiplied by a constant k_x . The slip surface is defined in the error state space to ensure state convergence. This verifies the convergence condition $\dot{s}(x) \cdot s(x) < 0$.

To demonstrate this, we consider the state system (II.1). We seek to determine the analogue expression of the control U .

The derivative of the surface $s(x)$ is:

$$\dot{s} = \frac{ds}{dt} = \frac{\partial s}{\partial x} \frac{\partial x}{\partial t} \quad (II. 7)$$

Replacing (II.1) and (II.6) in (II.7), we find :

$$\dot{s}(x) = \frac{ds}{dt} = \frac{\partial s}{\partial x} ([A][X] + [B]U_{eq}) + \frac{\partial s}{\partial t} [B]\Delta U \quad (II. 8)$$

Durant le mode de glissement et le régime permanent, la surface est nulle et par conséquent, sa dérivée et la partie discontinue sont aussi nulles. D'où on déduit l'expression de la commande équivalente.

$$U_{eq} = -(\frac{\partial s}{\partial x} [B])^{-1} (\frac{\partial s}{\partial x} [A][X]) \quad (II. 9)$$

For the equivalent command to take on a finite value, the following must be true:

$$\frac{\partial s}{\partial X}[B] \neq 0 \quad (\text{II. 10})$$

In convergence mode, replacing the equivalent command with the expression given in (II.8) yields a new derivative expression for the surface.

$$\dot{s}(x) = \frac{\partial s}{\partial X}[B]\Delta U \quad (\text{II. 11})$$

and the attractiveness condition $s(X) \cdot \dot{s}(X) \leq 0$ becomes :

$$s(x) \frac{\partial s}{\partial X}[B]\Delta U < 0 \quad (\text{II. 12})$$

To satisfy the condition, the sign of ΔU must be opposite to that of $s(x) \frac{\partial s}{\partial X}[B]$. The simplest form that discrete control can take is a **sign** function (see Figure II-4).

$$\Delta U = k_x \text{sign } s(x) \quad (\text{II. 13})$$

The sign of k_x must differ from that of $\frac{\partial s}{\partial X}[B]$

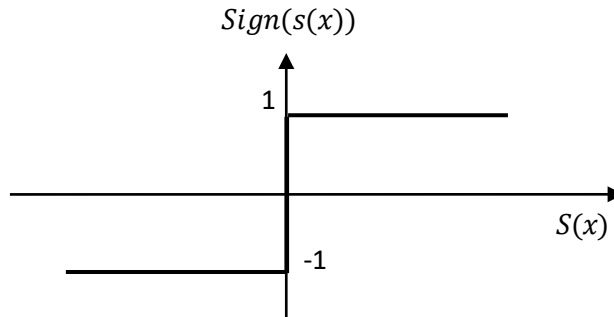


Figure II-4: Sign function (non-linear on/off).

II.4 Eliminating the chattering phenomenon

Equivalent control and technological imperfections, such as delays and hysteresis in actuators and sensors, can cause undesirable oscillations known as 'chattering'. This phenomenon poses an obstacle to developing a control system that can guide the trajectory towards the sliding surface with ideal oscillations and infinite frequencies [37].

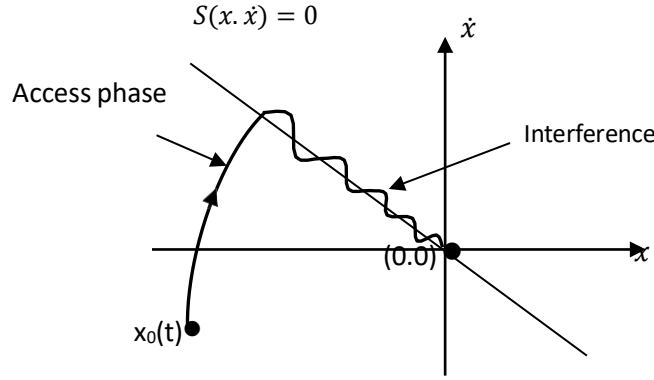


Figure II-5: Grazing phenomenon.

Chattering has a number of undesirable effects on the system's behaviour.

These include:

- the creation of unmodelled dynamics;
- reduced control accuracy;
- increased energy losses;
- increased fatigue of moving mechanical parts.

Several techniques have been proposed to reduce or eliminate these issues. Two of these solutions are presented below.

II.4.1 Slotine proposal

Replace the 'sign' function with a 'saturation' function characterised by one or two thresholds that attenuate ripple amplitudes, as shown in Figure II-6.

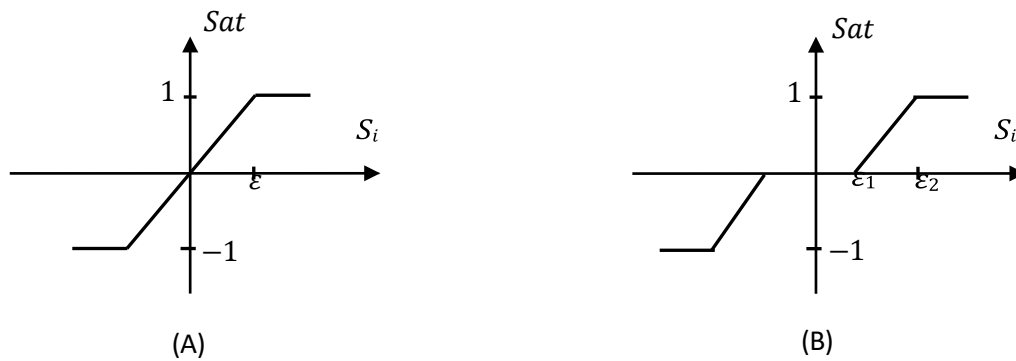


Figure II-6: SAT function: (A) - one threshold; (b) - two thresholds (dead zone).

In this case, the command becomes: $\Delta U = -k_x \text{cont } s(x)$.

