A New High-Frequency Injection Method for Sensorless Control of Doubly-Fed Induction Machines

DISSERTATION

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ABSTRACT

This research introduces a new method to solve the sensorless control problem for a grid-connected Doubly-Fed Induction Machine (DFIM). The proposed method is based on high-frequency signal injection and the fact that the rotor of a DFIM can be seen as a rotating secondary of an induction transformer.

The currently used sensorless techniques, besides being parameter sensitive, fail during fault ride-through conditions. This fact makes these techniques almost irrelevant for wind turbines applications since, due to new grid codes and regulations, wind turbines are required to stay connected during a fault. Hence, even if the machine in the wind turbine is equipped with a sensorless control algorithm, an encoder (or resolver) is still needed in order to be able to control the machine during fault ride-through conditions.

The proposed sensorless algorithm works by applying a high-frequency voltage to the rotor and measuring the produced high-frequency voltage on the stator. The high-frequency voltage applied to the rotor is independent of, and superimposed on, the control voltage applied to the rotor. It will be shown that the rotor position information is encoded in the phase of the space vector of the high-frequency voltage measured at the stator; hence, the proposed sensorless method is implemented by filtering out the
high-frequency voltage at the stator, and determining the phase of its space vector. This results in a relatively easy implementation.

The proposed sensorless method, besides being parameter independent, remains fully functional during grid faults. Moreover, as opposed to other high frequency injection methods, the proposed method does not require rotor saliency; hence, it is not necessary to induce saturation in the machine.

The mathematical principle of the proposed technique and its implementation are presented. Computer simulations and experimental results are also included for verification.
Dedicated to my family and educators...
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“In God we trust; everybody else must bring data.”

W. Edwards Deming

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TABLE OF CONTENTS

Abstract ............................................................................................................................... ii

Acknowledgments .......................................................................................................... v

Vita ..................................................................................................................................... vii

Table of Contents ............................................................................................................. ix

List of Tables .................................................................................................................... xii

List of Figures ................................................................................................................... xiii

Chapter 1: Introduction ...................................................................................................... 1

1.1 Renewable Energy ....................................................................................................... 4

1.2 Main Generator Topologies used for Wind Power ...................................................... 8

1.3 Grid Connection Codes for Wind Turbines ................................................................. 13

1.4 Sensorless Control of DFIG ..................................................................................... 20

1.5 Contributions of this Research Work ......................................................................... 22

1.6 Conclusions ............................................................................................................... 24

1.7 References ............................................................................................................... 25
Chapter 5: Comparison between currently used Sensorless Method and the newly Proposed Algorithm ........................................................................................................... 86

5.1 Performance during Fault-Ride Through (FRT) .................................................. 87

5.2 Sensitivity against Parameter Variations .............................................................. 100

5.3 Conclusions ........................................................................................................ 101

5.4 References ......................................................................................................... 102

Chapter 6: Conclusions and Future Work ................................................................. 103

6.1 Dissertation Conclusions ................................................................................... 103

6.2 Future Work ....................................................................................................... 105

Bibliography ............................................................................................................. 106
## LIST OF TABLES

<table>
<thead>
<tr>
<th>Table</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Table 1</td>
<td>Performance of MRAS observers</td>
<td>42</td>
</tr>
<tr>
<td>Table 2</td>
<td>Parameters used for Simulations and Experiments</td>
<td>78</td>
</tr>
<tr>
<td>Table 3</td>
<td>Parameters used for Simulations and Experiments</td>
<td>94</td>
</tr>
</tbody>
</table>
LIST OF FIGURES

Fig. 1.1 Global renewable power capacity excluding hydro ........................................ 2
Fig. 1.2 Biofueled world ........................................................................................................ 4
Fig. 1.3 Photovoltaic world ................................................................................................. 5
Fig. 1.4 Current-voltage wcharacteristics of a PV ............................................................ 7
Fig. 1.5 MPPT for Wind Turbine ........................................................................................ 9
Fig. 1.6 Scheme of a Wind Turbine equipped with a PMSG. ........................................ 11
Fig. 1.7 Scheme of a Wind Turbine equipped with a DFIG ........................................... 12
Fig. 1.8 Speed-torque characteristic of the DFIG ............................................................ 13
Fig. 1.9 Typical low voltage ride-through characteristic.................................................. 18
Fig. 1.10 Low voltage ride-through requirements in different countries ....................... 19
Fig. 2.1 Estimation and control of torque angle ............................................................... 30
Fig. 2.2 Phasor diagram of torque angle control ............................................................... 31
Fig. 2.3 Relationship between different angles in Morel’s Estimator ......................... 33
Fig. 2.4 Scheme of general MRAS Observer. ................................................................. 35
Fig. 2.5 Stator flux MRAS Observer .................................................................................. 36
Fig. 2.6 Rotor flux MRAS Observer .................................................................................. 38
Fig. 2.7 Rotor Current MRAS Observer .......................................................................... 39
Fig. 3.1 Two-phase winding system in stator coordinates ................................................. 47
Fig. 3.2 Two-phase winding system, magnet rotor aligned ............................................. 47
Fig. 3.3 Two-phase winding system, magnet rotor displaced ......................................... 49
Fig. 3.4 PM synchronous machine .............................................................................. 49
Fig. 3.5 Space vector diagram .................................................................................... 51
Fig. 3.6 Signal flow graph of sensorless scheme ............................................................ 53
Fig. 3.7 Initial rotor angle and polarity detection ............................................................ 54
Fig. 3.8 Geometric loci of space-vectors ..................................................................... 56
Fig. 4.1 Representation of a 3-ph DFIM ...................................................................... 60
Fig. 4.2 Phase relation between the stator and rotor voltages ....................................... 61
Fig. 4.3 Representation of a 3-ph DFIM in which the rotor is rotated by -35 deg .......... 61
Fig. 4.4 Phase relation between the stator and rotor voltages of previous figure ......... 61
Fig. 4.5 Pictorial representation of a DFIM ................................................................. 63
Fig. 4.6 Per phase equivalent circuit of a grid-connected DFIG .................................. 66
Fig. 4.7 Block diagram of proposed observer .............................................................. 69
Fig. 4.8 Frequency spectrum of proposed sensorless control algorithm ....................... 71
Fig. 4.9 Block diagram of complete system for small-signal analysis ............................. 72
Fig. 4.10 Block diagram of three-phase PLL ............................................................... 72
Fig. 4.11 Linearized model of three-phase PLL ........................................................... 73
Fig. 4.12 Configuration to simulate and test experimentally the proposed method ...... 79
Fig. 4.13 Simulation results at synchronous speed ...................................................... 79
Fig. 4.14 Experimental set-up used to test the proposed algorithm........................................ 80
Fig. 4.15 Experimental results at synchronous speed............................................................. 81
Fig. 4.16 Rotor currents at synchronous speed....................................................................... 82
Fig. 4.17 Experimental results showing the dynamic response.............................................. 83
Fig. 5.1 Configuration of the DFIG wind turbine system....................................................... 88
Fig. 5.2 Rotor-side equivalent circuit. .................................................................................... 88
Fig. 5.3 Simulation results of a 2-MW DFIG during a grid fault........................................... 90
Fig. 5.4 Block diagram of a typical MRAS observer .............................................................. 90
Fig. 5.5 Configuration to simulate and test experimentally the proposed method................. 91
Fig. 5.6 Simulation results for a 2-ph to ground fault with SFMO ........................................ 92
Fig. 5.7 Simulation results for a 2-ph to ground fault with proposed method ....................... 96
Fig. 5.8 Simulation results for a 3-ph to ground fault with proposed method ....................... 97
Fig. 5.9 Configuration to simulate and test experimentally the proposed method................. 98
Fig. 5.10 Experimental results for a 3-ph to ground fault with proposed method.................... 99
Fig. 5.11 Simulation when $L_s$ increases 10% in a SFMO.................................................. 100
Fig. 5.12 Simulation when $L_m$ increases 10% in a SFMO .................................................. 100
CHAPTER 1

INTRODUCTION

There has been an enormous increase in the global demand for energy in recent years as a result of industrial development and population growth. As a consequence, the rise in consumption of traditional fossil fuels has led to many serious problems, such as pollution, global warming, the shortfall of traditional fossil energy sources, and energy insecurity. These factors are driving the development of renewable energy technologies, which are considered as an essential part of a well-balanced energy portfolio able to tackle the aforementioned situation [1]. Wind power, a type of renewable energy, is thought to be the most promising alternative energy in the near future. In recent years, wind power is the fastest growing renewable source of electrical energy [2]. Fig. 1.1 shows the increase in different sources of renewable energy from 2004 to 2010.

Although wind power currently only provides about 3% of European electricity and 2% of the United States’ electrical energy demands, it is reasonable to expect a high penetration of wind power into the existing power system in the near future, e.g., by 2030 [2]. For instance, the European Wind Energy Association (EWEA) has set a target to satisfy 23% of European electricity needs with wind by 2030 [3]. In the United States, the Department of Energy (DOE) and the American Wind Energy Association (AWEA)
examined the feasibility of providing 20% of the nation’s electricity from wind by 2030, and the consensus is that this scenario is feasible [4]. In fact, some European countries have already achieved high levels of wind power penetration. For instance, in 2007, wind power accounted for approximately 19% of electricity production in Denmark, 9% in Spain and Portugal, and 6% in Germany and the Republic of Ireland [1].

The increased penetration of wind energy into the power system over the last decade has led to serious concern about its influence on the dynamic behavior of the power system. It has resulted in the power system operators revising the grid codes in several countries [5]. Several years ago, most grid codes did not require wind turbines to support the power system during a grid disturbance—wind turbines were only required to be disconnected from the grid when an abnormal grid voltage was detected.

Due to wind power’s recent increased capacity, a sudden high loss of power during grid faults resulting from wind turbine disconnection could occur, which may generate control
problems of frequency and voltage in the system, with a worst case scenario of system collapse [6]. Basically, for wind power, new grid codes require an operational behavior more similar to that of conventional generation capacity, and more responsibility in network. Among these grid codes, two main issues are of special concern for engineers in the area of Energy and Power: a) active and reactive power control in normal conditions, and b) low voltage ride-through (LVRT) capability during grid faults, or more succinctly, fault-ride through (FRT) capability.

Besides the progress made in the creation of adequate grid codes for the proper utilization of wind energy, significant improvement has been achieved in the design and implementation of robust energy conversion systems that transform wind energy in an efficient way. The two main energy conversion systems currently in the market are the Permanent Magnet Synchronous Generator (PMSG) and the Doubly-Fed Induction Generator (DFIG). The main contribution of this dissertation resides on the development of an algorithm that allows the DFIG to operate successfully without the need of a position sensor, even during fault conditions, effectively increasing the robustness of the overall system. These techniques are called sensorless control algorithms and, even though many have been proposed and studied, none has been proposed for the specific challenges presented by FRT conditions [8]. The main contribution of this dissertation lies on the development of a sensorless algorithm able to work during FRT conditions [8].

This chapter will provide an overview of different renewable energy sources, with special emphasis on wind power, followed by a general explanation of the benefits and disadvantages of the main topologies used to extract energy from the wind. The issue of
LVRT and grid operation codes for wind generation will be covered since they are one of the main causes for the need of the proposed algorithm. The chapter will end with an overview of the rest of the dissertation.

1.1 RENEWABLE ENERGY

Renewable energy is derived from natural processes that are replenished constantly. In its various forms, it derives directly from the sun, or from heat generated deep within the earth. Included in the definition is electricity and heat generated from solar, wind, ocean, hydropower, biomass, geothermal resources, and biofuels and hydrogen derived from renewable resources [9]. Climate change concerns, coupled with high oil prices, peak oil, and increasing government support, are driving increasing renewable energy legislation,
A world powered entirely by photovoltaics (solar power) would require 6% of the available land and 12% of the rainfall [12].

The increase in interest and investment in renewable energy has been accompanied by a surge in research interest in the ways to exploit them. Given the state of the art in energy conversion technology, bio-fuels, solar, and wind energy have been identified as a new important batch of renewable resources with great potential for energy extraction.

1.1.1 BIOFUELS

Biomass (plant material) can be converted directly into liquid fuels, or "biofuels," to help meet transportation fuel needs. The two most common types of biofuels in use today are ethanol and biodiesel. Ethanol is an alcohol, the same as in beer and wine (although ethanol used as a fuel is modified to make it undrinkable). It is most commonly made by fermenting any biomass high in carbohydrates through a process similar to beer brewing.
Today, ethanol is made from starches and sugars, but research work is in place in order to develop technology to allow it to be made from cellulose and hemicellulose, the fibrous material that makes up the bulk of most plant matter [11].

A simulation study published in 2009 [12] showed that, in its current form, biofuels utilize too much land and water resources when compared to other forms of renewable energy, like solar power. Fig. 1.2 shows the results of supplying all the electricity, transportation and heating demands for the whole world from switchgrass. Switchgrass was picked because it seemed a fair compromise among biofuels. It has a good shot at near-term economic viability, because it can provide more net energy for a given amount of carbon emissions than corn ethanol. The results show that, even though this scenario would result in zero carbon emissions, the amount of land (193% of the land available on the planet) and water (173% of annual rainfall) required would make this scenario impractical. On the other hand, when the world’s energy needs are supplied by photovoltaics (solar power) only 6% and 12% of land and rainfall are needed, as shown in Fig. 1.3. The scenario based on photovoltaics is achieved without any emissions of CO2.

1.1.2 SOLAR POWER

More energy from the sun falls on the earth in one hour than is used by everyone in the world in one year [13]. This has motivated research in the exploitation of solar energy.

Solar power is the conversion of sunlight into electricity, either directly using photovoltaics (PV), or indirectly using concentrated solar power (CSP). Concentrated solar
power systems use lenses or mirrors and tracking systems to focus a large area of sunlight into a small beam. Photovoltaic cells are normally connected in groups of parallel strings, called modules. The cells (or modules) convert light into direct current (DC) electric power using the photoelectric effect [15]. The power produced fluctuates with the intensity of the irradiated light and usually requires conversion to certain desired voltages or alternating current (AC), which requires the use of inverters [16].

Solar cells have a complex relationship between solar irradiation, temperature, and total resistance that produces a non-linear output efficiency known as the I-V curve. In order to obtain maximum power for any given environmental conditions, a controller samples the output of the cells and applies the proper resistance (load); this technique is normally called Maximum Power Point Tracking (MPPT). Fig. 1.4 shows different operating curves for a conventional solar cell as a function of the irradiation levels. The maximum points of operation are marked by an asterisk (*), and are located at different points in the I-V curves shown in Fig. 1.4.
1.1.3 WIND POWER

Wind power refers to the conversion of wind energy into a useful form of energy, such as using wind turbines to make electricity. The total wind energy $E$ flowing through an area $A$ during a time $t$ is given by

$$E = \frac{1}{2} m v^2 = \frac{1}{2} (Avt \rho) v^2 = \frac{1}{2} At \rho v^3,$$

(1.1)

where $v$ is the wind speed and $\rho$ is the wind density. Therefore, the available wind power, $P$, becomes $P = \frac{1}{2} A \rho v^3$. Eq. (1.1) represents the total kinetic wind energy available, but extracting all the energy from the wind would require the wind velocity to be zero after the energy conversion process. This is not possible for wind turbines, with Betz’s law establishing the maximum theoretical limit of 59.3%.

Wind power does not produce CO$_2$, is renewable, plentiful, clean, and widely distributed. Even though it is an intermittent source, it seldom creates problems when it supplies less than 20% of electricity demand [17].

1.2 MAIN GENERATOR TOPOLOGIES USED FOR WIND POWER

Nowadays, wind turbines (WT) on the market mix and match a variety of innovative concepts with proven technologies for both generators and power electronics [18]. There are different types of WT with various control philosophies, forms of operation, and connections to the grid. In the early stage of wind power development, most wind farms...
were equipped with fixed-speed wind turbines and induction generators. Since such wind
generators can only operate at a constant speed, the power efficiency is fairly low for most
wind speeds [19].

The fixed-speed turbines employ induction generators (squirrel cage) directly connected
to the electrical grid. They operate within a narrow range of speed slightly higher than the
synchronous speed, so they are called fixed speed wind turbines. This technology shows
advantages when compared to a variable speed device, including simplicity, mechanical
strength, cost, and ease of operation. On the other hand, they have less control flexibility
and very low power efficiency.

The main trend of modern wind turbines/wind farms is clearly the variable-speed
operation and a grid connection through a power converter interface. There are a lot of
advantages to this type of generator when it is compared to fixed-speed machines, such as
the reduction of mechanical stress, the improvement of power quality, and the increase in

![Image](image-url)  
**Fig. 1.5** Maximum power point tracking (MPPT) curve for a wind turbine [20].
the overall system efficiency. Those advantages are achieved through a higher cost per MW [19, 21]. Fig. 1.5 shows an example of the tracking characteristic for a wind turbine based on wind speed. Each turbine power characteristic, which is based on a particular value of wind speed, shows the relationship between the output power of the turbine and the rotational speed of the shaft. Hence, in order to extract the maximum output power from the turbine, its shaft speed should be a function of the wind speed. The tracking characteristic is the curve joining all the maxima of the turbine power characteristic curves; hence, it coincides with the MPPT curve.

There are two main variable-speed wind turbine concepts in the market today. One of them is the variable-speed wind turbine concept with partial-scale power converter, known as the doubly fed induction generator (DFIG) concept. The other is the variable-speed wind turbine concept with full-scale power converter and synchronous generator (PMSG). These two variable-speed wind turbine concepts compete against each other on the market, with their more or less weak and strong features [18]. The proposed sensorless algorithm is intended for DFIGs since this is the generator present in the majority of the high power wind turbines in the market.

1.2.1 PERMANENT MAGNET SYNCHRONOUS GENERATORS (PMSG)

Fig. 1.6 shows a diagram of a wind turbine based on a PMSG. As seen in Fig. 1.6, all the energy produced by the generator reaches the grid through two power converters, connected “back-to-back”. Between the two converters a dc-link capacitor is placed as
energy storage, in order to keep the voltage variations (or ripple) in the dc-link voltage small [22]. The back-to-back converter is required in order to convert the voltage and frequency of the energy provided by the PMSG to levels compatible with the grid. This is a consequence of allowing the wind turbine to rotate at different speeds in order to maximize the power extracted from the wind. Consequently, the back-to-back converters have to be rated at the power of the PMSG, which increases the initial cost and reduces the overall efficiency of this wind turbine system, as compared to wind turbine systems using DFIGs. On the other hand, wind turbines based on PMSGs tend to have lower weight and maintenance costs since they lack brushes and can be direct-driven (without a gear box), as opposed to DFIGs.

1.2.2 DOUBLY-FED INDUCTION GENERATORS (DFIG)

The most widely used generator type for units above 1 MW is the doubly fed induction
machine [19]. At present, the basic advantage of the DFIG concept is that only a percentage of power generated in the generator has to pass through the power converter. This is typically only 20–30% compared to full power (100%) for a synchronous generator-based wind turbine concept, and thus it has a substantial cost advantage compared to the conversion of full power.

For variable-speed systems with limited variable-speed range, e.g. ±30% of synchronous speed, the DFIG can be an interesting solution [21]. As mentioned earlier, the power electronics converter only has to handle a fraction (20–30%) of the total power [22]. This means that the losses in the power electronics converter can be reduced compared to a system where the converter has to handle the total power. In addition, the cost of the converter is decreased. The stator circuit of the DFIG is connected to the grid, while the rotor circuit is connected to a converter via slip rings, as seen in Fig. 1.7.

As seen from Fig. 1.7, the rotor side of the machine is connected to the grid through a

![Diagram of Wind Turbine with DFIG](image-url)
back-to-back power converter. With the machine-side converter, it is possible to control the torque or the speed of the DFIG and also the power factor at the stator terminals, while the main objective for the grid-side converter is to keep the dc-link voltage constant. The speed–torque characteristics of the DFIG system can be seen in Fig. 1.8 [24]. As also seen in the figure, the DFIG can operate both in motor and generator operation with a rotor-speed range of $\pm \Delta \omega_r^{\text{max}}$ around the synchronous speed, $\omega_1$.

1.3 GRID CONNECTION CODES FOR WIND TURBINES

Grid codes are documents that contain technical procedures and responsibilities regarding power system planning, operation, maintenance, and protection. Every generator (like a wind power generator) that successfully connects to the grid has to follow this set of
requirements. The major requirements of typical grid codes for operation and grid connection of wind turbines are summarized as follows [2]:

1) **Voltage operating range:** The wind turbines are required to operate within typical grid voltage variations.

2) **Frequency operating range:** The wind turbines are required to operate within typical grid frequency variations.

3) **Active power control:** Several grid codes require wind farms to provide active power control in order to ensure a stable frequency in the system and to prevent overloading of lines, etc. Also, wind turbines are required to respond with a ramp rate in the desired range.

4) **Frequency control:** Several grid codes require wind farms to provide frequency regulation capability to help maintain the desired network frequency.

5) **Voltage control:** Grid codes require that individual wind turbines control their own terminal voltage to a constant value by means of an automatic voltage regulator.

6) **Reactive power control:** The wind farms are required to provide dynamic reactive power control capability to maintain the reactive power balance and the power factor in the desired range.

7) **Low voltage ride through (LVRT):** In the event of a voltage sag, the wind turbines are required to remain connected for a specific amount of time before being allowed to disconnect. In addition, some utilities require that the wind turbines help support grid voltage during faults.
8) **High voltage ride through (HVRT):** In the event the voltage goes above its upper limit value, the wind turbines should be capable to stay on line for a given length of time.

9) **Power quality:** Wind farms are required to provide the electric power with a desired quality, e.g., maintaining constant voltage or voltage fluctuations in the desired range, maintaining voltage/current harmonics in the desired range, etc.

10) **Wind farm modeling and verification:** Some grid codes require wind farm owners/developers to provide models and system data, to enable the system operator to investigate by simulations the interaction between the wind farm and the power system. They also require installation of monitoring equipment to verify the actual behavior of the wind farm during faults, and to check the model.

11) **Communications and external control:** The wind farm operators are required to provide signals corresponding to a number of parameters important for the system operator to enable proper operation of the power system. Moreover, it must be possible to connect and disconnect the wind turbines remotely.

### 1.3.1 LOW VOLTAGE RIDE-THROUGH IN WIND TURBINES WITH DFIGS

In the past, most national grid codes and standards did not require wind turbines to support the power system during a disturbance. For example, during a grid fault or sudden drop in frequency, wind turbines were tripped off the system. However, as the wind power penetration continues to increase, the interaction between the wind turbines and the power
system has become more important. This is because, when all wind turbines would be
disconnected in case of a grid failure, these renewable generators will, unlike conventional
power plants, not be able to support the voltage and the frequency of the grid during and
immediately following the grid failure. This would cause major problems for the system’s
stability [2, 5, 25].

Therefore, wind farms will have to continue to operate during system disturbances and
support the network voltage and frequency. Network design codes are now being revised to
reflect this new requirement. A special focus in this requirement is drawn to both the fault
ride-through (FRT) capability and the grid support capability [5, 6, 18, 19, 21]. Fault
ride-through capability addresses mainly the design of the wind turbine controller in such a
way that the wind turbine is able to remain connected to the network during grid faults (e.g.
short circuit faults). Grid support capability represents the wind turbine’s ability to assist
the power system by supplying ancillary services, i.e. such as supplying reactive power, in
order to help the grid voltage recovery during and just after the clearance of grid faults.

Due to the partial-scale power converter, wind turbines based on the DFIG are very
sensitive to grid disturbances, especially to voltage dips during grid faults. Faults in the
power system, even far away from the location of the turbine, can cause a voltage dip at the
connection point of the wind turbine. The abrupt drop of the grid voltage will cause
over-current in the rotor windings and over-voltage in the DC bus of the power converters.
Without any protection, this will certainly lead to the destruction of the converters. In
addition, it will also cause over-speeding of the wind turbine, which will threaten the safe
operation of the turbine [26 - 28]. Thus, a significant amount of research has been carried
out on the LVRT ability of DFIG wind turbines under the grid fault [6, 19, 25, 27]. These LVRT strategies can be divided into two main types: the active method by improving control strategies and the passive scheme with additional hardware protective devices.

Low voltage ride-through is seen to be particularly important in terms of maintaining voltage stability, especially when there is a high local concentration of wind generation. Premature tripping of numerous wind generators due to local disturbances can further risk the stability of the system, contributing to amplification of the effect of the disturbance [29]. This is particularly important in the case of weak systems, where the wind generator should help to provide voltage support and contribute to the load at the end of the line.

Another related issue is the possibility of properly estimating the rotor position during grid-faults, which is needed for sensorless control. This dissertation presents a technique that is able to properly estimate the rotor position during several conditions, grid-faults included.

**LVRT CHARACTERISTIC**

A typical LVRT characteristic is shown in Fig. 1.9. The characteristic is more or less defined by a minimum voltage throughout the duration of the fault followed by a ramping up to nominal level as the voltage recovers. The width of this minimum is dictated by the length of time to normally clear a transmission level fault (10 to 20 cycles), which then ramps up to the normal operating range with a given slope. Although the width and magnitude of the minimum are dictated by protection technologies and the location and type of fault, the slope of the recovery likely depends on the strength of the interconnection.
and reactive power support, whereby stronger systems could afford a much steeper increase and thus minimize the ride-through requirements of the generators.

The functional operation of LVRT is based on the comparison of the characteristic with that of the terminal voltage. Essentially, once the voltage dips outside of the normal operating range, the control system then compares the recovery of the fault with that of the minimum voltage level as given by the LVRT characteristic. In the case in which the voltage falls below that of the LVRT characteristic, the generator may trip, however, anywhere above the minimum, the generator must remain connected and support the grid [29].

Minimum LVRT requirements were first established by Germany [30]. Later, other countries have adopted similar rules. Fig. 1.10 illustrates some LVRT characteristics adopted in different countries [30]. In these curves, it can be seen that each country adopts a different approach for minimum voltage during the event, protection system time, and voltage recovery. Germany and the UK grid operators issue LVRT curves with hard
constraints on minimum voltage: WT connected there have to withstand zero voltage for a period of time. USA/Ireland/Canada, Spain and Italy are more severe on the protection time duration. In six different LVRT curves presented, there are four different profiles for
the voltage recovery period. Even for the three countries with similar profiles, slopes and recovery times are different from each other [2, 30].

### 1.4 SENSORLESS CONTROL OF DFIG

As mentioned before, rotor-controlled Doubly-Fed Induction Machines (DFIMs) are widely used as generators in wind power systems in mega-watt ratings due to several advantages. The main one is the possibility of using power converters that are a fraction of the total power of the system. Since the reliability of the system increases with fewer components, significant effort has been made to solve the sensorless control problem for DFIMs. The efforts can be broadly divided into two main groups. The first group is open-loop rotor-position estimators [31-33], in which the rotor position is directly estimated from the measured voltages and currents by reference frame transformation. Unfortunately, open-loop estimators are highly sensitive to the machine parameters, and the accuracy of estimation is not guaranteed [34]. The second group is based on adaptive control theory [34-38]. Within this group, rotor-flux based model reference adaptive system (MRAS) observers [37, 38] have been widely studied. The accuracy of the rotor position estimation using MRAS, compared to open-loop estimators, is less influenced by variations of the machine parameters. However, both techniques lose accuracy when the rotor speed approaches synchronous speed, because the back-EMF in the rotor windings approaches zero around synchronous speed, analogous to the case of Singly-Fed Induction Machines running at zero speed. Hence, methods based on rotor flux integration cannot
give a satisfactory result [31, 39]. As a way to overcome this issue, several researchers
[40-41] have developed observers in which the stator-flux linkage, which is considered
very stable for grid-connected machines, is used in the estimation process, as opposed to
the rotor flux linkage. These methods are susceptible to machine parameter variations,

hence on-line estimators shall be used [42]. However, they are immune to the

aforementioned problem at synchronous speed due to the fact that the grid to which the

DFIGs are connected remains relatively stable, which is true in most cases. Nevertheless,

this assumption does not hold during fault ride through (FRT) conditions. In these

situations, the highly distorted voltages and currents introduce significant disturbances to

the observers, which cannot keep track of the reference quantities utilized to determine the

rotor position. As a consequence, these sensorless methods fail under FRT conditions.

During the last decade, a clever technique based on high frequency signal injection (HFI)

has emerged as a possible solution to the sensorless problem at zero speed for SFIMs [43].

Literature review indicates that the HFI-based method has been investigated only once for

DFIG applications [44]. As expected, the accuracy of the results is not dependent on

parameter uncertainties, and the method works at synchronous speed. Unfortunately, a

large amount of computational power was needed since a Fourier Transform based

algorithm was used. In contrast to the approach in [44], this dissertation presents a new HFI

method with a relatively simple algorithm to solve the sensorless control problem of

grid-connected DFIMs.
1.5 CONTRIBUTIONS OF THIS RESEARCH WORK

The main contribution of this dissertation is the development of a new sensorless control algorithm for grid-connected DFIGs. In the proposed algorithm, a high-frequency signal is injected to the rotor windings to determine the rotor position, while the rotor saliency information is not needed. The advantages of the proposed method include being immune to parameter variations, very good performance at synchronous speed and during FRT conditions, and ease of implementation.

As mentioned previously, due to current grid code regulations, DFIGs have to remain connected during grid faults. This requirement introduces new challenges, which in this dissertation have been categorized as two independent sub-problems. The first sub-problem would be the actual control of the machine. One of the objectives in this sub-problem is to be able to control the rotor currents during the fault, giving a limited DC bus voltage. Another related objective could be to protect the rotor-side converter from the high-currents that tend to appear during a grid-fault. Several papers have been written on these issues, and it will be explained in more detail in Chapter 5. A second sub-problem, which is separate from the first, is the possibility of properly estimating the rotor position during grid-faults, which is needed for sensorless control. This dissertation presents a technique that is able to properly estimate the rotor position during several conditions, grid-faults included. In that sense, this dissertation is not presenting a new control technique, but a new estimator.
Several papers dealing with the control of the rotor currents of a DFIM during FRT conditions have been written [45, 46], but the issue of rotor position estimation during FRT conditions has not been studied yet. The sensorless technique proposed in this dissertation solves the rotor position estimation problem during FRT and other conditions.

1.5.1 DISSENTATION OVERVIEW

This dissertation is structured in a way that explains the need, development, and testing of a new sensorless control algorithm. In this sense, Chapter 1 covered the current state of the art of important sources of renewable energy and how wind power fits in this picture. Also, this chapter provided a short introduction to the two main technologies used for wind power generation and how the newly developed grid codes create new challenges for DFIGs. In the last part of this chapter, the currently available sensorless algorithms for DFIGs were reviewed with special attention given to how the issue of sensorless control during FRT conditions has not been studied.

Chapter 2 will expand Section 1.4 by describing in more detail the different techniques of sensorless control for DFIMs with special attention to the current state of the art. Chapter 3 will introduce a technique (high-frequency injection) that serves as the basis for the sensorless technique proposed in this dissertation. Even though high-frequency injection has been widely used as a sensorless control method for other machines, this method has not been fully exploited in DFIMs. Hence, a study of high-frequency injection for DFIMs represents another important contribution of this dissertation.
Chapter 4 develops the proposed high-frequency injection method. An intuitive explanation is given, followed by a more rigorous mathematical development. Then a method for its actual implementation is proposed followed by simulation and experimental results. The small-signal modeling of the proposed implementation is included. This model becomes the foundation for the design and selection of the parameters of the proposed sensorless estimator.

Since performance during FRT conditions has been identified as one of the main advantages of the proposed sensorless method, Chapter 5 is devoted to this issue. The specific issue of the behavior of a DFIG during LVRT is studied in detail, followed by how currently used sensorless algorithms behave during these FRT conditions. This is followed by a rigorous experimental testing of the proposed algorithm during FRT conditions.

Finally, Chapter 6 concludes the dissertation by making final remarks and some suggestions about future research directions to be followed.