The two thresholds of this function are defined as follows:

$$sat(s) = \begin{cases} -1 & si \quad s < -\varepsilon \\ \frac{s}{\varepsilon} & si \quad |s| < \varepsilon \\ 1 & si \quad s > \varepsilon \end{cases} \quad (II. 14)$$

$$sat(s) = \begin{cases} 0 & si \quad |s| < \varepsilon_1 \\ \frac{s - \varepsilon_1}{\varepsilon_2 - \varepsilon_1} & si \quad \varepsilon_1 < |s| < \varepsilon_2 \\ \{sign(s)\} & si \quad s > \varepsilon_2 \end{cases} \quad (II. 15)$$

II.4.2 Harshima's proposal

In this proposal, the 'sign' function is replaced by a smoothing function called 'cont' (see figure II-17)

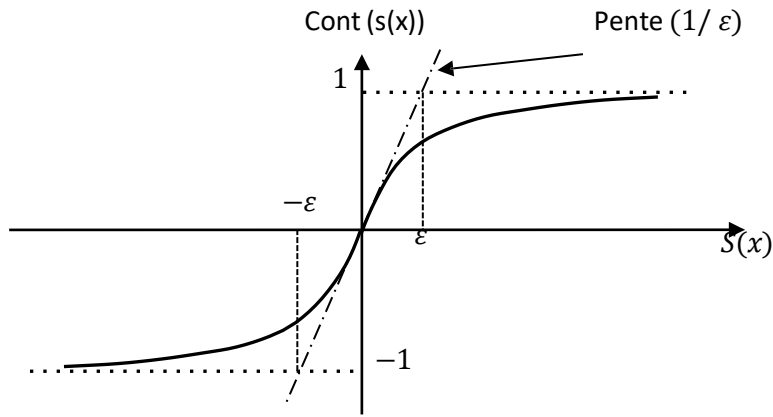


Figure II-7: Cont function ($\varepsilon >$ and small).

In this case, the command becomes: $\Delta U = -k_x cont s(x)$.

This function is defined as follows:

$$cont(s) = \begin{cases} \frac{s(x)}{|s(x)| + \varepsilon} & si \quad |s| < \varepsilon \\ \{sign(s)\} & sino \end{cases} \quad (II. 16)$$

II.5 Different structures of sliding mode control

In general, three basic configurations are used to control this mode. The first involves switching on the controller itself (known as the 'controller switching structure'), the second involves switching on state feedback (known as the 'state feedback switching structure'), and the third involves adding an equivalent control (known as the 'equivalent control structure'). The latter structure was chosen for the remainder of our study.

II.5.1 Switching structure at the actuator level

Figure II-8 uses a block diagram to show how the switching structure works at the actuator level.

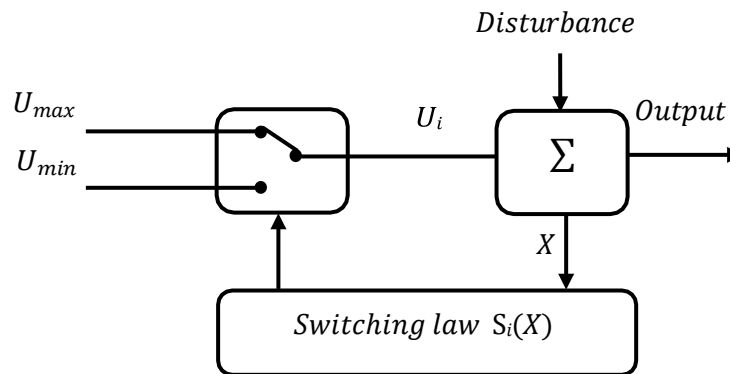


Figure II-8 : Control structure based on switching at the control unit.

This control structure is the most traditional and widely used. It corresponds to the 'digital' on/off operation of the power switches associated with variable speed drives in the vast majority of applications. It has been used to control stepper motors

II.5.2 Switching structure with state feedback

The block diagram representing the operating principle of the state feedback switching structure is given in Figure (II-9)

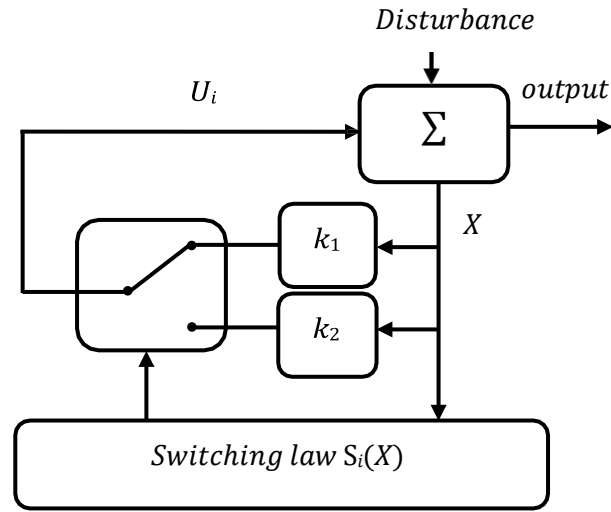


Figure II-9: Switching control structure with state feedback.

Since the non-linearity of the system comes from the switching between the gains, we have created a switching at the level of the dynamics of the system to be controlled, based on classical state feedback control. This structure is the least demanding in terms of control oscillations. It has been used to control DC and permanent magnet motors, as well as induction machines.

II.5.3 Control structure with the addition of equivalent control

The block diagram representing the operating principle of the control structure with added equivalent control is given in Fig. (II-10).

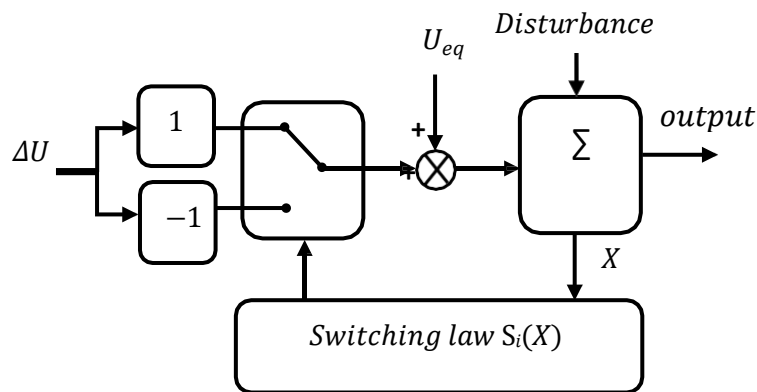


Figure II-10: Control structure by adding the equivalent control.

This control has the advantage of giving the future state of the system thanks to the equivalent control. The latter is none other than the desired value of the system in steady state.

II.6 Applications of sliding mode control to DFIG

Using sliding mode control in a wind energy conversion system with a double-fed asynchronous generator provides robust control and solves many problems related to uncertainties and variations in the controlled system. This study aims to apply this control technique to enable independent regulation of the active and reactive power generated by a flux-directed asynchronous machine. The numerical simulation results obtained demonstrate the growing interest in such control systems for wind energy [38,39].

II.6.1 DFIG double-fed asynchronous generator model

The model used is the stator flux-oriented model presented previously, with all electrical quantities expressed in a fixed reference frame linked to the stator in the (d, q) reference frame.