1.6 CONCLUSIONS

This chapter served as an introduction to this dissertation by explaining the current state of the art in wind energy conversion. This chapter also showed that, in view of the newly proposed grid codes for wind turbines, a new sensorless control method for grid-connected DFIGs is needed.
1.7 REFERENCES


Introduction


CHAPTER 2

SURVEY OF SENSORLESS CONTROL METHODS FOR DFIMS

This chapter will summarize the state of the art in sensorless control of grid-connected Doubly-Fed Induction Machines (DFIMs). As with many other cases in science and engineering, the development of new sensorless algorithms is the response to new challenges that the current available technology is unable to meet. This will be exemplified in this chapter. Along these lines, the proposed sensorless algorithm, which will be introduced in Chapter 4, is also a response to the new challenges raised by the current grid connection codes during FRT conditions.

This chapter will first cover open-loop estimators, which are the easiest to implement, but are relatively sensitive to parameter variations. A solution to the parameter variation problem is proposed in the so called Model Reference Adaptive System (MRAS) observer. The most salient features of this family of observers will be shown. A commonly used observer from this family will be used in Chapter 5 for comparison to the sensorless method proposed in this dissertation.
2.1 OPEN-LOOP OBSERVERS

Several estimators have been proposed in which, assuming perfect knowledge of the machine parameters, the rotor position angle can be determined. This section is not intended to provide an exhaustive list of all open-loop estimators in the literature, but aims at highlighting key examples that feature special characteristics.

2.1.1 XU’S OBSERVER

Xu et al. [1] presented the first sensorless algorithm for DFIMs. Since DFIMs are normally controlled through the rotor, accurate knowledge of the actual rotor angle becomes necessary for field-orientation control. This fact presents a higher degree of complexity when compared to sensorless control for singly-fed induction machines, since the latter only need the rotor speed information when field-orientation control is desired.

The work presented in [1] is based on the fact that the electromagnetic torque, $T_e$, can be viewed as the result of the interaction between the rotor current, $i_{dq}$, and the airgap flux, $\lambda_{dqm}$, regardless of the selection of the reference frame. That is,

$$T_e = \frac{3}{2} p \left( \lambda_{dqm} i_{dq} - \lambda_{qdr} i_{dr} \right) = \frac{3}{2} p \left| \lambda_{dqm} \right| \left| i_{dq} \right| \sin \delta$$

(2.1)

and

$$T_e = \frac{3}{2} p \left( \lambda_{qmr} i_{dr} - \lambda_{qmr} i_{dr} \right) = \frac{3}{2} p \left| \lambda_{qym} \right| \left| i_{dr} \right| \sin \delta$$

(2.2)
are equivalent. In (2.1) and (2.2), $\delta$ represents the torque angle, and the subscripts \textit{d-q} and \textit{x-y} are used to differentiate the selection of the reference frames tied to the synchronous and rotor circuit frame, respectively. Since (2.1) and (2.2) are equivalent, observation of the magnitudes of flux and current, as well as the electromagnetic torque, are independent of the reference frame selection. The torque angle can be estimated in one reference frame, and then the estimated results can be used in another reference frame. Furthermore, the airgap flux can be estimated from either the stator, rotor, or combined circuit. Therefore, it is possible to choose the most convenient circuit, that of the stator or rotor, in the most convenient reference frame, the rotor or synchronous frame, to compute the magnitudes of the flux and current and the torque angle $\delta$.

In the sensorless control strategy proposed in this method, the rotor currents and voltages are used as the basic input variables to compute the torque angle, because the rotor variables are already available for the purpose of current regulation. Assuming that the reference frame is tied to the rotor axis, the torque angle and the flux magnitude together

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Survey of Sensorless Control Methods for DFIMs

30
with the current magnitude can be estimated by the measured rotor voltages and currents according to the following equations:

\[
\dot{\lambda}_{x_m} = \int (v_{x_r} - r_i i_{x_r}) dt - L_{q_r} i_{y_r},
\]

\[
\dot{\lambda}_{y_m} = \int (v_{y_r} - r_i i_{y_r}) dt - L_{q_r} i_{x_r},
\]

\[
\dot{\lambda}_m = \sqrt{\dot{\lambda}_{x_m}^2 + \dot{\lambda}_{y_m}^2},
\]

\[
i_r = \sqrt{i_{x_m}^2 + i_{y_m}^2}.
\]

(2.3)

Thus, according to (2.2), the sine of the torque angle is

\[
\sin \hat{\delta} = \frac{\dot{\lambda}_{y_m} i_{x_r} - \dot{\lambda}_{x_m} i_{y_r}}{\dot{\lambda}_m i_r}.
\]

(2.4)

In (2.3) and (2.4), \(\hat{\delta}, \dot{\lambda}_{x_m}, \dot{\lambda}_{y_m},\) and \(\dot{\lambda}_m\) are used to indicate that they are the estimated values. Note, however, that \(\sin \hat{\delta}\) alone is not sufficient to uniquely determine the value of \(\hat{\delta}\), since the inverse of a sine function is dual-valued. Interpreting this observation to the machine physics, the information contained in the electromagnetic torque or active power cannot uniquely determine the torque angle. It is evident that additional information is necessary. In effect, by computing a quantity related to the instantaneous reactive power,
cos \delta \ can be found by the following equation:

\[ \cos \hat{\delta} = \frac{\hat{\lambda}_{ym} t_{yr} + \hat{\lambda}_{wm} t_{yr}}{\hat{\lambda}_{im} t_{r}}. \]  \hfill (2.5)

Taking the inverse of \( \sin \hat{\delta} \) and \( \cos \hat{\delta} \) terms, the torque angle \( \hat{\delta} \) is uniquely determined. Similar to the computation of the sine term, the necessary inputs for (2.5) are rotor side variables which are readily available and no additional sensing is needed.

It is interesting to note that by using the rotor voltages and currents, the torque angle \( \delta \) is estimated without any transformation. Nevertheless, the estimated angle \( \hat{\delta} \) can be used in any other reference frame. Using the estimated angle of \( \hat{\delta} \) as the feedback variable, it is possible to force the actual \( \delta \) to follow the desired value. The scheme of the torque angle estimation and control is summarized as shown in Fig. 2.1. In the control block of Fig. 2.1, a voltage controlled oscillator (VCO) with high gain is used to generate a desired slip frequency, so as to converge the actual torque angle to the desired one. It is in this way that the torque and reactive power control of the doubly fed induction machine is properly achieved. The torque angle control scheme can be best illustrated by the phasor diagram shown in Fig. 2.2. Note that as soon as an error in the torque angle is detected by the estimator, for example \( \delta < \hat{\delta} \) as shown in the phasor diagram, the VCO based controller will immediately slow down the slip frequency so that \( \theta_s \) is reduced. As a result, the torque angle increases. Integrating the torque angle estimation and control schemes into the system, the rotor speed can be estimated conveniently.

The main drawback of this method, besides being parameter sensitive, is its inability to work close to synchronous speed. As seen in (2.3), the rotor voltage is required to calculate
the airgap flux, but close to synchronous speed it becomes very low in magnitude and frequency due to lack of magnetic induction.

### 2.1.2 MOREL’S OBSERVER

As opposed to Xu’s algorithm mentioned before, which tries to estimate the flux and electromagnetic torque to calculate the rotor angle, Morel’s method [2] uses quantities from the stator frame to compute the rotor mechanical position. In summary, this method does not use the rotor voltage for the angle computation, it needs only the stator voltages and the rotor and stator currents. This becomes an advantage with respect to Xu’s method at synchronous position, besides the fact that acquiring the rotor voltages presents some challenges in case of a PWM inverter supply. The computation is achieved only with multiplication and division, Park’s transformation, and the stator inductance, which has to be known.

In the explanation that follows, $\theta_s$ and $\theta_r$ refer to the angle between a rotating reference

![Fig. 2.3 Relationship between different angles in Morel’s Estimator [2].](image-url)
frame and the stator and rotor frames axis, respectively. As shown in Fig. 2.3, these two angles are related to the position of the rotor, $\theta_m$, by $\theta_s = \theta_r + \theta_m$. The goal becomes to get the values of $\theta_r$ and $\theta_m$. To determine the electrical angle $\theta_r$, for Park’s rotor transformation, two other angles $\delta$ and $\gamma$ are computed. Indeed, the three rotor currents are a function of the angle $\delta(t)$ for a balanced direct system as in:

$$I_{abc r}(t) = \sqrt{2} \cdot I_s(t) \begin{pmatrix} \cos(\delta) \\ \cos \left( \frac{\delta - 2\pi}{3} \right) \\ \cos \left( \frac{\delta + 2\pi}{3} \right) \end{pmatrix}^T.$$  \hspace{1cm} (2.6)

After applying the Clark Transform to (2.6), the angle $\delta$ can be found from

$$\delta = \tan^{-1} \frac{I_{\beta r}}{I_{\alpha r}}.$$  \hspace{1cm} (2.7)

Determining $\gamma$ requires two steps. First, $\theta_s$ is determined by using (2.8) below:

$$\hat{\dot{\lambda}}_{\alpha s} = \int (v_{\alpha s} - r_s i_{\alpha s}) dt - L_{\alpha s} i_{\alpha s}, \quad \hat{\dot{\lambda}}_{\beta s} = \int (v_{\beta s} - r_s i_{\beta s}) dt - L_{\alpha s} i_{\beta s},$$

$$\theta_s = \tan^{-1} \frac{\hat{\dot{\lambda}}_{\beta s}}{\hat{\dot{\lambda}}_{\alpha s}}.$$  \hspace{1cm} (2.8)

Then, $I_{\theta r}(t)$, the rotor current in a rotating reference frame, is found by solving the equation below for $I_{abc r}^{\theta(s)}$,

$$\hat{\dot{\lambda}}_{abc r} = \begin{bmatrix} L_s \\ L_{sr} \end{bmatrix} \begin{bmatrix} I_{\theta r}^{\theta(s)} \\ I_{abc r} \end{bmatrix},$$  \hspace{1cm} (2.9)

where $[L_s]$ and $[L_{sr}]$ represent the stator and stator-rotor mutual inductance matrices, respectively. After $I_{\theta r}^{\theta(s)}(t)$ is found, $\gamma$ is obtained from

$$\gamma = \tan^{-1} \frac{I_{\beta s}}{I_{\alpha s}}.$$  \hspace{1cm}

Finally, as seen in Fig. 2.3, after $\delta$ and $\gamma$ are obtained, $\theta_m$ can be found.
Even though this method is capable of working at synchronous speed, it is very susceptible to machine parameter variations. The following sensorless algorithms are able to reduce the dependency on machine parameters due to the use of adaptive control techniques.

### 2.2 MRAS-BASED OBSERVERS

This family of methods [3] is based on a technique called Model Reference Adaptive Systems (MRAS). In summary, an MRAS–based estimator is composed of two different models of the system under study. These are, as shown in Fig. 2.4, a Reference Model and an Adaptive Model. The Reference Model does not depend on the variable to be estimated, whereas the Adaptive Model does. In the case of MRAS-based estimators for sensorless control of DFIMs, the variable to be estimated is the rotor mechanical angle, $\theta_r$ (notice that in this section, $\theta_r$ refers to the rotor mechanical angle, as opposed to its meaning in the previous section). The difference between the outputs of these two models creates an error that is used to drive an adjustment block (normally a PI). The output of the adjustment
block is supposed to converge on the true quantity to be estimated. Simulation results of an MRAS-based estimator will be shown in Chapter 5.

### 2.2.1 STATOR FLUX BASED MRAS OBSERVER (SFMO)

The SFMO described in [3] uses the stator flux, $\lambda_s$, as the comparison variable between the two models of Fig. 2.4. As shown in Fig. 2.5, these two models are the machine voltage model, which does not depend on the quantity to be estimated, and the machine current model, which does depend on $\theta_r$.

The Reference and Estimated stator flux is given by the following equations:

\[
\tilde{\lambda}^s_s = \int (\tilde{v}^s_s - R_s \tilde{i}^s_s) dt \tag{2.10}
\]

\[
\hat{\lambda}^s_s = L_s \tilde{i}^s_s + L_m \tilde{i}^{j0^s} e^{j\phi^s}. \tag{2.11}
\]

In (2.10) and (2.11), the rotor current is in the rotor reference frame, while the other quantities are in the stator reference frame, as indicated by the superscripts. As shown in

![Fig. 2.5 Stator flux MRAS observer [3]. In this section, $\theta$ refers to the rotor mechanical angle.](image)
Fig. 2.5, the error is defined as the vector product between the Reference and Estimated stator fluxes, \( \epsilon = \hat{\lambda}_{d_s} \hat{\lambda}_{q_s} - \lambda_{d_s} \hat{\lambda}_{q_s} \).

Fig. 2.5 shows the SFMO implementation. The voltage model is used to obtain \( \lambda_s \) using a band-pass filter as a modified integrator to block the dc components of the measured voltages and currents. Since the stator voltages and currents are at a frequency well above the filter cut-off frequency, there is no deterioration in integral action.

One of the disadvantages of the stator flux MRAS observer is that if \( i_{dr}' = 0 \), the observer fails. This is noticeable from the expansion of (2.11) into its respective d-q component, as shown in (2.12):

\[
\begin{bmatrix}
\hat{\lambda}_{d_s} \\
0
\end{bmatrix} = [L_s] \begin{bmatrix}
i_{d_s}' \\
i_{q_s}'
\end{bmatrix} + [L_m] \begin{bmatrix}
i_{dr}' \\
i_{qr}'
\end{bmatrix} \cdot e^{j(\hat{\theta}_r - \theta_r)}. \quad (2.12)
\]

Since \( i_{d_s}' \) is the only “link” between the error \( (\hat{\theta}_r - \theta_r) \) and the output of the Adaptive block, \( \hat{\lambda}_{d_s} \), there is no way to track the error if \( i_{dr}' = 0 \).

### 2.2.2 Rotor Flux Based MRAS Observer (RFMO)

The RFMO observer is widely used for sensorless control of cage induction machines [1, 4]. However, because DFIGs are excited from the rotor and stator, RFMOs have different small signal models and dynamic performance when sensorless control is considered [3]. Several versions of RFMOs have been written [5], but this dissertation will follow the development in [3] because of its generality.
Similarly to the SFMO, the RFMO uses voltage and current models. In the stator frame, the reference rotor flux is obtained as [6]

\[
\hat{\lambda}_r = \frac{L_r}{L_m} \left[ \hat{\lambda}_m - \sigma L_s i_s \right],
\]

where the leakage coefficient, \( \sigma \), is defined as \( \sigma = (L_s L_r - L_m^2) / (L_s L_r) \). In the stator frame, the rotor is determined using the stator currents

\[
\hat{\lambda}_r^s = L_m \tilde{i}_s^s + L_r \tilde{i}_r^s \cdot e^{j(\hat{\theta} - \theta_0)}.
\]  (2.13)

Transforming (2.13) to a rotating axis aligned with the stator flux yields

\[
\hat{\lambda}_{dpr} = L_m \tilde{i}_{dpr} + L_r \tilde{i}_{dpr} \cdot e^{j(\hat{\theta} - \theta_0)}.
\]

The implementation of the RFMO is shown in Fig. 2.6. It has been shown [3] that if \( \sigma \) or the current \( i_q^r \) is relatively low, then a rotor flux MRAS observer may produce a noisy or unstable estimation of the rotational speed and/or rotor position when \( \dot{\theta}_r = 0 \). When compared to the SFMO, the main disadvantage of the RFMO is that it has a relatively
complex expression for the gain when the system is oriented along the stator flux vector. Therefore, the variable gain required in the PI controller is not simple to implement and is affected by incorrect estimation of the machine parameters.

### 2.2.3 Rotor Current Based MRAS Observer (RCMO)

MRAS observers for sensorless DFIGs can also be implemented using the rotor and stator currents. This has the advantage that \( i_r \) and \( i_s \) are both measured quantities. For the RCMO, the Reference model will be the current measured by the transducer. In stationary reference frame, the flux is obtained by (2.11). Solving for \( i_r \) in (2.11) produces

\[
\tilde{i}_r = \frac{\tilde{\lambda}_r - L_r i_r}{L_m} e^{-j\omega_r t}.
\]

Hence, the Adaptive model is given by replacing \( \omega_r \) by its estimation, \( \hat{\omega}_r \), in (2.14), i.e.,

\[
\hat{i}_r = \frac{\tilde{\lambda}_r - L_r i_r}{L_m} e^{-j\hat{\omega}_r t}.
\]

Finally, the error is the cross product between the estimated and measured currents.

Fig. (2.7) shows the implementation of the proposed MRAS observer. As opposed to the

![Fig. 2.7 Rotor Current MRAS observer [3]. In this section \( \theta \) refers to the rotor mechanical angle.](image)
SFMO, which can be implemented with a PI functioning as the controller, in this observer the controller is nonlinear since it is a function of the rotor current, more specifically $|\vec{i}_r|^2$ [3]. This has advantages and disadvantages. On one hand, since the controller is a function of a measured quantity, it has more robustness against machine parameter variations. On the other hand, the controller is effectively nonlinear and becomes harder to implement.

### 2.2.4 STATOR CURRENT BASED MRAS OBSERVER (SCMO)

An MRAS observer can also be implemented using the stator current $\vec{i}_s$. In this case, the reference model uses $\vec{i}_s$ measured by the transducers. Solving for $\vec{i}_s$ in (2.11), the adaptive model becomes

$$\dot{\hat{\vec{i}}}_s = \vec{\lambda}_s - L_m \vec{i}_s e^{-j\phi} \frac{L_s}{L_m}.$$  (2.16)

The SCMO cannot be used when $\vec{i}_s$ is low or zero [3]. Therefore, unlike the previous MRAS observers, a load must always be connected to the DFIG stator for correct SCMO operation. This is a disadvantage when the DFIG is sourcing light loads in stand-alone operation, when the DFIG is synchronized with the utility before grid-connected operation, or when speed catching on the fly is required. Therefore, the SCMO is not appropriate for stand-alone operation. Another disadvantage of the SCMO is that the adjustment block has to have a variable gain, just like the RCMO. For the flux-based observers, the lowest gain in the adjustment block is obtained when the direct component of the rotor current is zero.
But for SCMO, there are several points in the plane \((i_{dr}, i_{qr})\) with \(i_{dr} \neq 0\) where the value of the adjustment block will be very low or zero [3]. However, in grid-connected operation when the current \(i_{dr} = 0\), the SCMO small signal gain is similar to that of the RCMO [3]. Therefore, the SCMO can provide tracking as good as that of the RCMO when the rotor current \(i_{dr} = 0\). A block diagram of the SCMO is obtained by replacing the adaptive model of Fig. 2.7 with (2.16) and replacing \(i_r\) with \(\tilde{i}_r\) in the reference model.

### 2.2.5 Parameter Sensitivity & Performance of MRAS Observers

From (2.11), (2.13), (2.15), and (2.16), it can be shown that the phase of \(\hat{\lambda}_r\) is affected by errors in \(L_s\) and \(L_m\), the phase of \(\hat{\lambda}_s\) by errors in \(L_s\) and \(L_m\), the phase of \(\hat{i}_r\) by errors in \(L_r\), and the phase of \(\hat{i}_s\) by errors in \(L_m\). RFMO performance is also affected by errors in \(\sigma L_s\).

Variations in \(R_s\) cause small errors in \(\hat{\lambda}_s\), as seen from (2.10), and in \(\hat{\lambda}_r\) [3]. However, they are negligible since the stator voltage is usually regulated at the nominal value and the voltage drop is small. The MRAS observers proposed in this section are not affected by rotor resistance variations.

Due to the dependence of these MRAS observers on machine parameters, several researchers have proposed machine parameter estimators to be used in conjunction with the MRAS observers. For instance, [7] proposes a self-tuning scheme for the stator inductance, whose value is required in the implementation of the RCMO observer described above. The tuner is based on the stator flux linkage estimation, on the stator...
TABLE 1  PERFORMANCE OF MRAS OBSERVERS [3]

<table>
<thead>
<tr>
<th>Application</th>
<th>SFMO</th>
<th>RFMO</th>
<th>SCMO</th>
<th>RCMO</th>
</tr>
</thead>
<tbody>
<tr>
<td>Grid connected operation with high $i_r$.</td>
<td>Good.</td>
<td>Good.</td>
<td>Good.</td>
<td>OK.</td>
</tr>
<tr>
<td>Grid connected operation with high $i_r \approx 0$.</td>
<td>No.</td>
<td>OK (with high $\sigma$).</td>
<td>Good.</td>
<td>Good.</td>
</tr>
<tr>
<td>Stand-Alone operation.</td>
<td>Good.</td>
<td>OK.</td>
<td>OK.</td>
<td>No.</td>
</tr>
<tr>
<td>Synchronization of the DFIG with the Grid.</td>
<td>Good.</td>
<td>Good.</td>
<td>Good.</td>
<td>No.</td>
</tr>
</tbody>
</table>

currents measurements, and on the squared norm of the rotor current space vector (measured). Hence, the need for these parameter estimators increases the computational effort required for the implementation of the MRAS observers.

The performance of the MRAS observers presented is summarized in Table 1. The performance quality is categorized as “Good,” “OK,” and “No”. “OK” is used to describe MRAS observers which have adequate tracking in: (a) a restricted operating range, (b) when the machine parameters are a certain value (e.g., large value of $\sigma$), or (c) for observers which are not simple to implement. It is assumed that before operating the machine as a grid-connected DFIG, speed catching on the fly and synchronization of the DFIG to the grid has to be realized. Therefore, the performance of the MRAS observers for these tasks is included in Table 1. As stated before, for stand-alone operation it is assumed that the DFIG is magnetized from the rotor. From the small signal analysis and the experimental results [3], it is concluded that most observers can accurately track $\omega$, and $\theta$, when the magnetizing current is provided from the rotor. The exception is the SCMO, which can become unstable in several operating points of the plane $(i_{dr}, i_{qr})$. For sensorless
stand-alone operation, the performance of the SFMO, RFMO, and RCMO are similar. However, as previously discussed, when the DFIG is magnetized from the rotor, the SFMO is simpler to implement because the gain of the adjustment block can be assumed to be constant. Consequently, this observer is considered the best choice for sensorless stand-alone operation of DFIGs. For grid-connected operation, when the rotor current $i_{dr} = 0$, the current based observers have good performance, provided that $i_{qr}$ is not too small or zero. As shown in Table 1, the RFMO can also operate under these conditions, but only when the value of $\sigma$ is relatively high.

As opposed to the open-loop observers presented previously [1, 5], the MRAS observers shown in this section have good steady state and dynamic performance when the machine is operating at synchronous speed.

### 2.3 CONCLUSIONS

This chapter presented a summary of the different solutions that have been proposed for the sensorless control of DFIMs. Instead of presenting an exhaustive compilation of sensorless methods, an effort was made to showcase the most representative solutions within a particular technological trend. It was also shown that these new technological trends appeared as a response to new problems that the available solutions were not able to meet. In this regard, open-loop methods were the first group presented. This group has a high dependency of machine parameter variations. The dependency of parameter

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Survey of Sensorless Control Methods for DFIMs

43
variations is reduced, but not eliminated, by adaptive methods based on MRAS technology. From this group, several versions of observers were presented. However, none of these methods have been tested during FRT conditions.

The final goal of this dissertation is to present a sensorless method that, besides being machine parameter independent, can work successfully during FRT conditions. The proposed sensorless algorithm will rely on a method that has not been used for DFIMs yet. Hence, Chapter 3 will introduce that technique; then the proposed method will be explained and tested in the following chapters.

2.4 REFERENCES


CHAPTER 3

HIGH-FREQUENCY SIGNAL INJECTION

During the last decade, a clever technique based on high-frequency signal injection (HFI) has emerged as a possible solution to the sensorless problem at zero speed for Permanent Magnet Synchronous Machines (PMSMs) and Singly-Fed Induction Machines (SFIMs) [1-8]. This approach incorporates two features deemed necessary for accurate and robust estimation independent of speed and load: 1) trackable magnetic saliency and 2) high-frequency signal injection [1]. The trackable magnetic saliency is inherent within buried permanent magnet synchronous machines as in [2], but is introduced in induction machines in [3] in the form of a spatial modulation in the rotor leakage inductance. This modulation causes a variation in the stator transient reactance which dominates the stator terminal impedance at high frequencies, thus enabling rotor position (and velocity) to be directly tracked. Finally, a balanced three-phase high-frequency signal is generated by the inverter and superimposed upon the fundamental frequency, enabling the tracking of the machine’s speed [1].

This dissertation introduces a new HFI method to solve the sensorless control problem of grid-connected DFIMs. In the proposed method, a high-frequency signal is injected to the rotor windings to determine the rotor position. As opposed to the conventional HFI
method, the rotor saliency information is not needed. The advantages of the proposed method include being immune to parameter variations, very good performance at synchronous speed and during FRT conditions, and ease of implementation [9].

This chapter will serve as an introduction to the generic HFI method that serves as the basis for the sensorless technique proposed in this dissertation. It is important to point out that there are several sensorless algorithms based on HFI. Hence, there is not such a thing as a generic HFI method. Nevertheless, the version shown in this chapter has many of the distinctive characteristics present in many of the versions proposed by researchers. In the remainder of the chapter, the term generic will be used to refer to the HFI algorithm being described.

Finally, it is important to highlight that even though high-frequency injection has been widely used as a sensorless control method for other machines, it has not been fully exploited in DFIMs. Hence, a study of high-frequency injection for DFIMs represents another important contribution of this dissertation.

3.1 HIGH-FREQUENCY SIGNAL INJECTION (HFI) FOR PMSMS

This section focuses on a summary of the algorithm for sensorless control of PMSM presented by Holtz [8]. This technique exploits the anisotropic properties of a PMSM and has several advantages over methods based on the utilization of the back-EMF.
3.1.1 PRINCIPLE OF OPERATION

A two-phase machine stator, without a rotor, is represented in Fig. 3.1. The winding arrangements, alpha and beta, are orthogonal to one another, and therefore there is no magnetic interaction between them. A PM-rotor is added to this basic arrangement, as seen in Fig. 3.2. The PM field is characterized by the magnetization vector $\vec{M}$. It saturates the stator iron locally, thus increasing the magnetic resistance of its flux paths. A magnetic anisotropy is thus created, which is modeled in Fig. 3.2 by a partially enlarged air gap. The flux density distribution $\vec{B}_{\alpha}$ is generated by the current $i_\alpha$. Its intensity is so low that the saturation of the iron is not changed. As long as the direction of the magnetic flux, $\vec{M}$, is aligned with the $\alpha$-axis, as shown in Fig. 3.2, the flux density distribution $\vec{B}_{\alpha}$ conserves its symmetry with respect to the $\alpha$-axis. Also, the flux linking the windings of phase $\beta$ remains zero. This situation changes when the rotor is in a different position, as shown in Fig. 3.3.

Fig. 3.1. Two-phase winding system in stator coordinates [8].

Fig. 3.2. Two-phase winding system, magnet rotor aligned with $\alpha$-axis [8].
Since magnetic flux gets deflected to regions of lower reluctance, the magnetic flux $\vec{B}_{\alpha}$ will now link the orthogonal phase winding $\beta$. As a consequence, a current $i_\beta$ is created in this winding if the current $i_\alpha$ is time varying. The direction of the current $i_\beta$ is determined by Lenz’s law: the induced current will generate a flux density distribution that will oppose the change in $\vec{B}_{\alpha}$ as experienced by the winding $\beta$. The resulting polarity of $i_\beta$ is shown in Fig. 3.3. In order to use the previous ideas to find the rotor position in a PM machine, the system in Fig. 3.2 is subjected to a coordinate transformation. The new reference frame is made to rotate in synchronism with the magnet rotor. The north-pole axis of the magnet rotor defines the real axis of a rotor oriented reference frame, marked R in Fig. 3.4. The excitation of the winding system is chosen as an AC voltage. It is injected in the spatial direction, R’, of an estimated rotor position, thus making use of existing a priori knowledge. The injected voltage, $u_c$, creates a carrier current in the stator windings, $i_c$, and a spatial flux density distribution, $\vec{B}_c$, both being dependent on the anisotropic conditions caused by the magnet’s field. Their spatial orientations deviate from the direction of the injected voltage. While the flux density distribution $\vec{B}_c$ shows a displacement away from the magnetization axis $\vec{M}$, the space vector of the carrier current $i_c$ will move closer to the true rotor (magnetization) axis. The reason is that the magnetic flux becomes attracted by regions of less magnetic resistance. It is deflected in a clockwise direction as seen from the injection axis R’. Therefore, by virtue of Lenz’s law, the carrier current vector is deflected in an anticlockwise direction, i.e., closer to the true rotor axis. Based on this principle, the rotor position can be found.
3.1.2 BASIC MATHEMATICAL ANALYSIS

As shown in Fig. 3.4, an AC carrier voltage, \( u_c \), is applied into the stator windings at an estimated displacement angle, \( \gamma_u \), with respect to the true field axis. From Fig. 3.4, \( \theta \) is the true rotor position angle, while \( \theta' \) is its estimate. Then, \( \theta' = \theta + \gamma_u \). The carrier voltage in stator coordinates is then

\[
\begin{align*}
u'_c &= u_c \cos \omega_c \cdot e^{j \theta} = r_s i'_c + L_s' \frac{di'_c}{dt},
\end{align*}
\]

(3.1)

where \( \omega_c \) is the carrier frequency. The term \( r_s i'_c \) is subsequently neglected since the carrier frequency is much higher than the fundamental frequency. To simplify the analysis, (1) is transformed to rotor coordinates by multiplying by \( e^{-j \beta} \). Then (3.1) becomes

\[
\begin{align*}
u'_c &= u_c \cos \omega_c \cdot e^{j (\theta - \beta)} = L_s' \frac{di'_c}{dt} + j \omega \cdot L_s' \cdot i'_c.
\end{align*}
\]

(3.2)
The last term in (3.2) can be neglected since the mechanical angular frequency \( \omega = \frac{d\bar{\theta}}{dt} \) is considered to be small. After the transformation, \( L' \) becomes the constant matrix

\[
L' = \begin{bmatrix}
L_d & 0 \\
0 & L_q
\end{bmatrix}
\] (3.3)

where \( L_d < L_q \), since the d-axis coincides with direction of the magnetic flux generated by the rotor magnet, \( \bar{M} \).

The solution of (3.2) is

\[
i' = \frac{u}{\omega} \sin \omega_t \cdot \left( \frac{1}{L_d} \cos(\bar{\theta} - \gamma) + j \frac{1}{L_q} \sin(\bar{\theta} - \gamma) \right).
\] (3.4)

From (3.4) it is possible to see that, due to the anisotropies of the machine, the space vector of the current \( i_c \) will be closer to the true magnetic axis than the injected voltage. If \( L_d \) were equal to \( L_q \), the angular displacement of the carrier current, \( \gamma_i \), would be

\[
\gamma_i = \tan^{-1} \left( \frac{\sin(\bar{\theta} - \gamma)}{\cos(\bar{\theta} - \gamma)} \right) = \bar{\theta} - \gamma = \gamma_u,
\] (3.5)

which is precisely the angular displacement of the injected carrier voltage, as seen in Fig. 3.5. But since \( L_d < L_q \), (3.5) becomes

\[
\gamma_i = \tan^{-1} \left( \frac{\sqrt{L_d} \sin(\gamma_u)}{\sqrt{L_q} \cos(\gamma_u)} \right) \rightarrow \gamma_i < \gamma_u.
\] (3.6)

Hence, the displacement of the carrier current space vector is closer to the true axis of the rotor than the displacement of the carrier voltage, which confirms the argument put forward in the previous section.
3.1.3 ALGORITHM FOR ROTOR POSITION ESTIMATION

Figure 3.5 is a space vector diagram showing the trajectories of the oscillating carrier signals $u_c$ and $i_c$ in rotor coordinates, R, and the estimated rotor coordinates, R’. $u_{s1}$ and $i_{s1}$ represent the fundamental stator voltage and current, respectively. The enlarged inset shows that the q-component, $i_{eq}$, of the carrier current vector in the estimated reference frame, R’, is an indicator for the angular displacement, $\Delta \gamma = \gamma_u - \gamma_i$, between the carrier voltage vector and current vector. For $\Delta \gamma = 0$, (3.6) implies that $\gamma_u = \gamma_i = 0$, hence R’→R for $\dot{i}_{eq} \rightarrow 0$. This means that the correct rotor position angle is obtained by minimizing the voltage injection angle $\gamma_u$. This is achieved by controlling the q-component $i_{eq}$ of the estimated carrier current vector to a zero value.