The state model is given as follows:

$$\begin{cases} V_{sd} = 0 \\ V_{sq} = V_s = \omega_s \cdot \varphi_s \\ V_{rd} = R_r \cdot I_{rd} + \frac{d\varphi_{rd}}{dt} - \omega_r \varphi_{rq} \\ V_{rq} = R_r \cdot I_{rq} + \frac{d\varphi_{rq}}{dt} + \omega_r \varphi_{rd} \end{cases} \quad (\text{II. 17})$$

❖ **The relationship between stator power and rotor currents.**

$$P_s^{ref} = - \frac{V_s \cdot M}{L_s} \cdot I_{rq}^{ref} \quad (\text{II. 18})$$

$$\begin{cases} Q_s^{ref} = \frac{V_s^2}{\omega_s \cdot L_s} - \frac{V_s \cdot M}{L_s} \cdot I_{rd}^{ref} \\ I_{rq}^{ref} = - \frac{L_s}{V_s \cdot M} \cdot P_s^{ref} \\ I_{rd}^{ref} = \frac{V_s}{\omega_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q_s^{ref} \end{cases} \quad (\text{II. 19})$$

❖ **Derivative equations for rotor currents**

$$\begin{cases} \frac{dI_{rd}}{dt} = \left(V_{rd} - R_r \cdot I_{rd} + g \cdot \omega_s \cdot L_r \cdot \sigma \cdot I_{rq} \right) \cdot \frac{1}{L_r \sigma} \\ \frac{dI_{rq}}{dt} = \left(V_{rq} - R_r \cdot I_{rq} - g \cdot \omega_s \cdot L_r \cdot \sigma \cdot I_{rd} - g \cdot \omega_s \cdot \frac{M \cdot V_s}{\omega_s \cdot L_s} \right) \cdot \frac{1}{L_r \sigma} \end{cases} \quad (\text{II. 20})$$

II.6.2 Selection of sliding surfaces

Active power is directly proportional to the q-axis rotor current, while reactive power is proportional to the d-axis rotor current. To control the power, $n = 1$ is used.

The expressions for the active and reactive power control surfaces are as follows:

$$s(P) = (I_{rq}^{ref} - I_{rq}) \quad (II.21)$$

$$s(Q) = (I_{rd}^{ref} - I_{rd}) \quad (II.22)$$

II.6.3 Convergence conditions

For the chosen variables to converge towards their reference values, the two sliding surfaces must be zero.

$$\begin{cases} s(P) = 0 \\ s(Q) = 0 \end{cases} \Rightarrow \begin{cases} \frac{d}{dt} (I_{rq}^{ref} - I_{rq}) = 0 \\ \frac{d}{dt} (I_{rd}^{ref} - I_{rd}) = 0 \end{cases} \quad (II.23)$$

When these conditions are met, the active and reactive powers will converge exponentially towards their reference values. To achieve this, it is sufficient to make the sliding surface attractive and invariant.

The sliding mode is achieved when the Lyapunov attractivity relation is less than zero $s(X) \cdot \dot{s}(X) \leq 0$

II.6.4 Sliding mode control law

The control algorithm is defined by the following relations:

$$V_{rq} = V_{rq-Equi} + V_{rq-attr} \quad (II.24)$$

$$V_{rd} = V_{rd-Equi} + V_{rd-attr} \quad (II.25)$$

With : V_{rq}, V_{rd} : the control voltages;

$V_{rq-Equi}, V_{rd-Equi}$: the equivalent control voltages;

$V_{rq-attr}, V_{rd-attr}$: are the switching control voltages.

❖ Active power control

To control the active power, we set $r = 1$. The expression for the sliding surface becomes:

$$s(P) = (I_{rq}^{ref} - I_{rq}) \quad (II.26)$$

Its derivative is:

$$\dot{s}(P) = (I_{rq}^{ref} - I_{rq}) \quad (II.27)$$

Substituting the expressions for the currents and reference currents I_{rq}^{ref} And I_{rq} given in equations (II.19) and (II.20) for their derivatives yields the following:

$$\dot{s}(P) = \left(-\frac{L_s}{M.V_s} \cdot P_s^{ref} - \frac{1}{L_r\sigma} \cdot (V_{rq} - R_r \cdot I_{rq} - g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rd} - g \cdot \frac{M.V_s}{L_s}) \right) \quad (II.28)$$

$$\dot{s}(P) = \left(-\frac{L_s}{M.V_s} \cdot P_s^{ref} - \frac{1}{L_r\sigma} \cdot V_{rq} - \frac{1}{L_r\sigma} \cdot (-R_r \cdot I_{rq} - g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rd} - g \cdot \frac{M.V_s}{L_s}) \right) \quad (II.29)$$

$$\dot{s}(P) = -v_1 \operatorname{sgn}(s(P)) \quad (II.30)$$

$$-\frac{L_s}{M.V_s} P_s^{ref} - \frac{1}{L_r\sigma} V_{rq} - \frac{1}{L_r\sigma} (-R_r \cdot I_{rq} - g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rd} - g \cdot \frac{M.V_s}{L_s}) = -v_1 \operatorname{sgn}(s(P)) \quad (II.31)$$

$$V_{rq} = -\frac{L_s L_r \sigma}{M.V_s} \cdot P_s^{ref} - (-R_r \cdot I_{rq} - g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rd} - g \cdot \frac{M.V_s}{L_s}) + L_r \cdot \sigma \cdot v_1 \operatorname{sgn}(s(P)) \quad (II.32)$$

$$V_{rq} = -\frac{L_s L_r \sigma}{M.V_s} \cdot P_s^{ref} + R_r \cdot I_{rq} + g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rd} + g \cdot \frac{M.V_s}{L_s} + L_r \cdot \sigma \cdot v_1 \operatorname{sgn}(s(P)) \quad (II.33)$$

$$V_{rq-Equi} = -\frac{L_s L_r \sigma}{M.V_s} \cdot P_s^{ref} + R_r \cdot I_{rq} + g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rd} + g \cdot \frac{M.V_s}{L_s} \quad (II.34)$$

$$\{ V_{rq-attr} = L_r \cdot \sigma \cdot v_1 \operatorname{sgn}(s(P))$$

❖ Reactive power control

As above, set r to 1 to control the reactive power. This gives the following expression for the sliding surface:

$$s(Q) = (I_{rd}^{ref} - I_{rd}) \quad (II.35)$$

Its derivative is:

$$\dot{s}(Q) = (I_{rd}^{ref} - I_{rd}) \quad (II.36)$$

By replacing the derivatives of the reference currents I_{rd}^{ref} & I_{rd} with the values given in equations II.19 and II.20, respectively, we obtain the following:

$$\dot{s}(Q) = \left(\left(\frac{V_s}{w_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q^{ref} \right) - \frac{1}{L_r \sigma} \left(V_{rd} - R_r \cdot I_{rd} + g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rq} \right) \right) \quad (\text{II. 37})$$