Fig. 3.6 shows the signal flow graph of the machine control scheme and also the elements for sensorless position estimation. The oscillating carrier voltage $u_c$ is applied in
estimated rotor coordinates. The response of the machine is an oscillating carrier current, $i_c$, superimposed to the stator current $i_s$. The measured stator current is transformed to rotor coordinates. The carrier frequency components are separated by a band-pass filter (BPF) from the fundamental current and from the switching harmonics of the inverter. According to Fig. 3.5, the angle $\Delta \gamma$ is assessed through the imaginary component $i_{cq}$ of the carrier current, which, as an AC signal, does not directly reveal the sign of $\Delta \gamma$. The sign is rather hidden in the phase relationship between $i_{cd}$ and $i_{cq}$. The trajectory of $i_{cq}$ in the enlarged inset of Fig. 3.5 shows that, for the error being positive, the instantaneous values of $i_{cd}$ and $i_{cq}$ have the same sign; a negative error exists when $i_{cd}$ and $i_{cq}$ have opposed signs. Hence, the displacement angle becomes

$$\Delta \gamma = i_{cq} \cdot \text{sign}(i_{cd}). \quad (3.7)$$

The error signal $\Delta \gamma$ thus obtained is passed to the PI controller in Fig. 3.6. The controller adjusts its output such that $\Delta \gamma \to 0$, thus keeping the estimated rotor coordinates aligned with the $d$-axis of the machine. The controller output is interpreted as the angular mechanical velocity $\hat{\omega}$, the integral of which is therefore the estimated rotor position angle $\hat{\theta}$. This angle is used both for coordinate transformation and for position control.

### 3.1.4 INITIAL ROTOR ANGLE AND POLARITY DETECTION

The position and polarity of the rotor of a PMSM is needed before it can be started. Electromagnetic torque must not be produced during the identification process. The current
reference signal $i_s^*$ of the control system in Fig. 3.6 is therefore set to zero, while a low amplitude carrier signal, $u_c$, is injected. It creates an alternating carrier current in an arbitrary direction. A nonzero displacement angle, $\Delta \gamma$, between the injected voltage, $u_c$, and the carrier current, $i_c$, is therefore detected. It activates the closed-loop position estimation scheme Fig. 3.6, thus making the estimated rotor position align with one of the rotor axes. The rotor angle thus estimated is $\hat{\vartheta}$. It either indicates the correct direction of the magnet flux, or it is in error by a displacement of $\pi$ rad.

The next step is to determine the polarity of the rotor. For this purpose, two short voltage pulses of the same volts-seconds magnitude are applied to the machine. One is applied in the spatial direction of $\hat{\vartheta}$, and the other in the spatial direction $\hat{\vartheta} + \pi$. Since the resulting current pulses are going to be aligned with the direct axis, they do not produce torque. Effectively, one of the pulses will align with the direction of the magnet flux, thus
increasing the magnetization of the stator iron and driving the direct axis inductance $L_d$ into deeper saturation. The other current pulse will desaturate the stator iron and, as a consequence, increase the value of $L_d$.

Consequently, since both voltage pulses are of the same volts-seconds value, the value of the resulting currents will be influenced by the value of the respective inductances. The current pulse with the higher magnitude will indicate the positive direction of the rotor axis as the inductance in this direction is lower.

The plots of Fig. 3.7 illustrate the initialization process of the position control scheme. The machine remains at standstill throughout this process. The rotor position estimation scheme of Fig. 3.6 first changes the orientation of the oscillating carrier to coincide with the estimated $d$-axis of the machine. A certain value of $\hat{\theta}$ is thus obtained at $t = t_2$. The carrier injection is subsequently discontinued to determine the magnet polarity. The example of Fig. 3.7 shows that the first current pulse exhibits higher amplitude, meaning
that the estimated rotor position value is correct. After the polarity is detected, the injection of the carrier is resumed at $t > t_3$, and the position control scheme is subsequently started.

### 3.2 DIFFERENCES BETWEEN GENERIC AND PROPOSED HFI METHODS

The generic HFI algorithm just presented, and the proposed algorithm in this dissertation (which will be introduced in Chapter 4) are both based on the injection of a probing signal in order to determine the rotor position of an electromechanical machine. But beyond that common point, several differences arise:

- **MACHINE TYPE AND SCOPE:**

The generic HFI algorithm needs some level of saliency. Surfaced-mounted PMSMs do not tend to have structural saliency, but normally have magnetic-based saliency. Given the machine parameters provided in [8], the amount of saliency seems to come from a surface-mounted PMSM, $L_q/L_d = 1.22$. Therefore, interior PMSMs could also benefit from this rotor-position identification method, since these motors normally have even higher saliency ratios. This method will be mainly used at very low speeds since at that point back-EMF methods fail.

The method proposed in this dissertation [9] is mainly tailored for DFIMs close to synchronous speed since back-EMF becomes null at this speed. As such, no saliency is required.
**INJECTED VOLTAGE:**

The generic HFI algorithm injects an alternating high-frequency voltage, whereas the method proposed in this dissertation [9] injects a rotating high-frequency signal. In order words, if a PMSM using the generic HFI algorithm is at standstill, the locus of the space-vector of the carrier voltage will describe a straight line, as in Fig. 3.8, whereas the locus of the space-vector of the rotating voltage always describes a circle.

**DYNAMIC RESPONSE:**

Since the injected voltage in the generic HFI algorithm is a pulsating signal mostly aligned with the d-axis, the resulting high-frequency current will be mostly on the d-axis. It is an advantage of this injection technique that the dynamic performance of the speed and torque control system is not impaired, as the carrier signal does not appear in the torque building current component $i_q$. Therefore, the $q$-current in the current feedback path does
not need to be low-pass filtered. Such filter is required when using a rotating carrier. It reduces the bandwidth for the control of the machine torque. According to Fig. 3.6, a low-pass filter is only provided for the component $i_d$ in the excitation axis.

- **IMMUNITY TO NONLINEAR INVERTER EFFECTS:**

The injected voltage is needed for the algorithm suggested in this dissertation [9]. Nevertheless, due to nonlinear inverter effects, the commanded carrier voltage is different from the actual injected voltage. These nonlinear effects are caused by the signal delay of the pulsewidth modulator, by the storage time delay, and the threshold voltages of the semiconductor switches.

On the other hand, the generic HFI algorithm [8] avoids such an adverse effect by not using the injected voltage as a reference signal when extracting the rotor position information from the carrier current. The carrier current itself is used instead as the reference signal. The advantage of such an approach is that both the position information and the reference signal undergo the delays and nonlinear distortions of the inverter. The resulting effect is that the position information remains undisturbed.

### 3.3 CONCLUSIONS

This chapter introduced the HFI method, which serves as the basis for the sensorless technique proposed in this dissertation. It was shown that at the core of this method a
probing voltage, with a frequency relatively high with respect to the fundamental frequency, is applied to the machine while the currents are monitored. The high-frequency voltage is independent of the voltage used to control the machine. With the exception of measuring the resulting high-frequency currents, the proposed HFI method shares many of the principles described here, as will be shown in the next chapter.

3.4 REFERENCES


CHAPTER 4

A NEW HIGH-FREQUENCY INJECTION METHOD FOR DFIMS

This chapter introduces a new method to solve the sensorless control problem for a grid-connected Doubly-Fed Induction Machine (DFIM). The proposed method is based on high-frequency signal injection and the fact that the rotor of a DFIM can be seen as the rotating secondary of an induction transformer. As explained in the previous chapter, the currently used high-frequency injection techniques apply a voltage to the machine and then measure the resulting current since the rotor position information will be encoded in it. The proposed method, on the other hand, measures the voltage generated in the stator of the machine. This chapter will show that the rotor position information is related to the phase difference between the space vector of the injected voltage and the space vector of the recovered high-frequency voltage at the stator. An intuitive explanation of how the system works will be shown first, followed by a more rigorous mathematical exposition of the proposed method, and then the small-signal modeling and the sensitivity of the proposed method to various working conditions will be presented. The results of several simulation and experimental results have been included in this chapter in order to verify the validity of the proposed method. The chapter ends with the conclusions and references.
4.1 THEORETICAL ANALYSIS OF THE PROPOSED METHOD

This section presents a newly proposed HFI method with a relatively simple algorithm to solve the sensorless control problem of grid-connected DFIMs. As in standard HFI methods, a high-frequency signal is injected to the rotor windings to determine the rotor position, but rotor saliency information is not needed. The advantages of the proposed method include being immune to parameter variations, very good performance at synchronous speed, and ease of implementation.

4.1.1 PRINCIPLE OF OPERATION

The proposed method is based on the principle that a DFIM works as a transformer in which the relative position between the stator and rotor windings changes as the rotor rotates. In other words, the phase difference between the stator and rotor voltages in a

![Fig. 4.1](image-url)  
**Fig. 4.1** Representation of a 3-ph DFIM in which the rotor is rotated by 35 deg. Only phase a windings are shown.
Fig. 4.2 Phase relation between the stator and rotor voltages for the rotor position of Fig. 4.1.

Fig. 4.3 Representation of a 3-ph DFIM in which the rotor is rotated by -35 deg. Only phase \(a\) windings are shown.

Fig. 4.4 Phase relation between the stator and rotor voltages for the rotor position of Fig. 4.3.
DFIM is a function of the rotor positions.

Hence, if a high frequency signal is injected into the rotor winding, the corresponding signal obtained from the stator winding will contain the rotor position information. This idea is exemplified in Figs. 4.1 and 4.2, in which the rotor is displaced by 35 deg. with respect to the stator axis. The diagrams of Fig. 4.1 and 4.3 represent the phase $a$ of the stator and rotor windings of a 3-phase, 2-poles, DFIM. As seen in Fig. 4.2, the phase difference between the phase voltages is a function of the angular displacement of the rotor. If the machine were single-phase, the induced voltage would be zero, but for a 3-phase machine the induced voltage maintains a constant magnitude regardless of the angular displacement of the rotor.

Similarly, Fig. 4.4 shows the phase relation between the stator and rotor voltages of the machine for which the rotor has been rotated by -35 deg., as shown in Fig. 4.3. Here, again, the phase difference between stator and rotor voltages is a direct function of the angular displacement of the rotor.

If the machine were single-phase, the induced voltage magnitude would vary as the rotor moves, but for a 3-phase machine the induced voltage maintains a constant magnitude regardless of the angular displacement of the rotor.

**4.1.2 MATHEMATICAL ANALYSIS**

The winding arrangement for a 2-pole, 3-phase, wye-connected symmetrical induction machine is shown in Fig. 4.5. Three-phase stator windings are identical, sinusoidally...
distributed windings, displaced by $120^\circ$, with $N_s$ equivalent turns and resistance $r_s$. The rotor windings are also three identical, sinusoidally distributed windings, displaced by $120^\circ$, with $N_r$ equivalent turns and resistance $r_r$. As shown in Fig. 4.5, $\theta_r$ represents the angular difference between the axis of the winding related to phase $a$ in the stator and the axis of the winding related to phase $a$ in the rotor. It is assumed that a high frequency voltage source is connected to the rotor circuit as shown in Fig. 4.6.

Referring all the rotor variables to the stator, in abc reference frame voltage equations can be written as

\begin{equation}
\mathbf{\vec{V}}_{\text{abc}} = r_s \cdot \mathbf{\vec{i}}_{\text{abc}} + \frac{d \mathbf{\vec{\lambda}}_{\text{abc}}}{dt},
\tag{4.1}
\end{equation}

\begin{equation}
\mathbf{\vec{V}}_{\text{abcr}} = r_r \cdot \mathbf{\vec{i}}_{\text{abcr}} + \frac{d \mathbf{\vec{\lambda}}_{\text{abcr}}}{dt},
\tag{4.2}
\end{equation}

where

\begin{align*}
(\mathbf{\vec{V}}_{\text{abc}}) &= (V_a \ V_b \ V_c), \\
(\mathbf{\vec{V}}_{\text{abcr}}) &= (V'_a \ V'_b \ V'_c) \\
(\mathbf{\vec{\lambda}}_{\text{abc}}) &= (\lambda_a \ \lambda_b \ \lambda_c), \\
(\mathbf{\vec{\lambda}}_{\text{abcr}}) &= (\lambda'_a \ \lambda'_b \ \lambda'_c) \\
(\mathbf{\vec{i}}_{\text{abc}}) &= (i_a \ i_b \ i_c), \\
(\mathbf{\vec{i}}_{\text{abcr}}) &= (i'_a \ i'_b \ i'_c).
\end{align*}
The inductance matrices are defined as

\[
[\mathbf{L}_s] = \begin{pmatrix}
L_{is} + L_{ms} & -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} \\
-\frac{1}{2}L_{ms} & L_{is} + L_{ms} & -\frac{1}{2}L_{ms} \\
-\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} & L_{is} + L_{ms}
\end{pmatrix},
\quad
[\mathbf{L}_r] = \begin{pmatrix}
L_{ir} + L_{ms} & -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} \\
-\frac{1}{2}L_{ms} & L_{ir} + L_{ms} & -\frac{1}{2}L_{ms} \\
-\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} & L_{ir} + L_{ms}
\end{pmatrix},
\]

\[
[\mathbf{L}_{sr}'] = [\mathbf{L}_{rs}'] = L_{ms} \begin{pmatrix}
\cos \theta_r & \cos(\theta_r + \frac{2\pi}{3}) & \cos(\theta_r - \frac{2\pi}{3}) \\
\cos(\theta_r - \frac{2\pi}{3}) & \cos \theta_r & \cos(\theta_r + \frac{2\pi}{3}) \\
\cos(\theta_r + \frac{2\pi}{3}) & \cos(\theta_r - \frac{2\pi}{3}) & \cos \theta_r
\end{pmatrix},
\]

and the fluxes are

\[
\begin{pmatrix}
\vec{\lambda}_{abc} \\
\vec{\lambda}_{abcr}'
\end{pmatrix} = \begin{pmatrix}
[L_s] & [L_r']
\end{pmatrix} \begin{pmatrix}
\vec{i}_{abc} \\
\vec{i}_{abcr}'
\end{pmatrix}.
\]

(4.3)

Since the frequency of the injected high frequency signal, \(\omega_{hf}\), is several times higher than the operational frequency of the machine, \(r_r \ll X_r\), and Eq. (4.2) becomes

\[
\vec{V}_{abcr}' \approx \frac{d\vec{\lambda}_{abcr}'}{dt}. \quad \text{Hence, if a balanced voltage set,}
\]

\[
\vec{V}_{abcr}' = \begin{pmatrix}
V_{hf} \cos(\omega_{hf} t) & V_{hf} \cos(\omega_{hf} t - \frac{2\pi}{3}) & V_{hf} \cos(\omega_{hf} t + \frac{2\pi}{3})
\end{pmatrix}^T,
\]

(4.4)

is injected at the rotor’s terminals, in steady-state the rotor flux linkages will be

\[
\vec{\lambda}_{abcr}' = \frac{1}{\omega_{hf}} \begin{pmatrix}
V_{hf} \sin(\omega_{hf} t) & V_{hf} \sin(\omega_{hf} t - \frac{2\pi}{3}) & V_{hf} \sin(\omega_{hf} t + \frac{2\pi}{3})
\end{pmatrix}^T,
\]

(4.5)

From (4.3), the rotor current becomes

\[
\vec{i}_{abcr}' = [L_r']^\dagger \vec{\lambda}_{abcr}' - [L_r']^\dagger [L_s']^T \vec{i}_{abcr}'.
\]

(4.6)
Hence, replacing $\vec{i}_{abc}$ by (4.6) when solving for $\vec{\lambda}_{abc}$ in (4.3),

$$\vec{\lambda}_{abc} = [L_s][\vec{i}_{abc} - [L_{sr}][L_r]^{-1}[L_{sr}]\vec{\omega}_r + [L_{sr}][L_r]^{-1}\vec{\omega}_r].$$

Therefore, the stator voltage becomes

$$\vec{V}_{abc} \approx \frac{d\vec{\lambda}_{abc}}{dt} = [L_s]\frac{d\vec{i}_{abc}}{dt} - [L_{sr}][L_r]^{-1}[L_{sr}]\vec{\omega}_r \frac{d\vec{i}_{abc}}{dt} + [L_{sr}][L_r]^{-1}\frac{d\vec{\lambda}_{abc}}{dt}$$

$$- \frac{d[L_{sr}][L_r]^{-1}[L_{sr}]\vec{i}_{abc} - [L_{sr}][L_r]^{-1}\frac{d[L_{sr}][L_r]^{-1}d\vec{\lambda}_{abc}}{dt}}{dt} + \frac{d[L_{sr}][L_r]^{-1}d\vec{\lambda}_{abc}}{dt}.$$ (4.7)

The first three terms of (4.7) represent the so-called transformer voltage, whereas the last three terms are responsible for the speed voltage, a voltage proportional to the speed of variation of the mutual inductance.