$$\dot{s}(Q) = \left(\left(\frac{V_s}{w_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q^{ref} \right) - \frac{1}{L_r \sigma} V_{rd} - \frac{1}{L_r \sigma} \left(-R_r \cdot I_{rd} + g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rq} \right) \right) \quad (\text{II. 38})$$

$$\dot{s}(Q) = -v_2 \cdot \text{sgn}(s(Q)) \quad (\text{II. 39})$$

$$\left(\frac{V_s}{w_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q^{ref} \right) - \frac{1}{L_r \sigma} V_{rd} - \frac{1}{L_r \sigma} \left(-R_r \cdot I_{rd} + g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rq} \right) = -v_2 \cdot \text{sgn}(s(Q)) \quad (\text{II. 40})$$

$$V_{rd} = L_r \sigma \left(\frac{V_s}{w_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q^{ref} \right) - \left(-R_r \cdot I_{rd} + g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rq} \right) + L_r \cdot \sigma \cdot v_2 \cdot \text{sgn}(s(Q)) \quad (\text{II. 41})$$

$$V_{rd} = L_r \sigma \left(\frac{V_s}{w_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q^{ref} \right) + R_r \cdot I_{rd} - g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rq} + L_r \cdot \sigma \cdot v_2 \cdot \text{sgn}(s(Q)) \quad (\text{II. 42})$$

$$V_{rd-Equi} = L_r \sigma \left(\frac{V_s}{w_s \cdot M} - \frac{L_s}{V_s \cdot M} \cdot Q^{ref} \right) + R_r \cdot I_{rd} - g \cdot w_s \cdot L_r \cdot \sigma \cdot I_{rq} \quad (\text{II. 43})$$

$$\{V_{rd-attr} = L_r \cdot \sigma \cdot v_2 \cdot \text{sgn}(s(Q))\}$$

Equations (II.34) and (II.43) allow us to create a block diagram of a sliding-mode controller for a double-fed asynchronous generator (see Figure II-11).

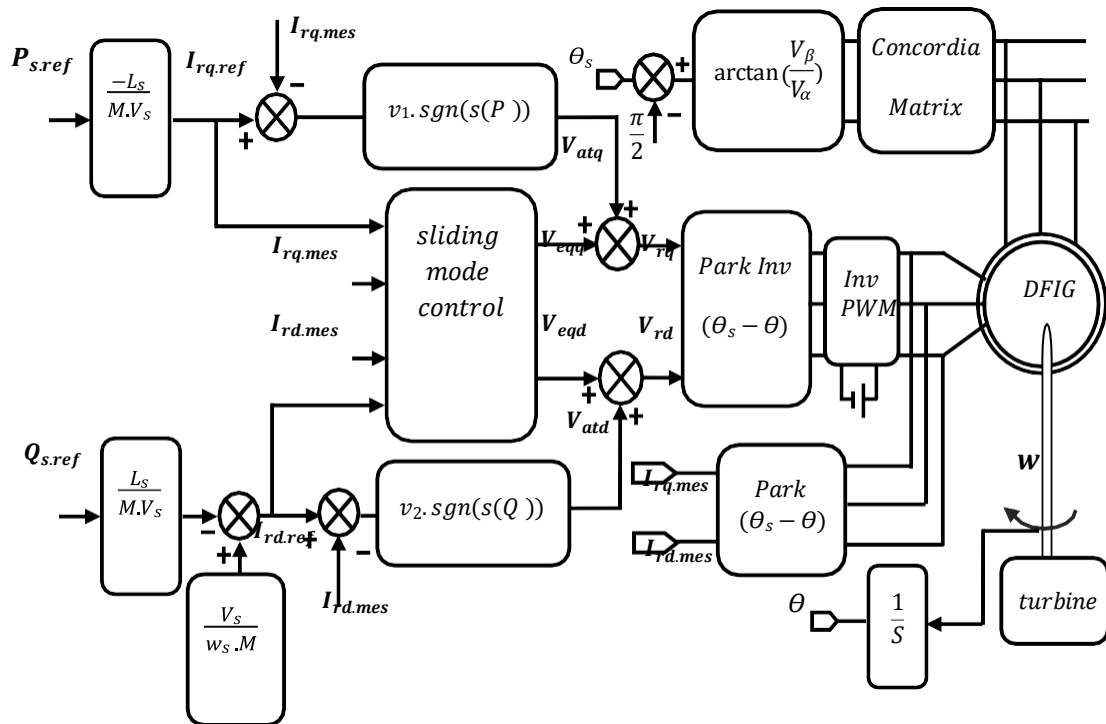


Figure II-11: block diagram illustrates the sliding-mode control structure.

II.7 Simulation results

II.7.1 The results of a numerical simulation of the system at a fixed speed

The figures below show the performance of indirect, power-loop-free vector control and sliding-mode control of the stator's active and reactive power applied to the DFIG. This test was carried out under the following conditions:

- ❖ The machine is connected to the mains;
- ❖ The machine is driven at a fixed speed of $1440 \frac{tr}{min}$;
- ❖ Between $t = 1s$ et $t = 3s$: the active power (P_{sref} undergoes a negative step (decreasing from 0 to $-3000W$));
- ❖ Between $t = 2s$ et $t = 4s$: the reactive power (Q_{sref} increases from 0 à $1000 VAR$).