If the stator current is zero, (4.7) becomes

$$\vec{V}_{abc} \approx \frac{d\vec{\lambda}_{abc}}{dt} = [L_{sr}][L_r]^{-1}\frac{d\vec{\lambda}_{abc}}{dt} + \frac{d[L_{sr}][L_r]^{-1}d\vec{\lambda}_{abc}}{dt}.$$ (4.8)

In (4.8), the first term represents the transformer voltage, the second term the speed voltage, and $\omega_r = d\theta/dt$ is the rotor mechanical speed. For a constant speed, $\theta = \omega_r t$, the argument of the cosine of both terms in (4.8) is the same, but the second term is written in terms of $\omega_r$ to emphasize its dependence on the rotor speed. Finally, since $\omega_{sf} \gg \omega_r$, the second term could be neglected since it is exactly the first term times the ratio of $\omega_r$ over $\omega_{sf}$.
The circuit shown in Fig. 4.6 is assumed if the constraint imposed to obtain (4.8), that is, the stator current is not identically zero, is relaxed. The grid is represented by an ideal voltage source in series with an inductance representing the line inductance, \( V_g \) and \( L_g \), respectively. As mentioned earlier, a high frequency voltage source, \( V_{hf} \), is connected to the rotor, and assuming that the machine is not in saturation, the principle of superposition can be used. Therefore, the grid voltage is assumed to be zero for the analysis at the injected high frequency, i.e., \( V_g = 0 \).

Hence, the stator voltage and current are related by

\[
\vec{V}_{abc} = -L_g \frac{d\vec{i}_{abc}}{dt},
\]

where the minus sign is needed since the current was initially defined for motor operation.

After some algebraic manipulation and using the trigonometric identity

\[
\cos u \cdot \sin v = \frac{1}{2} \left[ \sin(u + v) - \sin(u - v) \right],
\]

the stator voltage will be

\[
\vec{V}_{abc} = \frac{3L_{ms}V_{hf}L_g}{\omega_{hf} \left( 2L_{ls}L_{th} + 3L_{ls}L_{ms} + 3L_{th}L_{ms} + 3L_{ms}L_{ms} + 2L_{ls}L_g \right)} \left\{ \begin{array}{c}
\cos(\omega_{hf}t + \theta_x) \\
\cos(\omega_{hf}t - \frac{2\pi}{3} + \theta_x) \\
\cos(\omega_{hf}t + \frac{2\pi}{3} + \theta_x)
\end{array} \right\}.
\]

**Fig. 4.6** Per phase equivalent circuit of a grid-connected DFIG with high-frequency injected in the rotor. \( V_g = 0 \) for the analysis at high-frequency.
Due to the physical characteristics of DFIMs, reasonable assumptions can be made in order to simplify (10). For instance, assuming that \( L_{ls} = L_{r} \), Eq. (4.10) becomes

\[
\vec{V}_{abcs} = \frac{3L_{ms}V_{hf}L_{g} (\omega_{r} + \omega_{hf})}{\omega_{hf} \left(2L_{ls}^2 + 6L_{ls}L_{ms} + 3L_{g}L_{ms} + 2L_{ls}L_{g}\right)} \begin{bmatrix}
\cos(\omega_{hf} t + \theta_{r}) \\
\cos(\omega_{hf} t - \frac{2\pi}{3} + \theta_{r}) \\
\cos(\omega_{hf} t + \frac{2\pi}{3} + \theta_{r}) \\
\end{bmatrix}.
\] (4.11)

Finally, if the value of the line inductance is approximately equal to the machine leakage stator inductance, \( L_{g} \approx L_{ls} \), Eq. (4.11) can be further simplified to

\[
\vec{V}_{abcs} = \frac{3L_{ms}V_{hf} (\omega_{r} + \omega_{hf})}{\omega_{hf} \left(4L_{ls} + 9L_{ms}\right)} \begin{bmatrix}
\cos(\omega_{hf} t + \theta_{r}) \\
\cos(\omega_{hf} t - \frac{2\pi}{3} + \theta_{r}) \\
\cos(\omega_{hf} t + \frac{2\pi}{3} + \theta_{r}) \\
\end{bmatrix} \approx \frac{V_{hf}}{3} \begin{bmatrix}
\cos(\omega_{hf} t + \theta_{r}) \\
\cos(\omega_{hf} t - \frac{2\pi}{3} + \theta_{r}) \\
\cos(\omega_{hf} t + \frac{2\pi}{3} + \theta_{r}) \\
\end{bmatrix},
\] (4.12)

where \( \omega_{hf} \gg \omega_{r} \) and \( 9L_{ms} >> 4L_{ls} \) are assumed.

The result on (4.12) agrees with the assessment that at high frequency the stator currents are governed predominately by the stator and rotor leakage inductances [1] since, after assuming similar values for leakage and line inductances, the stator voltage is one-third of the injected voltage. Comparing (4.4) and (4.12), it becomes clear that the rotor position, i.e. \( \theta_{r} \), is encoded in the high frequency voltage at the stator.
4.2 EXTRACTION OF THE ROTOR POSITION INFORMATION

4.2.1 ALGORITHM IMPLEMENTATION

The rotor angular information is obtained by implementing the simple algorithm shown in Fig. 4.7. The measured stator voltage is filtered by a band-pass filter centered at the high frequency injected in the rotor, i.e., $\omega_{hf}$. The low-pass filter is an analog filter needed to filter out frequencies above the Nyquist frequency of the DSP system. Later, a phase compensation block is used to advance (delay) the signal by a value equal to the phase-delay (phase-advance) introduced by the band-pass and low-pass filters. The phase compensation block is an all-pass filter with a flat magnitude response in the frequencies of interest and a phase response that cancels out the phase-delay (phase-advance) introduced by the band-pass and low-pass filters. Then, the angular position of the space vector of the filtered stator voltage, $\phi_s$, is obtained due to the use of a phase-locked loop (PLL), as shown in Fig. 4.7. According to (4.10), when measured at the stator, the frequency of the injected signal will be changed by the rotational speed of the rotor. This is so because $\theta_r = \omega_r t$; therefore, it follows from (4.10) that at the stator, the frequency of the injected signal is $\omega_{hf} + \omega_r$. Eq. (4.10) shows that the high frequency component of the stator voltage is going to lead the injected high frequency signal (whose space vector has an angular position of $\phi_i$) by the rotor angular position, $\theta_r$, i.e.,

$$\theta_r = \phi_s - \phi_i. \quad (4.11)$$
Hence, $\phi_r$ has to be subtracted from the output of the PLL. Finally, if $\theta_r < 0$ in (4.11), a wrapping block is implemented to keep $\theta_r$ within bounds, i.e., $0 < \theta_r < 2\pi$.

As a final point, as seen in Fig. 4.7, no machine parameters are needed in order to implement the proposed algorithm. The proposed algorithm is analogous to phase modulation in the sense that the magnitude of the measured high-frequency signal on the stator does not contain any valuable information. The only requirement is that the magnitude has to be high enough to be properly detected. According to (4.10), the machine parameters only influence the magnitude of the high-frequency voltage on the stator, not the phase. Therefore, it follows that the proposed algorithm is parameter independent.

### 4.2.2 SELECTION OF THE FREQUENCY OF THE INJECTED SIGNAL

Since the PWM converter generates the high-frequency signal injected on the rotor, its frequency has to be at least an order of magnitude below the switching frequency. If this is not the case, the PWM converter will introduce unwanted harmonics. On the other hand, the high-frequency signal on the stator will have to be properly filtered from the line.
voltage. Therefore, it is recommended to place the frequency of the injected signal at a reasonable spectral distance from the grid frequency.

Finally, since the grid voltage normally exhibits a significant harmonic content, the high-frequency signal on the stator has to be located such that no interference arises at the frequencies at which these harmonics occur. Since the position of the 8th harmonic (480Hz in the U.S.) is vacant in a 3-phase system, this frequency is chosen to be injected on the rotor. As a result, the measured high-frequency signal on the stator will be 540Hz (the 9th harmonic) at synchronous speed and will be limited between 522Hz and 558Hz (assuming that the speed of the DFIM is limited between 0.7 p.u. and 1.3 p.u.). This location guarantees the least filtering effort, since the nearest grid harmonics are at 420Hz and 660Hz (the 7th and 11th harmonics, respectively). Fig. 4.8 shows a graphical representation of this discussion.

4.3 SMALL-SIGNAL ANALYSIS

In order to find the linearization of the proposed algorithm, Fig. 4.9 shows the block diagram of the whole system, including the DFIM. In the diagram of Fig. 4.9, a $3\phi$ high-frequency signal is generated and injected into the rotor of the DFIM, thanks to the power converter shown. It is important to notice that the high-frequency signal used for position estimation is going to be superimposed to the current command signal resulting
from the implemented control algorithm (not covered in this dissertation). Later, the stator voltage is measured and these results are inputted to the block diagram of Fig. 4.7.

In order to proceed with the linearization of the proposed system, several key ideas are highlighted. First, as seen in Eq. (4.11), there is a memory-less (algebraic) relationship between the phase of the injected high-frequency signal at the rotor and the phase of the measured high-frequency signal at the stator. This relationship is analogous to the relationship between the stator frequency, the slip frequency, and the rotor mechanical speed, i.e., \( \omega_{\text{slip}} = \omega_s - \omega_m \).

It is also worth noting that, according to Fig. 4.7, the phase delay introduced by the band-pass and low-pass filters is going to be cancelled by the phase compensation block. This last statement deserves some clarification. It is possible to design FIR filters (equalizers) with a relatively arbitrary phase response in a relatively small bandwidth. Since in this application the band-pass filter has a bandwidth of 36Hz (558Hz – 522Hz), designing the FIR filter is entirely possible. Nevertheless, since magnitude variations can

![Figure 4.8](image.png)

**Fig. 4.8** Frequency spectrum of proposed sensorless control algorithm.
be modeled as a unit step function, and the frequency signature of a unit step function has an imprint in all the frequency range, during transients the FIR equalizer will not be able to perfectly cancel the phase delay introduced by the low- and band-pass filters. Additionally, the time delay introduced by the FIR (phase compensation block) of Fig. 4.7 will be considered negligible due to the fast operating frequency of the DSP, relative to the frequency range of the high-frequency signal injected and recovered.

Finally, the wrapping block of Fig. 4.7 does not have any dynamics; it simply represents the angular information in an unbounded way. As a result, the dynamics of the proposed algorithm will be the dynamics of the PLL shown in Fig. 4.7.
4.3.1 SMALL-SIGNAL MODELING AND ANALYSIS OF 3-Ø PLL

Three-phase PLL systems are able to accurately estimate the phase of the space-vector of a three-phase signal. Due to their great usefulness for grid-connection applications, PLL systems have been studied by different authors [2-3]. The following analysis is based on [2].

The block diagram of a 3-Ø PLL system is shown in Fig. 4.10. In this PLL, the phase angle is detected by synchronizing the PLL rotating reference frame with the high-frequency voltage vector. When the quadrature axis reference voltage \( V_q \) is set to zero, the PLL output locks-in to the phase angle of the high-frequency voltage vector. In addition, the instantaneous frequency and amplitude of the voltage vector are also determined. The feed forward frequency command (wff) can be introduced to improve the overall tracking performance of the PLL, but will be disregarded in the following analysis.

The input voltage to the PLL can be represented as

\[
\vec{V}_{abc} = V_{hfs} \begin{pmatrix} \cos(\theta_a) & \cos(\theta_a - \frac{2\pi}{3}) & \cos(\theta_a + \frac{2\pi}{3}) \end{pmatrix}^T,
\]

where \( \vec{V}_{abc} = [V_a \ V_b \ V_c]^T \), \( V_{hfs} \) is the magnitude of the voltage at the output of the phase

![Fig. 4.11 Linearized model of three-phase PLL [2].](image-url)
compensation block, and \( \theta \) is the phase of this voltage. By using the transformation matrix

\[
T_s(\varphi) = \frac{2}{3} \begin{pmatrix}
\cos(\varphi) & \cos(\varphi - \frac{2\pi}{3}) & \cos(\varphi + \frac{2\pi}{3}) \\
-\sin(\varphi) & -\sin(\varphi - \frac{2\pi}{3}) & -\sin(\varphi + \frac{2\pi}{3}) \\
\frac{1}{2} & \frac{1}{2} & \frac{1}{2}
\end{pmatrix}
\]

the voltages \( V_d \) and \( V_q \) are obtained. Then

\[
\hat{V}_{dq} = \begin{bmatrix} V_d & V_q \end{bmatrix}^T = V_{hfs} \begin{bmatrix} \cos(\theta_s - \phi_s) & \sin(\theta_s - \phi_s) \end{bmatrix}^T,
\]

where \( \theta_s \) is the phase of the space vector of the voltage at the input of the PLL, and \( \phi_s \) is the estimate of \( \theta_s \).

Let, \( \delta = \theta_s - \phi_s \), then, for \( \delta \approx 0 \), the voltage \( V_q \) becomes

\[
V_q = V_{hfs} \sin \delta \\
\approx V_{hfs} \cdot \delta \\
= V_{hfs} \cdot (\theta_s - \phi_s).
\]

It follows that the input to the PI controller of Fig. 4.10 is basically the error between the phase of the space vector of the high-frequency injected voltage and its estimation. The output of the PI controller is the frequency of the high-frequency injected voltage. Finally, an integration block is utilized to get the phase. Hence, the PLL can track the frequency and phase of the three-phase voltage at its input.

The linearized model of the three-phase PLL can be seen in Fig. 4.11. The transfer function of the closed loop system can be represented as

\[
H_{PLL}(s) = \frac{\Phi(s)}{\Theta(s)} = \frac{V_{hfs} PI(s)}{s + V_{hfs} PI(s)},
\]
where \( \text{PI}(s) \), \( \Phi(s) \), and \( \Theta(s) \) denote the Laplace transforms of the PI controller, \( \phi \), and \( \theta \), respectively.

Since the Laplace transform of a PI controller is, as seen in Fig. 4.10,

\[
\text{PI}(s) = \frac{k_i + k_p s}{s}
\]

then the transfer function of the closed loop system can be written out as a function of the Laplace operator as

\[
H_{PLL}(s) = \frac{\Phi(s)}{\Theta(s)} = V_{hi} \frac{k_p s + k_i}{s^2 + V_{hi} k_p s + V_{hi} k_i}.
\]  

Eq. (4.13) shows that the proposed estimation method behaves as a second order system when a conventional three-phase PLL is utilized. This similarity facilitates the design of the PI gains, as will be shown in the next sub-section.

**4.3.2 CONTROLLER DESIGN**

The transfer function of the complete system can be rewritten in the general form of the second order loop as

\[
H_{PLL}(s) = \frac{\Phi(s)}{\Theta(s)} = \frac{2\zeta \omega_n s + \omega_n^2}{s^2 + 2\zeta \omega_n s + \omega_n^2}.
\]  

(4.14)

Then, by comparing (4.13) and (4.14), the gains of the PI controller can be chosen for accordingly as

\[
\omega_p = \sqrt{V_{hi} k_i} \quad \text{and} \quad \zeta = \frac{V_{hi} k_p}{2 \omega_n} = \sqrt{\frac{V_{hi} k_i^2}{2k_i}}.
\]
4.4 SENSITIVITY ANALYSIS

In practice, the three-phase high-frequency signal recovered at the stator is not purely sinusoidal, but distorted every time the fundamental 60 Hz voltage experiences a step change in magnitude, etc. There are also other sources of disturbance, like nonlinearities in the measuring devices and signal conversion errors. All these sources of error result in phase unbalancing, harmonics, and errors in the three-phase high-frequency signal which, in turn, produce errors in the output of the PLL. The following analysis is based on [2].