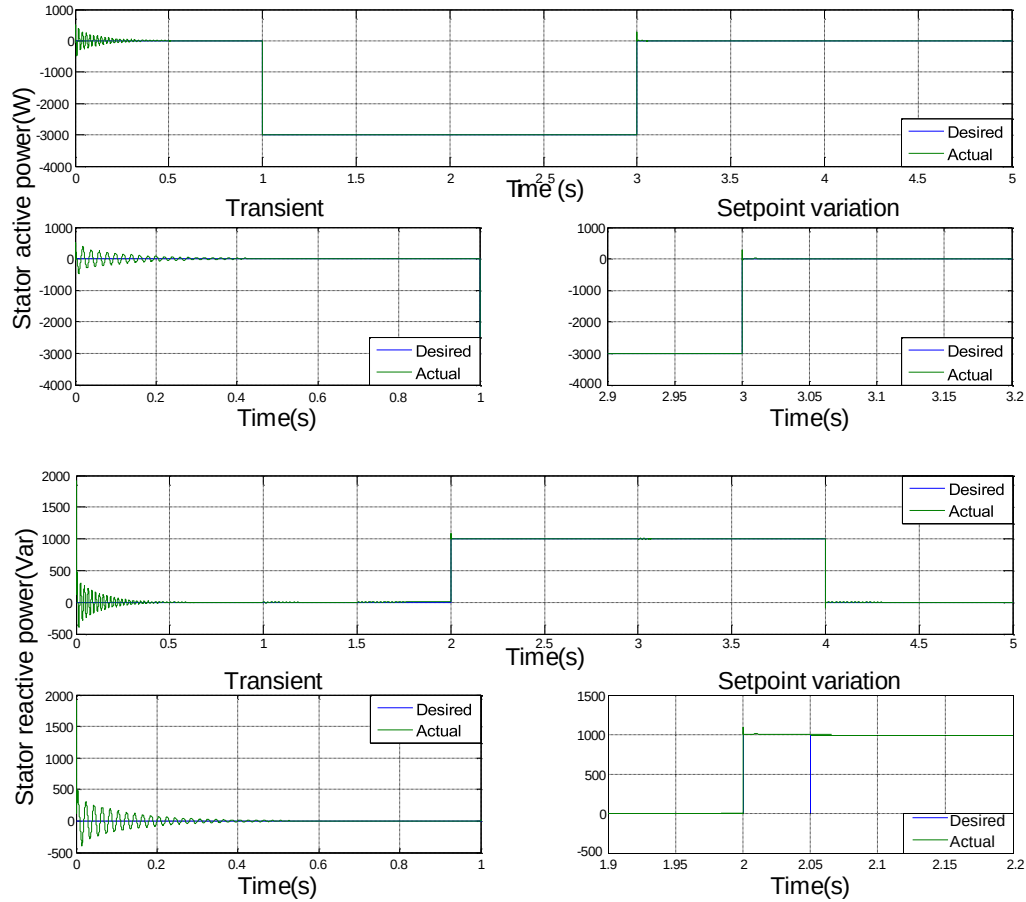


Figure II-12: Active and reactive stator power for the sliding mode control with a zoom

II.7.2 Interpretation of the results

The simulation results clearly show that the sliding mode control provides perfect decoupling between the two stator power components (active and reactive). In addition, we can see that :

- ✓ There is a rapid response in the transient regime and when setpoints are changed.
- ✓ The error between the setpoints and the measured values is minimised (almost zero).
- ✓ There is also a significant reduction in power oscillations.

II.7.3 The results of a numerical simulation of the system at a variable speed

In the most practical case, the machine is coupled directly to the grid through the stator and controlled by the rotor magnitudes through two bi-directional PWM converters, and driven by a turbine with variable blade pitch. The stator active power setpoint is determined from the turbine power, while the stator reactive power is kept zero to keep the unit power factor on the stator side of the DFIG. The time during which the measurements were performed is 50s.

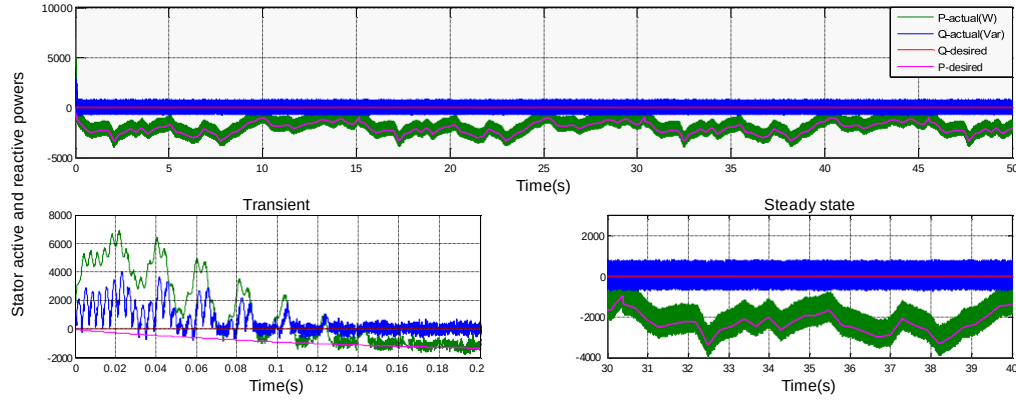


Figure II-13: Active and reactive stator powers with a zoom

II.7.4 Interpretation of the results

As shown in Figure II-13, the stator active and reactive power profiles generated by the machine follow the reference profiles well in the enhanced sliding mode control. This results in a lower response time in the transient regime, minimum overshoot of the power curves and less oscillation and disturbance of the active and reactive power values, which quickly return to their reference curves.

II.8 Conclusion

This chapter outlines the design and implementation of a sliding mode control strategy for regulating active and reactive power. This control method is highly robust in the face of system uncertainties and external disturbances, making it ideal for dynamic power systems. Precise control of power flow improves the stability, reliability, and performance of the energy conversion process when using the sliding mode approach. The chapter also establishes the basis for real-time validation, which will be explored further in the next chapter using Processor-in-the-Loop simulation.

III.1 Introduction

This chapter presents an alternative design method by describing an emerging method that is widely used in the industry. This method is called 'processor in the loop' (PiL), belonging to the 'in-the-loop' family of methods [40].

In-the-loop methods are techniques used to develop systems consisting of both a plant and a controller. There are five such techniques. These are:

- Model in the Loop (MiL);
- Software in the Loop (SiL);
- Processor in the Loop (PiL);
- Hardware in the Loop (HiL);
- Rapid Prototyping (RP)

III.2 The different loop methods

Consider the V-cycle for a controller, as shown in Figure III-1. Each step can be assigned to a particular phase. This method can be applied at every stage of the design process, from system-level design to integration.

III.2.1 MiL

Is a technique that only involves models and can be performed at an early design stage, e.g. system-level modelling.

III.2.2 RP

Can be performed during controller design using code generation. This allows the controller to be quickly tested on board.

III.2.3 SiL

Can be performed during software design and involves integrating existing C code into the modelling tool.

III.2.4 PiL

Can be performed just before the integration phase. It is used to integrate the processor behaviour directly within the modelling tool.

III.2.5 HiL

Is used to test the entire real-time system, including the plant and controller.

All of these methods are primarily performed using MATLAB/Simulink tools and several toolboxes, particularly code generation toolboxes such as MATLAB Coder, Simulink Coder and Embedded Coder.

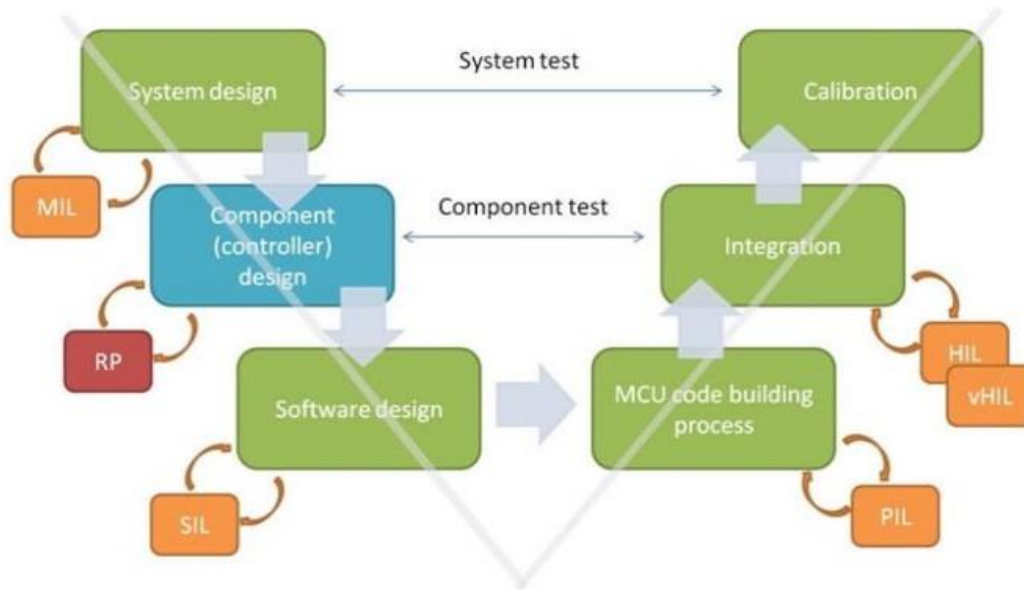


Figure III-1: V-cycle outlines the methods used to design controllers.

III.3 Processor in the loop

The Processor-in-the-Loop (PiL) is a simulation method that compiles and runs the generated code of a model on a specific target. Therefore, both a model and a real processor are required. Figure III-1 illustrates the workflow from the system level to implementation. The PiL method occurs just before the integration phase, once valid and stable models have been developed. Several modelling tools support PiL, including MATLAB/Simulink, Plexim and dSPACE.

PiL is one of the loop methods employed in model-based design. The other methods are the model-in-the-loop (MiL), software-in-the-loop (SiL) and hardware-in-the-loop (HiL) methods. (respectively). After a brief introduction, we will then focus on the principle of PiL, its advantages and disadvantages, and the added value of this method in model-based design.

III.3.1 Principle of the PiL method

Figure III-2 shows that PiL consists of two parts.

- 1) A Simulink model containing a controller model.
- 2) A target board containing a processor (e.g. a C2000 or an STM32).

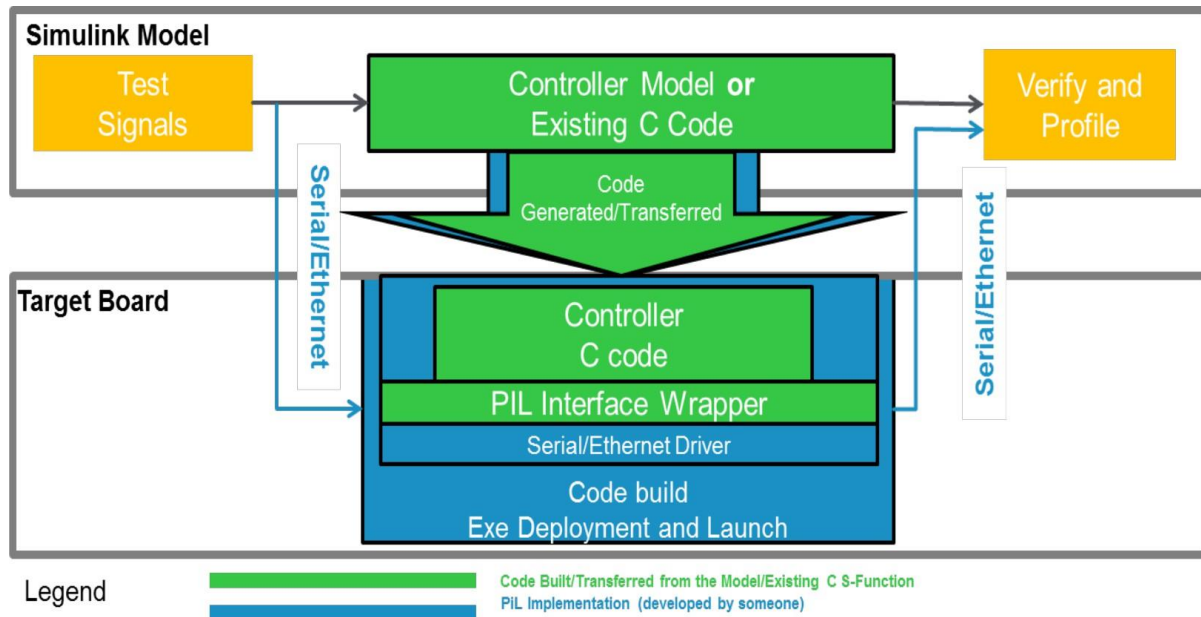


Figure III-2: Schematic describing the principle of PiL.

The primary difference between a PiL simulation and a 'classical' simulation is as follows:

When you click on the 'Launch Simulation' button, this method determines what happens next.

- Build the code to generate the executable corresponding to DUT1.
- Deploy the executable on the target.
- Execute it on the target.
- Communicate with the target (Ethernet, serial link, etc.).

The controller can be either a model or existing C code.

Performing these functions requires a specific implementation, which is displayed in blue around the green implementation in Figure III-2.

This implementation is surrounded by blocks managed by a modelling tool, such as MATLAB/Simulink. These blocks are code generation and a wrapper. These help to link the implementation with the generated model. Once the PiL implementation is complete, an integration phase with the wrapper is then required.

III.3.2 Benefits of the PiL method

The PiL method provides users with additional features that enable more precise modelling than earlier in-the-loop (MiL or SiL) methods.

The key added values provided by this method are:

- Accurate timing
- Testing compiler options and optimisations from a model
- Allowing firmware and hardware validation in advance
- Providing profiling data
- Helping you choose the most suitable processor for your application
- Safely design the controller

III.4 Processor-in-the-loop testing

The proposed SMC and NNC controllers were validated using a Processor-in-the-Loop (PIL) test. This involved using MATLAB/Simulink software and the LAUNCHXL-F28069M DSP board (see Figure III-3 and Figure III-4) In this study, the PIL technique was applied to the LAUNCHXL-F28069M board as follows:

- First, a control algorithm subsystem was created in Matlab/Simulink to interface with the DSP board. This enabled the integrated control algorithm to be executed, the DSP board was then selected as the target model.
- Next, the inputs and outputs of the control system were defined. The control algorithm was developed and uploaded to the DSP board, a USB communication link was set up between the DSP board and Simulink, along with an appropriate time step, the configuration was then set up.
- Finally, the control model was transferred to the DSP board. MATLAB/Simulink then generated the necessary programming code. The control algorithm was then run on the DSP board. The DSP board was then used to run the system, this process uses the MATLAB/Simulink platform.