4.4.1 PHASE UNBALANCING

An unbalanced three-phase high-frequency signal recovered at the stator can be expressed as follows,

$$\vec{V}_{abc} = \left( V_{hfs} \cos \theta_s \right) \left( 1 + \beta \right) \cos \theta_s \left( \frac{2\pi}{3} \right) \left( 1 + \gamma \right) V_{hfs} \cos \left( \theta_s + \frac{2\pi}{3} \right),$$

where $\beta$ and $\gamma$ are constants and $\theta_s$ is the phase of the high-frequency vector at the input of the PLL. These voltages can be transformed to $V_{dq}$ using the Park transformation shown in (4.12). Then, $V_d$ becomes

$$V_d = V_{hfs} \sin \delta$$

$$-V_{hfs} \left[ \frac{\beta - \gamma}{2\sqrt{3}} \left( \cos \theta_s \cos \phi_s - \sin \theta_s \sin \phi_s \right) + \frac{\beta + \gamma}{6} \left( \sin \theta_s \cos \phi_s + \cos \theta_s \sin \phi_s \right) \right],$$

(4.15)

where, as defined before, $\phi_s$ represents the phase estimation result of the PLL.
Under the assumption that the error $\delta = \theta_s - \phi_s$ is very small, and $\theta_s + \phi_s \approx 2\theta_s$, (4.15) can be rewritten as

$$V_q \approx V_{hfs}\delta - V_{hfs}\left[\frac{\beta - \gamma}{2\sqrt{3}} \cos 2\theta_s + \frac{\beta + \gamma}{6} \sin 2\theta_s\right]. \quad (4.16)$$

Since the PLL will control the q-axis to zero, the resultant error due to the phase imbalance will be

$$\delta \approx \sqrt{\left(\frac{\beta - \gamma}{2\sqrt{3}}\right)^2 + \left(\frac{\beta + \gamma}{6}\right)^2} \cos \left(2\theta_s - \tan^{-1}\left(\frac{1}{\sqrt{3}}\frac{\beta + \gamma}{\beta - \gamma}\right)\right).$$

Then, the error introduced by phase unbalancing produces an error component with a frequency double the frequency of the one detected on the stator.

### 4.4.2 VOLTAGE OFFSET

The voltage offset is often produced by errors in the measuring circuit or actual voltage offsets in the signal to be measured. The high-frequency stator signal with offset can be expressed as

$$\tilde{V}_{abc} = \begin{pmatrix} V_{hfs}\cos(\theta_a) + V_{ao} \\ V_{hfs}\cos(\theta_a - \frac{2\pi}{3}) + V_{bo} \\ V_{hfs}\cos(\theta_a + \frac{2\pi}{3}) + V_{co} \end{pmatrix}^T,$$

where $V_{ao}$, $V_{bo}$, and $V_{co}$ represent the offset present in each of the phases.

After applying the transformation matrix of (4.12), $V_q$ becomes

$$V_q = V_{hfs}\sin \delta + \left(\frac{2}{3} (V_{ao} + V_{bo} + V_{co})\right)\sin \phi + \left(\frac{1}{3\sqrt{3}} (V_{co} - V_{bo})\right)\cos \phi.$$
Therefore, since the PI controller in the PLL will make $V_q$ equal to zero, and with the assumption that $\delta \approx 0$, the error caused by the voltage offset can be shown as

$$\delta \approx -\sqrt{\left(\frac{2}{3}(V_{ao} + V_{bo} + V_{co})\right)^2 + \left(\frac{1}{\sqrt{3}}(V_{co} - V_{bo})\right)^2 \cos \theta_s - \tan^{-1}\left(\frac{2\sqrt{3}}{3} \left(\frac{V_{ao} + V_{bo} + V_{co}}{V_{co} - V_{bo}}\right)\right)}.$$ 

Hence, the error introduced by the voltage offset has the same frequency of that of the high-frequency signal recovered on the stator.

### 4.5 SIMULATION AND EXPERIMENTAL RESULTS

#### 4.5.1 SIMULATION RESULTS

A Matlab/Simulink simulation model was developed to test the proposed method. The simulation describes a grid-connected, 4-poles, Doubly-Fed Induction Generator controlled by a back-to-back PWM converter. The configuration used is shown in Fig. 4.12. Without any loss of generality, it is assumed that the line reactance has the same order of magnitude as the stator leakage reactance, i.e., $L_g \approx L_s$. The machine parameters used in the simulations are shown in Table 2.

<table>
<thead>
<tr>
<th>TABLE 2</th>
<th>PARAMETERS USED FOR SIMULATIONS AND EXPERIMENTS</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>DFIM Parameters</strong></td>
<td><strong>in (Ω)</strong></td>
</tr>
<tr>
<td><strong>Power</strong></td>
<td><strong>Poles</strong></td>
</tr>
<tr>
<td>5 HP</td>
<td>4</td>
</tr>
</tbody>
</table>

a. Leakage inductance for stator and rotor have same numerical value.
The frequency and voltage of the injected high-frequency signal are 480Hz and 10% of the magnitude of the rotor voltage, respectively. A second assumption is that the machine rotational speed is close to the machine synchronous speed, i.e., \( \omega_r \approx \omega_{\text{sync}} \).

The main parameters for the simulated Matlab/Simulink model are the system frequency, 60 Hz; the frequency and voltage of the injected high-frequency signal, 480 Hz; and 10% of the magnitude of the rotor voltage, respectively. Fig. 4.13 shows the simulation results of the proposed method when the machine is running at a speed very close to synchronous speed. As seen in the simulation results of Fig. 4.13, the estimated rotor position information basically overlaps the actual rotor position information.

**Fig. 4.12** Configuration used to simulate and test experimentally the proposed sensorless algorithm.

**Fig. 4.13** Simulation results for situation in which the rotor speed is approximately equal to synchronous speed. The estimated result overlaps the measured values.
4.5.2 EXPERIMENTAL RESULTS

Experimental results have been pursued to verify the feasibility of the proposed method at synchronous speed. The system setup for experimental testing was shown in Fig. 4.14, and its block diagram is shown in Fig. 4.12. Two PWM back-to-back inverters are connected to the rotor of the machine. The rotor-side PWM inverter is controlled using a frequency of 10 kHz, which is also the sampling frequency of the sensors and of the interrupt subroutine.

Current transducers are used to measure the rotor currents. Two voltage transducers are used to measure the stator voltage. A 2nd order analog low-pass filter with a cut-off frequency of 863Hz filters out the switching noise but not the high frequency injected for sensing purposes. The band-pass filter of Fig. 4.7 is implemented as a digital filter. A speed encoder of 1024 pulses per revolution is used to measure the rotational speed and rotor angle. The mechanical encoder is used only for comparison purposes. Since a new control technique during FRT conditions is outside the focus of this paper, the conventional stator
field oriented control technique is used to control the DFIM [4].

However, the current loop bandwidth has to be lower than the frequency of the injected signal; otherwise the current controllers will disturb the injected signal. A low-pass filter for the measured currents could be used that would serve the dual purpose of filtering out the injected high frequency signal, as well as the PWM noise. Low-pass filters with a cut-off frequency of 100Hz have been implemented in this experimental set-up. The DFIM is speed controlled as a motor (no prime mover is used), with the stator flux being obtained from the stator voltages and currents.

The machine parameters used for the experimental testing are shown in Table 2. A high-frequency signal of 480Hz is injected to the rotor-side windings while the stator-side voltages are measured. An estimate of the rotor angular position, which is used to control the machine, is obtained by applying the algorithm described in section 4.2 and Fig. 4.7. A microcontroller board, based on the TMS320F2812 from TI, is used to implement the proposed observer and the whole sensorless vector-control system. Since the

![Fig. 4.15 Experimental results obtained with Code Composer Studio™. The top plot shows the measured rotor position (p.u.) for the machine under test when it is running at synchronous speed. The bottom plot shows the estimated rotor position using the proposed algorithm.](image)
Fig. 4.16 Rotor phase currents when the machine under test goes from 0.8 p.u. to synchronous speed. It is shown that the machine remains at synchronous speed for 18 secs. Time/div equals 4 secs.

Therefore, unlike previous publications based on rotor flux estimation [6], the proposed observer has good performance at synchronous speed. It is important to point out that the microcontroller utilized does not have DACs, the results of the proposed algorithm (like rotor angular position and speed) are saved in the microcontroller memory and then exported to Matlab for display, or taken directly from the screen of Code Composer Studio™.

Fig. 4.15 shows the result of the proposed algorithm, obtained with Code Composer Studio™, when the machine is running at synchronous speed. Fig. 4.16 shows the rotor currents for phase \( a \) and phase \( b \) when the machine under test runs at 0.8 p.u. and synchronous speed. The machine is initially running at 0.8 p.u., and after 6 seconds the speed command changes to 1.0 p.u. At that point, the controller commands an increase in current magnitude in order to boost the machine torque. After the transient phase, the rotor phase currents become DC values and remain as such for 18 seconds. Note that the time/div is 4 seconds.
number of poles of the machine is the only machine parameter needed in order to implement the proposed algorithm.

The performance of the proposed algorithm has been tested for changes in the speed command. Fig. 4.17 shows the measured, command, and estimated speed, as well as the estimation error. Initially, the machine is running at 1.2 p.u. After 1.8s the speed command starts to change until it reaches 0.8 p.u. at \( t = 5.5 \)s. This speed command is kept for 4s, until it is changed again to 1.2 p.u. These values were chosen since a DFIM used for wind power generation normally works in this speed range. As seen in Fig. 11-a, the proposed algorithm follows the real speed very closely, with an error, shown in Fig. 11-b, of less than 2\%.

![Fig. 4.17 Experimental results showing the dynamic response of the proposed algorithm to changes in the speed command. The results show (a) the measured, command, and estimated rotor speed; and (b) the error between the measured and estimated speed.](image)

New High-Frequency Injection Method
4.6 CONCLUSIONS

In this chapter, a new method of rotor position identification for sensorless control of grid-connected DFIGs was proposed and discussed. The mathematical foundations of the proposed method, its small signal modeling, and an algorithm for its implementation were presented. Computer simulation and experimental results were shown in order to confirm the high accuracy of the proposed technique. Special testing emphasis was placed on the synchronous speed condition, since other sensorless methods have failed at this particular operating condition. The analysis and experimental results show that knowledge of the machine parameters, with the exception of the number of poles, is unnecessary in order to apply the proposed sensorless method.

The results presented dealt with the implementation, steady-state and small signal analysis, experimental dynamic performance, and feasibility during synchronous speed of the proposed sensorless observer. The next chapter will show that the proposed sensorless technique is also able to operate successfully under grid fault-ride through conditions.

4.7 REFERENCES


New High-Frequency Injection Method


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New High-Frequency Injection Method

85
CHAPTER 5

COMPARISON BETWEEN CURRENTLY USED SENSORLESS METHOD AND THE NEWLY PROPOSED ALGORITHM

This chapter compares the newly proposed algorithm against one of the most commonly used methods for sensorless control of DFIMs. The stator-flux MRAS Observer (SFMO) is chosen for comparison given its wide spread use, ease of implementation, and the fact that it has been thoroughly investigated [1-3]. Comparison studies are performed in two major areas that are relevant to the sensorless control of DFIMs: a) performance during fault-ride through (FRT) conditions and b) sensitivity against parameter variations. These studies are validated with the use of computer simulations and, when permitted, with experimental results.

The commonly used sensorless techniques based on stator-flux linkage, besides being parameter sensitive, do not work during fault ride through (FRT) conditions. As a way to illustrate the necessity for a new sensorless control algorithm for DFIMs, this chapter will explain the fault ride-through (FRT) problem for grid-connected DFIMs. Later, the performance of SFRO will be evaluated during FRT conditions. Then, the newly proposed algorithm will be evaluated during the same strenuous conditions. Experimental results will be included to confirm the validity of the simulations shown.
The chapter concludes with highlights of the most important differences between the proposed sensorless control method and SFMO.

5.1 PERFORMANCE DURING FAULT-RIDE THROUGH (FRT)

5.1.1 THE LOW VOLTAGE RIDE-THROUGH (LVRT) PROBLEM IN DFIMs

The problem of LVRT has been widely studied in the literature [4-6]; the following development follows the presentation in [4]. Fig. 5.1 shows the block diagram of a DFIG wind turbine system. The generator has a three-phase wound rotor supplied via slip rings, from a four-quadrant pulse width modulation (PWM) converter, with voltage of controllable amplitude and frequency. The generator can operate at variable speed, while the stator remains at a constant grid frequency. The machine usually operates in the vector-control mode oriented to the stator-flux linkage, which can be calculated as shown in Eq. (5.1) by using the measured stator voltages and currents,

\[
\tilde{\lambda}_{abc} = \int (V_{abc} - R_s \cdot i_{abc}) dt. \tag{5.1}
\]

The stator- and rotor-flux linkages are related to the currents as

\[
\tilde{\lambda}_s^s = (L_{ls} + L_m)\tilde{i}_s^s + L_m\tilde{i}_r^s \tag{5.2}
\]

\[
\tilde{\lambda}_r^s = L_m\tilde{i}_s^s + (L_{lr} + L_m)\tilde{i}_r^s \tag{5.3}
\]

where \(L_{ls}\) and \(L_{lr}\) are the leakage inductances, and \(L_m\) is the magnetizing inductance. During normal operation, both the stator- and rotor-flux linkages rotate at synchronous speed with...
respect to the stator. In the rotor reference frame, the voltage equation of the rotor windings is

\[ \ddot{u}_r = R_r \dot{i}_r + \frac{d}{dt} \ddot{\lambda}_r. \tag{5.4} \]

The rotor-flux linkage can be expressed in terms of stator-flux linkage and rotor current according to Eqs. (5.2) and (5.3) as

\[ \ddot{\lambda}_r = \frac{L_m}{L_s} \ddot{\lambda}_s + \frac{L_r L_s - L_m^2}{L_s} \ddot{i}_r. \tag{5.5} \]

The equivalent circuit of the DFIG machine viewed from the rotor side, corresponding to

![Fig. 5.1 Configuration of the DFIG wind turbine system [4].](image1)

![Fig. 5.2 Rotor-side equivalent circuit. [4].](image2)
Comparison between Proposed Method and the State of the Art in Sensorless Control

(5.4) and (5.5), is shown in Fig. 5.2. The rotor current is jointly decided by the injected rotor voltage and the induced electromotive force (EMF), which is the derivative of the stator-flux linkage with respect to time. In normal operation, the space vector $\tilde{\lambda}_r$ rotates at slip speed $s\omega_e$, with respect to the rotor winding, where $s$ is the slip and $\omega_e$ is the synchronous speed. Therefore, $d\tilde{\lambda}_r / dt = j s\omega_e \tilde{\lambda}_r$. This is predominantly balanced by the rotor voltage of the same frequency and determines the voltage rating range of the rotor-side converter. Now, consider a three-phase fault on the grid side, which brings the stator terminal voltage down to zero. In such an extreme case, the space vector of the stator-flux linkage, $\tilde{\lambda}_s$, will “freeze” and stop rotating with respect to the stator winding, but its magnitude will remain relatively unchanged, as implied by Eq. (5.1). As a result, $\tilde{\lambda}_s$ rotates at speed $\omega_r$, rotor speed, with respect to the rotor winding. The induced EMF will then become $j\omega_s \tilde{\lambda}_s$. This voltage value tends to be higher than what the rotor-side converter can generate, which causes a voltage mismatch resulting in large rotor currents. These currents could potentially damage the rotor-side power converter.

As a way to illustrate the severity of the problem, Fig. 5.3 shows the simulated fault current for a 2-MW DFIG system whose parameters have been described in [4]. The machine initially operates with a full load and is at 30% super synchronous speed. A three-phase grid fault brings the generator terminal voltage down to 0.3 p.u. The control implemented in the simulation model is the vector-control algorithm for terminal voltage regulation and the maximum active power tracking. The rotor-side converter is represented
as controlled voltage sources. No additional action or control limit is included to constrain the fault current. It is observed that the rotor current increases from about 1.0 to 7.2 p.u.