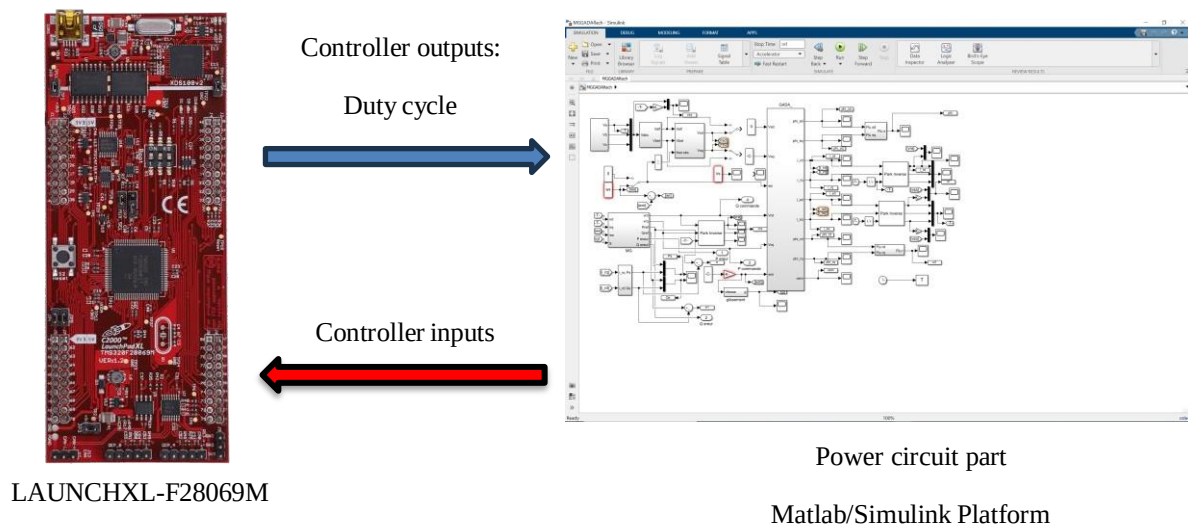


Figure III-3: Processor in the loop platform.

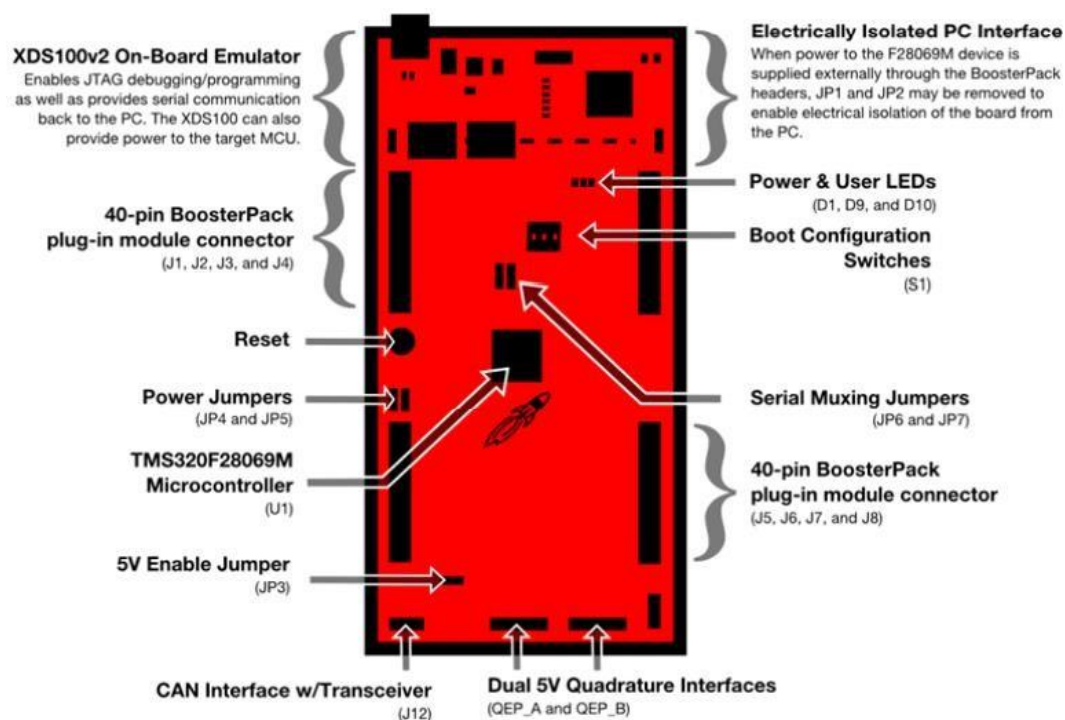


Figure III-4: LAUNCHXL-F28069M Board.

III.5 Device pin out

Table 1: F28069M LaunchPad Pin Out and Pin Mux Options - J1, J3

Mux Value				J1 Pin	J3 Pin	Mux Value			
3	2	1	0			0	1	2	3
			+3.3V	1	21	+5V			
			ADCINA6	2	22	GND			
			J1.3	3	23	ADCINA7			
			J1.4	4	24	ADCINB1			
SPISIMOB	SCITXDA	TZ1	GPIO12	5	25	ADCINA2			
			ADCINB6	6	26	ADCINB2			
XCLKOUT	SCITXDB	SPICKLA	GPIO18	7	27	ADCINA0			
SCITXDB	MCLKXA	EQEP1S	GPIO22	8	28	ADCINB0			
ADCSOCBO	EPWMSYNCO	SCLA	GPIO33	9	29	ADCINA1			
ADCSOCOA	EPWMSYNCI	SDAA	GPIO32	10	30	NC			

Table 2: F28069M LaunchPad Pin Out and Pin Mux Options - J4, J2

Mux Value				J4 Pin	J2 Pin	Mux Value			
3	2	1	0			0	1	2	3
Rsvd	Rsvd	EPWM1A	GPIO0	40	20	GND			
COMP1OUT	Rsvd	EPWM1B	GPIO1	39	19	GPIO19	SP1STE1A	SC1RXDB	ECAP1
Rsvd	Rsvd	EPWM2A	GPIO2	38	18	GPIO44	MFSRA	SC1RXDB	EPWM7B
COMP2OUT	SPISOMIA	EPWM2B	GPIO3	37	17	NC			
Rsvd	Rsvd	EPWM3A	GPIO4	36	16	RESET#			
ECAP1	SPISIMOA	EPWM3B	GPIO5	35	15	GPIO16	SPISIMOA	Rsvd	TZ2
SPISOMIB	Rsvd	TZ2	GPIO13	34	14	GPIO17	SPISOMIA	Rsvd	TZ3

Table 2: F28069M LaunchPad Pin Out and Pin Mux Options - J4, J2 (continued)

Mux Value				J4 Pin	J2 Pin	Mux Value			
3	2	1	0			0	1	2	3
			NC	33	13	GPIO50	EQEP1A	MDXA	TZ1
			DAC1	32	12	GPIO51	EQEP1B	MDRA	TZ2
			DAC2	31	11	GPIO55	SPISOMIA	EQEP2B	HRCAP2

Table 3: F28069M LaunchPad Pin Out and Pin Mux Options - J5, J7

Mux Value				J5 Pin	J7 Pin	Mux Value			
3	2	1	0			0	1	2	3
			+3.3 V	41	61	+5 V			
			NC	42	62	GND			
			J7.3	43	63	ADCINB7			
			J7.4	44	64	ADCINB4			
COMP1OUT	MDXA	EQEP1A	GPIO20	45	65	ADCINA5			
			NC	46	66	ADCINB5			
SPICLKB	SCITXDB	TZ3	GPIO14	47	67	ADCINA3			
COMP2OUT	MDRA	EQEP1B	GPIO21	48	68	ADCINB3			
SCIRXDB	MFSXA	EQEP1I	GPIO23	49	69	ADCINA4			
HRCAP1	EQEP2A	SPISIMOA	GPIO54	50	70	NC			

Table 4: F28069M LaunchPad Pin Out and Pin Mux Options – J8, J6

Mux Value				J8 Pin	J6 Pin	Mux Value			
3	2	1	0			0	1	2	3
EPWMSYNCO	EPWMSYNCI	EPWM4A	GPIO6	80	60	GND			
ECAP2	SCIRXDA	EPWM4B	GPIO7	79	59	GPIO27	HRCAP2	EQEP2S	SPISTEB
ADCSOCAO	Rsvd	EPWM5A	GPIO8	78	58	GPIO26	ECAP3	EQEP2I	SPICLKB
ECAP3	SCITXDB	EPWM5B	GPIO9	77	57	NC			
ADCSOCBO	Rsvd	EPWM6A	GPIO10	76	56	RESET#			
ECAP1	SCIRXDB	EPWM6B	GPIO11	75	55	GPIO24	ECAP1	EQEP2A	SPISIMOB
			NC	74	54	GPIO25	ECAP2	EQEP2B	SPISOMIB
			NC	73	53	GPIO52	EQEP1S	MCLKXA	TZ3
			DAC3	72	52	GPIO53	EQEP1I	MFSXA	Rsvd
			DAC4	71	51	GPIO56	SPICLKA	EQEP2I	HRCAP3

III.6 Hardware in the loop

Hardware-in-the-loop (HIL) simulation is used to evaluate and validate the performance of control systems in a realistic simulated environment. HIL systems integrate real-time

environment simulation with physical hardware components, known as the device under test (DUT), or simulated models.

During HIL testing, the real-time simulation platform replicates the operational environment with which the control system or DUT would normally interact. This enables detailed testing under conditions that closely mimic real-world scenarios. The platform provides real-time inputs to the control system or DUT and receives outputs.

This allows the system's responses to be tested and monitored dynamically [41].

III.7 Conclusion

In this chapter, we validated the LAUNCHXL-F28069M development board, used to implement sliding mode control applied to a dual-fed asynchronous machine (DFIG), this architecture was integrated into a wind energy conversion system, making it possible to efficiently exploit the energy produced by a variable-speed turbine, the control applied by the card produced satisfactory results in terms of dynamic response, response time reduction and disturbance rejection, validating the feasibility of the approach for renewable energy conversion applications.

Our work aimed to develop and implement a control algorithm capable of ensuring high performance and the production of high-quality electrical energy by a wind energy conversion System (WECS). The control strategy employed was the sliding mode control technique.

In the first chapter, we developed a comprehensive mathematical and simulation model of the WECS. We analyzed the system's dynamic behavior and the interactions between its components under various wind conditions.

The second chapter focused on the design and implementation of a sliding mode control strategy for regulating both active and reactive power within the WECS. This approach demonstrated the system's robustness across a range of operating scenarios.

In the third chapter, we presented an overview of the digital signal processing (DSP) card and explained its operational principles. We examined its internal mechanisms and processing architecture before introducing the Processor-in-the-Loop (PIL) simulation approach. Finally, we deployed the control algorithm onto the DSP card and performed real-time simulations to validate the proposed control strategy which ultimately gave very good results compared to the matlab results.

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Abstract:

Today, the importance of implementing non-linear controls on DFIG based wind energy conversion systems has prompted specialists to carry out extensive research to develop control maps used to validate the controls applied by simulation in order to actually control these complex systems. The objective of this work is to develop and implement a non-linear control on a DFIG based wind energy conversion system in order to validate that the control system works optimally, maintaining the stability, safety, efficiency and performance of the wind turbine under varying conditions.

Key words:

Wind energy - Turbine - DFIG - bidirectional converter - Modelling - Control - Sliding Mode.

Résumé :

Aujourd'hui, l'importance de l'implémentation des commandes non linéaires sur les systèmes de conversion de l'énergie éolienne basés sur une MADA a incité les spécialistes à mener des recherches approfondies pour développer des cartes de commande utilisées pour valider les commandes appliquées par simulation afin de contrôler réellement ces systèmes complexes. L'objectif de ce travail est de développer et d'implémenter une commande non linéaire sur un système de conversion de l'énergie éolienne basé sur MADA afin de valider que le système de contrôle fonctionne de manière optimale, en maintenant la stabilité, la sécurité, l'efficacité et les performances de l'éolienne dans des conditions variables.

Mots clés:

Energie éolienne - Turbine - MADA - Convertisseurs bidirectionnels - Modélisation - Commande - Mode Glissant.

ملخص:

اليوم، دفعت أهمية تطبيق اللوازم غير الخطية على أنظمة تحويل طاقة الرياح القائمة على آلة غير متزامنة ثنائية التغذية المتخصصين إلى إجراء بحوث مكثفة لتطوير بطاقات التحكم المستخدمة للتحقق من صحة اللوازم المطبقة عن طريق المحاكاة من أجل التحكم الفعلي في هذه الأنظمة المعقدة. والهدف من هذا العمل هو تطوير وتنفيذ لوازم غير خطية على نظام تحويل طاقة الرياح القائم على آلة غير متزامنة ثنائية التغذية من أجل التحقق من أن نظام التحكم يعمل على النحو الأمثل، والحفاظ على استقرار وسلامة وكفاءة وأداء توربينات الرياح في ظل ظروف متغيرة.

كلمات مفتاحية:

طاقة هوائية - ناعورة - مكنة لزامية ثنائية التغذية - محول ثنائية النّجاه - نمذجة - تحكم - بالزلزلة.