The issues concerning DFIMs during FRT could be divided into two sub-problems. The first sub-problem would be the actual control of the machine. One of the objectives in this sub-problem is to be able to control the rotor currents during the fault giving a limited DC bus voltage. Another related objective could be to protect the rotor-side converter from the high-currents that tend to appear during a grid-fault. Several papers have been written on these issues [4, 6]. A second sub-problem, which is separate from the first, is the possibility

![Fig. 5.3 Simulation results of a 2-MW DFIG during a grid fault. [4].](image)

![Fig. 5.4 Block diagram of a typical MRAS observer [1].](image)
Comparison between Proposed Method and the State of the Art in Sensorless Control

of properly estimating the rotor position during grid-faults, which is needed for sensorless control. This dissertation presents a technique that is able to properly estimate the rotor position during several conditions, grid-faults included. In this sense, this dissertation is not presenting a new control technique, but a new estimator.

The following sub-section will show the poor performance in angle estimation of a commonly used sensorless algorithm during FRT conditions.

5.1.2 PERFORMANCE OF COMMONLY USED MRAS ALGORITHM DURING LVRT

Fig. 5.4 shows the block diagram of a typical MRAS observer. In summary, two signals that represent the same physical quantity, \( x \) and \( \hat{x} \), are generated by two different blocks.

The first block does not depend on the quantity to be estimated, which in the case of sensorless applications is the rotor position, whereas the second block generates a signal that depends on it. Finally, the error between these two signals, \( e \), is used to improve the accuracy of the estimation. In grid-connected applications, the stator flux is normally taken as a reference signal since it depends on the grid-voltage, which is normally very stable.
Unfortunately, the stator-flux based observer relies on the fact that the reference signal, $x$, changes relatively slowly after a reference frame transformation. This has to be the case so that a linear controller, normally a PI, can be able to track it successfully. Nevertheless, during unbalanced FRT conditions, the phase and magnitude of the machine stator-flux linkage changes relatively fast due to the now appearing negative component, which rotates negatively at double grid frequency when seen from the positive component reference frame.

Fig. 5.6 shows the simulation results of a Stator Flux MRAS Observer (SFMO) [1] during a 2-phase to ground fault (on phases $b$ and $c$). The configuration used for the
simulation is shown in Fig. 5.5, and the machine parameters are given in Table 2, which is the same from last chapter but repeated here as Table 3. The system frequency of the simulated Matlab/Simulink model is 60 Hz. The rotor side converter is controlled to maintain a constant speed, which would be determined by a Maximum Power Point Tracking (MPPT) algorithm, whereas the line side converter is controlled to maintain a constant DC-Bus voltage. The fault is introduced at \( t = 0.8 \)s and is removed 200ms later. As seen in Fig. 5.6-b, during the fault the magnitude of the stator flux is not constant. Moreover, the stator flux does not rotate at a constant angular velocity, as shown in Fig. 5.6-c, where the angular position of the stator flux is shown. As a consequence, the observer fails to properly estimate the rotor position, \( \theta_r \). Fig. 5.6-d shows the actual and estimated rotor position. Before \( t = 0.8 \)s, they overlap each other and the error is almost zero, as shown in Fig. 5.6-e. Nevertheless, by \( t \approx 0.83 \)s, the error is close to 100 deg. In this simulation, the actual rotor position, not the estimated, is used to control the rotor side converter.

### 5.1.3 SIMULATION OF PROPOSED ALGORITHM DURING FRT

According to the results of Chapter 4, the proposed sensorless algorithm should be immune to Fault Ride-Through (FRT) conditions. A Matlab/Simulink simulation model was developed to test the proposed method during FRT situations. The simulation describes a grid-connected, 4-poles, Doubly-Fed Induction Generator controlled by a back-to-back PWM converter. The configuration used was shown in Fig. 4.12, which has
been repeated again as Fig. 5.5 for convenience. Without any loss of generality, it is assumed that the line reactance has the same order of magnitude as the stator leakage reactance, i.e., \( L_g \approx L_i \). The machine parameters used in the simulations are shown in Table 2, which is repeated below as Table 3:

<table>
<thead>
<tr>
<th>TABLE 3</th>
<th>PARAMETERS USED FOR SIMULATIONS AND EXPERIMENTS</th>
</tr>
</thead>
<tbody>
<tr>
<td>DFIM Parameters</td>
<td>in (Ω)</td>
</tr>
<tr>
<td>Power</td>
<td>Poles</td>
</tr>
<tr>
<td>5 HP</td>
<td>4</td>
</tr>
</tbody>
</table>

a. Leakage inductance for stator and rotor have same numerical value.

The frequency and voltage of the injected high-frequency signal are 480Hz and 10% of the magnitude of the rotor voltage, respectively. Fig. 5.7 shows the simulation results when a 5hp DFIM is running at a speed of 30% below synchronous and a 2-ph to ground fault (in phases b and c) is introduced at \( t = 0.3s \), as shown in Fig. 5.5. The fault is removed at \( t = 0.5s \). Before the fault is removed, the voltages and currents of the machine are greatly distorted due to the now appearing negative and zero sequences. Phase \( a \) of the stator voltage goes from 170 volts-peak to 90 volts-peak, whereas the other two phases decrease to 45 volts-peak each. If any of these quantities is used as a reference in an adaptive system, they would be very difficult to track with accuracy by the conventional technique of a PI controller in a synchronously rotating reference frame. Nevertheless, the result obtained with the proposed observer shows its great degree of immunity to these kinds of disturbances, since, as shown in Fig. 5.7-d, the error between the actual and the estimated rotor position is bounded to 25 deg. This error happens due to two reasons. First, during a transient state. Second, during a transient.
state the PLL output is also a function of the magnitude of the input signal, as opposed to only the phase. Nevertheless, if more advanced filtering techniques are utilized for the high-frequency voltage on the stator, this error can be greatly reduced.

Fig. 5.8 is similar to Fig. 5.7, but a 3-phase to ground fault is simulated, rather than a 2-phase to ground fault.

5.1.4 EXPERIMENTAL VERIFICATION OF PROPOSED ALGORITHM DURING FRT

Experimental results have been pursued to verify the feasibility of the proposed method during synchronous speed, changes in the speed command, and FRT conditions.

EXPERIMENTAL SET UP

The system setup for experimental testing is shown in Fig. 5.9, and its block diagram was shown in Fig. 5.5. Two PWM back-to-back inverters are connected to the rotor of the machine.

The rotor-side PWM inverter is controlled using a frequency of 10 kHz, which is also the sampling frequency of the sensors and of the interrupt subroutine. Current transducers are used to measure the rotor currents. Two voltage transducers are used to measure the stator voltage. A 2nd order analog low-pass filter with a cut-off frequency of 863Hz filters out the switching noise but not the high frequency injected for sensing purposes.
The band-pass filter of Fig. 4.7 is implemented as a digital filter. A speed encoder of 1024 pulses per revolution is used to measure the rotational speed and rotor angle. The mechanical encoder is used only for comparison purposes.

Since a new control technique during FRT conditions is outside the focus of this paper, the conventional stator field oriented control technique is used to control the DFIM [7]. However, the current loop bandwidth has to be lower than the frequency of the injected signal; otherwise the current controllers will disturb the injected signal. A low-pass filter for the measured currents could be used that would serve the dual purpose of filtering out the injected high frequency signal, as well as the PWM noise. Low-pass filters with a cut-off frequency of 100Hz have been implemented in this experimental set-up. The DFIM

![Fig. 5.7](image)

*Fig. 5.7* Simulation results, using the proposed algorithm, in which the machine is running at a speed 30% below synchronous speed. As shown in Fig. 5.5, a 2-ph to ground fault is introduced at $t = 0.3s$ for 200ms. The results show (a) the line to neutral stator voltage, (b) the stator currents, (c) the actual and estimated rotor position, and (e) the error between these two.
is speed controlled as a motor (no prime mover is used), with the stator flux being obtained from the stator voltages and currents.

The machine parameters used for the experimental testing are shown in Table 3. A high-frequency signal of 480Hz is injected to the rotor-side windings while the stator-side voltages are measured. An estimate of the rotor angular position, which is used to control the machine, is obtained by applying the algorithm described in chapter 4 and Fig. 4.7. A microcontroller board, based on the TMS320F2812 from TI, is used to implement the proposed observer and the whole sensorless vector-control system. Since the microcontroller utilized does not have DACs, the results of the proposed algorithm (like

Fig. 5.8 Simulation results, with the proposed algorithm, in which the machine is running at a speed 30% below synchronous speed. As shown in Fig. 5.5, a 3-ph to ground fault is introduced at t = 0.3s for 200ms. The results show (a) the line to neutral stator voltage, (b) the stator currents, (c) the actual and estimated rotor position, and (d) the error between these two.
rotor angular position and speed) are saved in the microcontroller memory and then exported to Matlab for display.

TEST UNDER SYMMETRICAL 3-Φ TO GROUND FAULT

As shown in the block diagram of Fig. 5.5, a symmetrical 3-phase to ground fault is produced manually by the closing of a breaker. Fault impedances limit the maximum fault current; as a consequence, the stator voltage decreases to 0.3 p.u. during the fault rather than to zero. No new control algorithm for FRT conditions is used, but the rotor currents remain limited because a sufficiently high DC bus voltage is used. The machine is initially running at 0.9 p.u. when the fault condition is produced at \( t = 0.43 \)s. Fig. 5.10-a shows the voltage of phase \( a \) in p.u. during the instance of the fault, during the recovery, and in between. Fig. 5.10-b shows the measured and estimated rotor position information for the time range of the first enclosed area of Fig. 5.10-a, i.e., at the fault. Fig. 5.10-d shows the error, in deg., between the estimated and the measured rotor position. Figs. 5.10-c and 5.10-e present similar information to Figs. 5.10-b and 5.10-d, but focus on the instant of the

![Experimental set-up used to test the proposed algorithm. The mechanical encoder is used for verification purposes.](Image)
recovery, which is the second enclosed area of Fig. 5.10-a.

Fig. 5.10-d shows that during the instant of the fault the error remains limited to ±15 deg., whereas at the instant of the recovery, at t=1.48s, the error reaches a maximum value of 25 deg. These results are in good agreement with the simulation results shown in Fig. 5.8.

It has been observed that the error is inversely proportional to the magnitude of the injected high-frequency signal. Nevertheless, since the higher the magnitude of such signal the higher the harmonic distortion being injected to the grid, the magnitude of the injected high-frequency signal has to be limited.

Fig. 5.10 Experimental results of the proposed algorithm, for a symmetrical 3-phase to ground fault. The rotor speed is initially running at 0.9 p.u. At t=0.43s, a 3-phase to ground fault is produced, decreasing the stator voltage to 0.3 p.u. The fault is removed at t=1.47s. The results show (a) the stator voltage of phase α, (b) the measured and estimated rotor position for the time range enclosed by the blue square in (a), i.e., few milliseconds before and after the fault. The error between the measured and estimated rotor position for the same time range as (b) is shown in (d). Results analogous to (b) and (d) are shown in (c) and (e), but focusing on the instant of the recovery from the fault, which is the time range enclosed by the red square in (a).
Intuitively, better accuracy without sacrificing harmonic distortion can be achieved thanks to the utilization of better detection and filtering mechanisms for the high-frequency signal at the stator.

### 5.2 Sensitivity Against Parameter Variations

As shown in Fig. 4.7, the proposed algorithm is insensitive to parameter variations. This is a significant difference with respect to most commonly used observers, which depend on

![Figure 5.11](image1.png)

**Fig. 5.11** Simulation results showing the sensitivity to parameter variation of the commonly used SFMO when $L_s$ is increased by 10% at $t=1.0s$.

![Figure 5.12](image2.png)

**Fig. 5.12** Simulation results showing the sensitivity to parameter variation of the commonly used SFMO when $L_m$ is increased by 10% at $t=1.0s$. 

Comparison between Proposed Method and the State of the Art in Sensorless Control

100
the knowledge of the machine parameters to a certain extent. Specifically, SFMO is dependent on \( L_m \) and \( L_s \) [1]. Eq. (5.2), which is repeated below for convenience, clarifies this issue:

\[
\tilde{\lambda}_s = (L_{ij} + L_m)\tilde{i_s} + L_m\tilde{i_r}.
\]

Figs. 5.11 and 5.12 show simulation results of the position error when an SFMO observer is utilized to determine the rotor position of the system of Fig. 5.5 and Table 3. In Fig. 5.11, the SFMO is running with perfect knowledge of \( L_s \) until \( t = 1.0s \), at which point the value of \( L_s \) is increased 10%. Even though the error was bounded before this point, after the change in \( L_s \) the error in the estimated position grows up to 50 deg. by \( t = 3.0s \). It is noted that if the value of \( L_s \) is increased by a higher percentage, the observer diverges at a higher rate. Fig. 5.12 shows similar results as Fig. 5.11, but with an increase of 10% in \( L_m \) as opposed to \( L_s \).

The proposed high-frequency injection algorithm is completely immune to parameter variations since no parameters are needed in its implementation, as shown in Fig. 4.7 of last chapter.

5.3 CONCLUSIONS

This chapter compared the newly proposed high-frequency injection algorithm against one of the most widely studied sensorless control algorithms in the research literature, the Stator-Flux MRAS Observer (SFMO). It was shown that, in its standard form, the SFMO
Comparison between Proposed Method and the State of the Art in Sensorless Control

does not work during FRT conditions and that it is not immune to parameter variations. Nevertheless, the high-frequency algorithm remains fully functional during FRT conditions and is completely immune to machine parameter variations since the machine parameters are not needed for its implementation. These statements were confirmed by simulations and experimental results. This chapter also summarized the issues concerning grid connected DFIMs during low voltage ride-through conditions.

5.4 REFERENCES


CHAPTER 6

CONCLUSIONS AND FUTURE WORK

The objective of this research was to develop a new sensorless control technique for grid connected Doubly-Fed Induction Generators (DFIGs). This chapter summarizes the conclusions reached and indicates the future research opportunities made available by the development of this new technique.

6.1 DISSERTATION CONCLUSIONS

This dissertation presented a general picture of how improvement in wind power can alleviate the current energy crisis. In this regard, this dissertation introduced a new method that allows DFIGs (one of the most commonly used generators in wind power systems) to work during grid faults without the need of a position sensor, i.e. sensorless. Due to recently enacted grid codes and regulations, wind power generators should not disconnect from the grid when a fault occurs. By reviewing the currently available sensorless algorithms for DFIGs, this dissertation showed that they fail to work during grid faults. The method proposed in this dissertation solves this problem while also being immune to
The proposed sensorless algorithm works by applying a high-frequency voltage to the rotor and measuring the produced high-frequency voltage on the stator. The high-frequency voltage applied to the rotor is independent of, and superimposed on, the control voltage applied to the rotor. It was shown that the rotor position information is encoded in the phase of the space vector of the high-frequency voltage measured at the stator; hence, the proposed sensorless method is implemented by filtering out the high-frequency voltage at the stator, and determining the phase of its space vector. This results in a relatively easy implementation.

The mathematical foundations of the proposed method, as well as an algorithm for its implementation, were presented. The small-signal modeling of the proposed sensorless method, in addition to guidelines for the design of its parameters, was also included. Computer simulation and experimental results were shown in order to confirm the high accuracy of the proposed technique under grid faults and synchronous speed, respectively. The analysis and experimental results show that knowledge of the machine parameters, with the exception of the number of poles, is unnecessary in order to apply the proposed sensorless method.

Finally, several simulations were presented in which the proposed sensorless method was compared against a leading sensorless technique. The results of the simulations confirmed the superiority of the proposed method during grid faults and against machine parameter variations.
6.2 FUTURE WORK

In the small-signal analysis of Section 4.3, it was assumed that the time delay introduced by the FIR filter (phase compensation block of Fig. 4.7) can be considered negligible due to the fast operating frequency of the DSP. As a next step, this condition can be relaxed and the small-signal analysis can be performed with the inclusion of a constant time-delay block. Simulation and experimental testing of the small-signal analysis of Section 4.3, and the sensitivity analysis of Section 4.4, should be performed before this method is deployed commercially.

Finally, the impact and possible cross-interference resulting from utilizing the proposed sensorless control method in a set of DFIMs connected in parallel, like the ones found in wind farms, is an important issue that should be considered further.
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CHAPTER 1:


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CHAPTER 2:


CHAPTER 3:


Bibliography


CHAPTER 4:


CHAPTER 5:


