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Doctoral Thesis

**Model Predictive Control (MPC) for the Power Converters
for Renewable Energy Generation Systems with Switch Fault
Tolerance Capability**

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*This thesis is dedicated
to my parents for their
love, endless support
and encouragement*

Abstract

Renewable energy sources such as wind, solar, wave and geothermal, have appeared quite recently as new sources to supply the electrical energy demand. These renewable energy sources are more interesting than fossil fuel sources since they are clean, free to use, available in many places, naturally replenished and have lower bad impact on the climate and our planet. The electrical energy obtained from these renewable energies is produced by means of power electronic converters and electrical machines. Power converters should be controlled with microprocessors in order to extract this energy. Among the different control strategies that exist for the power converters, the model predictive control (MPC) is a promising strategy that has a fast dynamic response. However, it suffers from the variable switching frequency and needs a high sampling frequency.

In this Ph.D. thesis, a novel model predictive control (MPC) for the power converters of doubly fed induction generators (DFIGs) and surface permanent magnet synchronous generators (SPMSGs) is presented. In this method, four voltage vectors are selected in every switching period and their duration times are estimated to track the reference values. The appropriate voltage vectors in each period are recognized when the estimated duration times of the selected active vectors are positive. The vectors are selected in this method without using the position of the grid voltage or the machine fluxes. Since four voltage vectors are employed by the converter in this method, fixed switching frequency, and low current THD are achieved, while the fast dynamic response of the MPC remains in this strategy. Moreover, there is no need to use a phase-locked loop (PLL) for the proposed predictive power controls. First, the proposed predictive direct power control (PDPC) for the rotor side converter (RSC) of DFIG is presented, both for balanced and unbalanced grid voltage. The unbalanced grid voltage situation is considered for the RSC of DFIG since the DFIG is very vulnerable to the unbalanced grid voltage due to the direct connection of the stator windings to the grid. The proposed PDPC can easily follow the power references under normal and abnormal voltage conditions, even if the power references contain oscillation terms. Without any additional controller, PI or resonant controllers, the proposed PDPC strategy can obtain smooth stator active and reactive powers or smooth electromagnetic torque or could inject sinusoidal and balanced currents into the grid when unbalanced voltage appears in the stator windings of the DFIG. In order to validate the performance of the PDPC, simulation and experimental studies are carried out.

Second, the MPC is extended to the machine side converter (MSC) of the SPMSG. As a result, a model predictive direct current control (MPDCC) for the MSC, which tracks the current references in the synchronous frame, is presented. Moreover, a simple model of an oscillating water column (OWC) power plant is presented, in order to later validate the proposed MPDCC applied to an OWC system. This application is chosen because it is a difficult case study that demands a fast torque control to handle the power take-off (PTO) system. The proposed method for the MSC of SPMSG is tested and analyzed through simulations in the Matlab/Simulink environment and in a customized SPMSG based laboratory setup.

Third, a novel model predictive power control is proposed for the GSC of the DFIG and the SPMSG. In this method, the voltage vectors are selected without using the grid voltage sector, and the bidirectional operation of the GSC is considered. Simulation studies for a GSC of 2 MW DFIG and a GSC of 5kW SPMSG are carried out. Furthermore, the performance of the proposed control for the GSC is evaluated in a laboratory setup. The simulation and experimental results for the proposed MPC methods for the two-level power converters of DFIG and SPMSG show excellent performance during transient and steady-state conditions. Moreover, the proposed methods are compared with previous MPC methods and the results indicate that the proposed MPC controls have lower current THD and faster dynamic response.

Moreover, in this thesis, the fault tolerance capability for the power converters of DFIG and PMSG is considered and a new model predictive control for the fault-tolerant topology is presented. The four-switch three phase (FSTP) converter has been chosen as fault-tolerant topology and the proposed MPC control is designed according to this new converter topology. In these methods, three voltage vectors are selected in each switching period to achieve a fixed switching frequency.

First, the proposed MPC control for the FSTP MSC of SPMSG is presented. The proposed method follows the current references with great accuracy whereas the switching frequency of the IGBTs is fixed and low. This method minimizes the current reference tracking error, and its fast response makes it suitable for the power take-off (PTO) systems, present in wave energy converters (WECs). The system features a fast capacitor voltage offset suppression control without using any filter. The dynamic performance and the voltage offset control of the proposed strategy for FSTP converter feeding a SPMSG, are evaluated in the Simulink environment and on a laboratory SPMSG prototype. Furthermore, the capability of the proposed method to harvest the maximum energy from irregular waves is assessed using an OWC power plant emulator.

Second, predictive power control for FSTP GSC is presented. In this new method, the ripple of the active and reactive powers of the GSC is minimized. Furthermore, a compensation power to eliminate the DC voltage deviation in the capacitors, which can be achieved without using any low pass filter, is presented. The proposed strategy has been evaluated in Matlab/Simulink environment and afterward, it was implemented in a laboratory prototype.

The simulation and experimental results of the proposed predictive strategies for the FSTP MSC of SPMSG and FSTP GSC of DFIG show that the proposed methods are capable to suppress the DC link voltage offset, featuring balanced currents and a fast-dynamic response, while keeping a low current THD. Moreover, the proposed MPC methods are compared with two other MPC methods and the results showed that the proposed MPC strategies for FSTP converters are suitable to control FSTP converters having a better current THD.

Resumen

Cada vez está más extendido el empleo de las energías de origen renovable, como la eólica, solar, geotérmica y, en menor medida, las energías marinas, como fuentes de generación para abastecer la demanda de energía eléctrica. Las características de estas fuentes de generación, como la ausencia de emisiones, su libre disponibilidad, y su reposición natural, son factores que contribuyen a reducir el impacto que tiene la producción de energía eléctrica sobre el medioambiente y su repercusión sobre el cambio climático.

La gran mayoría de los sistemas de generación con energías renovables, tienen en común que precisan convertidores electrónicos de potencia en su sistema de conversión en energía eléctrica, habitualmente conectados a máquinas eléctricas rotativas. Dichos convertidores electrónicos, precisan microprocesadores para controlar adecuadamente la cantidad y la calidad de la energía eléctrica que suministran estas instalaciones al sistema eléctrico. Entre las estrategias de control que se implementan en los microprocesadores de los convertidores para conseguir este objetivo, el control predictivo basado en modelos (MPC) supone un método muy prometedor, debido a la rapidez que consigue en la respuesta dinámica del sistema. No obstante, esta técnica tiene como inconveniente que las señales de disparo que reciben los IGBT's del convertidor conmuten a frecuencia variable, con los inconvenientes que de ello se derivan, como una mayor distorsión armónica, mayores pérdidas en conmutación y elevada frecuencia de muestreo.

En la primera parte de esta Tesis Doctoral, se presenta un novedoso control predictivo basado en modelos (MPC) aplicable al sistema de control de los convertidores electrónicos para sistemas de generación de velocidad regulable que estén constituidos, tanto por Generadores de Inducción Doblemente Alimentados (DFIGs), como por Máquinas Síncronas de Imanes Permanentes Superficiales (SPMSGs). Esta estrategia de control consiste en seleccionar una secuencia de cuatro vectores espaciales de tensión en cada periodo de conmutación, donde, además, se calcula el tiempo de aplicación de cada uno de ellos para seguir al valor de referencia de la variable de control. Para comprobar que la secuencia de vectores tensión elegida en cada periodo es la correcta, el tiempo de aplicación calculado para cada vector debe ser positivo. Este método tiene importantes ventajas. Por una parte, la selección de la secuencia de vectores se realiza sin necesitar conocer el ángulo del vector tensión de red ni el campo magnético en la máquina. Esto implica evitar la inserción de sistemas del tipo phase-locked loop (PLL) para realizar esa tarea. Por otra parte, el hecho de que este método emplee una secuencia de cuatro vectores en cada periodo, permite conmutar a frecuencia fija, lo que reduce la tasa de contenido de armónicos THD en la corriente inyectada por el convertidor, a la vez que mantiene las extraordinarias prestaciones que presenta la tecnología MPC en su respuesta dinámica. En esta tesis, aplicando la técnica descrita, se han desarrollado e insertado sistemas de control específicos para cada uno de los convertidores de las instalaciones de generación del tipo DFIG y SPMSG, a la vez que comprobado su eficacia y operatividad.

En primer lugar, se presenta el desarrollo de un nuevo Control Directo de Potencia Predictivo (PDPC) para el Convertidor conectado en el Rotor (RSC) en un sistema de Generación del tipo DFIG, a la vez que se analiza su funcionamiento tanto en redes con sistemas de tensión equilibrados como desequilibrados. Se ha considerado especialmente esta última situación de desequilibrio, porque la tecnología DFIG es especialmente vulnerable a desequilibrios de tensión en la red, al estar las bobinas del estator de la máquina, directamente conectadas a dicho sistema de tensiones. Se ha comprobado como el sistema PDPC desarrollado, con ausencia de reguladores adicionales, ni del tipo PI ni Resonantes, sigue perfectamente la señal de referencia de la potencia activa y reactiva, tanto en condiciones de equilibrio como frente a fuertes desequilibrios en el sistema de tensiones aplicado por la red, incluso en el caso de que las propias señales de referencia presenten oscilaciones. Se comprueba asimismo, que el sistema PDPC propuesto consigue una respuesta amortiguada, tanto en el seguimiento de las referencias de potencia activa y reactiva, en el seguimiento de la referencia del par electromagnético, así como en conseguir inyectar un sistema de corrientes por las tres fases totalmente senoidal y equilibrado, si así se le requiere. Esto se consigue, tanto cuando hay aplicado un sistema de tensiones equilibrado, como desequilibrado, en el estator del DFIG. Todo ello se ha validado tanto en estudios con modelos de simulación como en ensayos reales sobre instalaciones experimentales en el laboratorio.

En segundo lugar, el empleo de método MPC se hace extensivo para desarrollar un nuevo Control Directo de la Intensidad Predictivo basado en modelos (MPDCC) ahora para aplicarlo al Convertidor conectado en el lado de la Máquina Eléctrica (MSC) de un sistema de Generación del tipo SPMSG. En este caso el sistema de control sigue las referencias establecidas para las ondas de corriente en las tres fases en un sistema de referencia en sincronismo con el giro de la máquina. Adicionalmente, para validar el funcionamiento del sistema MPDCC, se ha aplicado al convertidor de un generador SPMSG instalado en un accionamiento que emula el funcionamiento de un sistema de generación marina por extracción de energía de las olas del tipo Columna de Agua Oscilante (OWC). Se ha elegido esta aplicación, por los requisitos especialmente exigentes en la velocidad de respuesta del sistema de control, para regular de forma adecuada el punto de funcionamiento del dispositivo de extracción de energía de las olas (power take-off (PTO) device). En este caso, también se ha validado el funcionamiento tanto en estudios, empleando modelos de simulación, como en ensayos reales sobre instalaciones experimentales en el laboratorio.

En tercer lugar, se emplea el método MPC, para realizar un Control Directo de Potencia Predictivo (PDPC) para el Convertidor conectado en el lado de la Red, para ser instalado, indistintamente, en cualquier de los dos tipos de sistema de generación DFIG o SPMSG. La particularidad de este control es que, para seleccionar la secuencia de vectores tensión a aplicar en cada periodo de conmutación, no precisa conocer el sector de posición del vector espacial de la tensión de la red. Adicionalmente, el sistema de control permite el funcionamiento bidireccional del convertidor (rectificador-inversor) indistintamente. Se ha validado el funcionamiento empleando modelos de simulación de dos convertidores GSC, de dos niveles de potencia 2MW y 5kW instalados en sistemas de generación DFIG y SPMSG.

Por otro lado, también se ha validado el sistema en ensayos reales sobre instalaciones experimentales en el laboratorio.

Como conclusión de esta primera parte, decir que los resultados de las pruebas realizadas, tanto sobre modelos de simulación, como sobre equipos experimentales de laboratorio, permiten afirmar que el nuevo método de control MPC, permite desarrollar sistemas de control específicos para convertidores de dos niveles, cuya respuesta muestra un excelente comportamiento tanto en régimen permanente como durante grandes perturbaciones transitorias. Las pruebas comparativas con métodos de control anteriores, también de tipo MPC, muestran que el nuevo método consigue corrientes con menor contenido de armónicos (THD), a la vez que permite más rapidez en la respuesta dinámica del sistema.

En la segunda parte de esta Tesis Doctoral, se analiza la capacidad de tolerancia al fallo, para diversas topologías de los convertidores electrónicos de dos niveles instalados en un sistema de generación del tipo SPMSG, en caso de fallo en alguna de sus ramas. Se selecciona la topología de tres fases y cuatro interruptores, four-switch three phase, (FSTP) como topología más idónea para la operación del convertidor en caso de fallo. Posteriormente se desarrolla un nuevo modelo de control predictivo (MPC) para aplicar a dichos convertidores con la topología resultante en condiciones de fallo. De forma similar a los casos anteriores, en este control se realiza una selección de secuencia de tres vectores espaciales de tensión a aplicar en cada periodo de conmutación, lo que permite conmutar a frecuencia fija. Aplicando la técnica descrita, se han desarrollado e insertado estos sistemas de control en convertidores específicos.

En primer lugar, se aplica a un Control Directo de Corriente Predictivo (PDCC) en el Convertidor conectado en el estator de la Máquina (MSC) en un sistema de Generación del tipo SPMSG. El método propuesto opera a un valor constante y reducido de la frecuencia de conmutación de los IGBTs, con lo que logra seguir con gran precisión la referencia de las corrientes en las tres fases. El sistema logra un control muy rápido de la supresión de offset de tensión de los condensadores del enlace de DC, sin precisar filtro alguno. Tanto el funcionamiento dinámico, como las prestaciones del control de offset de tensión del condensador, han sido validados tanto con modelos de simulación como en ensayos reales sobre instalaciones experimentales en el laboratorio. Adicionalmente, para validar especialmente su capacidad de respuesta dinámica, se ha aplicado este control al convertidor del accionamiento experimental que emula el funcionamiento de un sistema de generación marina (OWC), anteriormente descrito. Se ha comprobado cómo el sistema es capaz de seguir con precisión y rapidez la consigna de operación en el punto de extracción de máxima potencia (MPTT) del dispositivo, incluso para un perfil de olas totalmente irregular.

En segundo lugar, se aplica ahora un Control Directo de Potencia Predictivo (PDPC) al Convertidor conectado en el lado de red (GSC). Se consigue aquí minimizar el rizado de la potencia activa y reactiva inyectada a la red por el GSC. Se consigue eliminar completamente el desequilibrio de tensión en los condensadores de la etapa de continua, sin necesidad de filtrado. Todo ello se ha validado tanto en estudios con modelos de simulación como en ensayos reales sobre instalaciones experimentales en el laboratorio.

Como conclusión de esta segunda parte, decir que se han desarrollado con éxito nuevos sistemas de control para grupos de generación tipo SPMSG, para aplicar a los convertidores, tanto del lado de máquina (MSC) como de red (GSC), cuya topología responda a un esquema FSTP para posibilitar continuar su operación en caso de fallo de una rama del convertidor. Los resultados de las pruebas realizadas, tanto sobre modelos de simulación, como sobre equipos experimentales de laboratorio, han permitido validar su excelente funcionamiento, en tanto a que, suprimen el offset en la tensión de los condensadores en el enlace de continua, mantienen el sistema de corrientes trifásica totalmente equilibrado, consiguen una rápida respuesta dinámica, a la vez que un bajo contenido en armónicos (THD). Así mismo los estudios comparativos con métodos de control anteriores, también de tipo MPC, muestran que el nuevo método consigue corrientes con menor contenido de armónicos (THD), a la vez que más rapidez en la respuesta dinámica del sistema.

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Introduction

1.1. Introduction

In the last few decades, the use of power electronics converters and microcontrollers has increased considerably in a wide range of applications such as wind turbines, photovoltaic systems, driving electrical machines and energy storage systems. This increment of using power converters is due to the features brought by the power converters to these systems, such as an improved performance and a higher efficiency together with a better control of different variables. As a result, power electronic converters have become one of the main actors in industry when converting electrical energy and when harvesting the maximum energy available from a renewable energy source. In order to contribute to increment the benefits of power electronics in the renewable electric generation field, new control strategies for power electronic devices, including certain fault tolerant topologies, have been developed.

In order to control the power converters of the renewable energy generation systems, different control methods have been proposed, such as linear and hysteresis controllers, artificial intelligence, sliding mode control and predictive controls. Generally, the classical controllers, like linear PI and nonlinear hysteresis controls (such as direct power control or direct torque control), are very simple to be implemented and are robust, while the newer control strategies like fuzzy, neural and sliding mode controls are more complicated and need more calculations and estimations when compared to the classical ones. The following paragraphs present some of the main characteristics of these two controllers.

Linear controllers like PI controllers with pulse width modulation or space vector modulation are applied to control the power converters of renewable generation power systems. The field-oriented control (FOC) and voltage-oriented control (VOC) are some of this type of controls that are widely used in many power converters connected to wind generators. The fixed switching frequency is achieved in this control by the modulation stage. On one hand, the coordinate transformations are vital in this type of controller and on other hand, employing a linear controller for a nonlinear system could result in an unacceptable performance of the system along the different range of operation points. Moreover, in these controllers, the PI controllers should be tuned and have a slow dynamic response. Additionally, the different system constraints and requirements cannot be employed directly in the linear controllers.

Hysteresis controllers use the switching states of the semiconductors and consider the nonlinear behavior of the converters by comparing the references variables with their measured values and, according to a hysteresis band, the switches states are selected. Although it is robust to parameters variation in the system, a high sample time and a high switching frequency is required to have an acceptable performance. Moreover, these

controllers produce a spread harmonic spectral content. As a result, expensive filters are necessary to be installed to reduce the harmonics. Recently, some modifications to obtain a fixed switching frequency for this type of controllers have been reported. However, the proposed controls are still far away to be employed by the industry.

As a result, the new controls that could overcome the disadvantages of classical controllers have become more necessary than ever. With the improvement of the microprocessors, new control strategies such as predictive controls have been developed. This type of control strategy seems to be an interesting method to control the power converters since the model of the system is considered to predict the future behavior of the variables and the error in the predicted variables is removed by the use of an optimization function. Among different types of predictive controls, the deadbeat predictive control and model predictive control are the more renowned.

1.2. Objectives of the doctoral thesis

Model predictive control (MPC) is a very powerful control strategy which has been employed in power converters in recent years. This control can bring many advantages to the system such as 1-fast dynamic response, 2-constraints and nonlinearity can be included in the control system 3-easy to understand 4-can be extended to different systems. However, it suffers from a high number of calculations and a variable switching frequency. Hence, in order to have good performance, model predictive control should work with fast microprocessors and with high sampling time and switching frequency. As a result, the switching power losses are high in this method. Many researchers tried to remove these disadvantages from the MPC and proposed to use multi-vectors MPC in order to improve the current THD and to decrease the switching losses in the converter. However, the obtained duration time for the multi-vector MPC could result in a negative duration value of a vector which doesn't have any meaning and will make the controller more complicated. In this type of multi-vectors model predictive controls, the vectors are selected according to the grid voltage sector or machine flux sector. As a result, when the power converter is connected to a weak grid, which could have a negative sequence or harmonics, the vectors are not selected correctly.

Consequently, it is essential to improve the MPC in order to work with a fixed switching frequency and to work perfectly during different grid voltage disturbances and switch faults. Otherwise, the switching power losses for high power converters would be enormous and could not be applied in industry, meaning that during the disturbances and switch faults, the wind turbine or the power converters should be disconnected from the grid, which is not acceptable according to the grid codes.

In accordance with the above paragraph, in this thesis, new MPC strategies for power converters of doubly fed induction generators (DFIGs) and surface permanent magnet synchronous generator (SPMSGs) have been developed in order to reach the three main objectives: first, to have fixed switching frequency and low current THD with the fast dynamic response characteristic. Second, to operate perfectly in disturbed grids regarding the rotor side converter (RSC) of DFIG. Finally, to be able to operate correctly in case of partial fault in the converter switches.

These new MPC control strategies can be applied to variable speed wind turbine generation (WTG) systems (DFIG and SPMSG wind turbines). The proposed method has constant switching frequency feature while the fast dynamic response of model predictive control remains in the control. The proposed method for the rotor side converter of DFIG can easily follow the power references under normal and abnormal voltage conditions, even if the power references contain ac terms. Without any additional controller, such as proportional resonant controller, the proposed technique can obtain smooth stator active and reactive powers or smooth electromagnetic torque or could inject sinusoidal and balanced currents into the grid when unbalanced voltage appears in the stator windings of a DFIG. Moreover, the MPC has also been developed for the machine side converter (MSC) of SPMSG as well as for the grid side converter of both types of the mentioned wind generation systems.

Furthermore, in the proposed new MPC strategies, a fault tolerance capability of the system has also been considered for the case of losing one of the legs of the power converter due to the breakdown of a switch. This case is especially important for the offshore wind turbines and for wave energy applications, such as oscillating water column (OWC), which are difficult to be repaired in a short time. In this case, the four-switch three phase converter topology is considered and the MPC is improved according to this power converter topology. As a result, the renewable energy system can continue injecting the harvested power to the grid even under a breakdown of one of the switches.

Moreover, the OWC emulator is used in this thesis in order to check the performance of the MPC control for the SPMSG during a large variation of the rotor speed and to see if the proposed method is capable to fulfill the requirements that are needed for the OWC power generation systems.

During the Ph.D. study and while working to improve and implement the MPC for the DFIG in the laboratory, testing the different controls, protecting the rotor and stator windings became also a big concern. As a result, after investigations and studies, a new differential protection was proposed in order to protect the whole induction machine during the internal faults. This new DFIG protection method resulted very effectively and quick in response. Although this protection was not the main objective of the Ph.D. study, a novel differential protection was carried out. The idea of this protection is explained in Appendix.B of the doctoral thesis since it is out of the scope of the thesis. However, this protection is very interesting.

1.3. Structure of the doctoral thesis

This document describes in detail the content and scope of the work developed in this thesis and has been structured as follows:

In the present chapter1, the problem to be solved and its context has been described. In chapter 2, first, different renewable energy systems and their topologies (wind energy and wave energy systems) are briefly explained. Second, different control strategies that have been proposed for the three phase voltage source converters of renewable energies are described. Finally, the switch fault tolerance topologies for the three phase power converters and their control strategies are clarified. In Chapter 3, the proposed new four-voltage vector

MPC for the power converters of DFIG and SPMSG, which could work perfectly in weak grids, is presented. In Chapter 4, three-voltage vector MPC strategy is presented for the switch fault tolerant topology, four-switch three phase converter, of the DFIG and SPMSG. Simulation and experimental results of the proposed methods for the six switches and four switches three phase systems are carried out in Chapter 5 and Chapter 6, respectively. In these chapters, the effectiveness of the proposed MPC strategies is evaluated and validated by many tests and is also compared to previous MPC controls. Finally, the conclusions and future works are presented in Chapter 7. In the Appendix, the proposed differential protection for the DFIG are explained.

1.4. Summary of publications

Some of the contributions around the control strategies that are presented in this doctoral thesis have been presented and published in different IEEE journals and conferences while some others are under review. The journal articles that have been published are listed below:

- 1- M. E. Zarei, C. Véganzones Nicolás and J. Rodríguez Arribas, "Improved Predictive Direct Power Control of Doubly Fed Induction Generator During Unbalanced Grid Voltage Based on Four Vectors," **IEEE Journal of Emerging and Selected Topics in Power Electronics**, vol. 5, no. 2, pp. 695-707, June 2017. DOI: 10.1109/JESTPE.2016.2611004
- 2- M. E. Zarei, C. V. Nicolás, J. R. Arribas and D. Ramírez, "Four-Switch Three-Phase Operation of Grid-Side Converter of Doubly Fed Induction Generator With Three Vectors Predictive Direct Power Control Strategy," **IEEE Transactions on Industrial Electronics**, vol. 66, no. 10, pp. 7741-7752, Oct. 2019. DOI: 10.1109/TIE.2018.2880672
- 3- M. E. Zarei, D. Ramírez, C. Véganzones Nicolas and J. Rodriguez Arribas, "Three-phase four-switch converter for SPMS Generators based on Model Predictive Current Control for Wave Energy Applications," **IEEE Transactions on Power Electronics**, vol., no, pp., 201. DOI: 10.1109/TPEL.2019.2911209
- 4- M. E. Zarei, C. A. Platero, C. V. Nicolás and J. R. Arribas, "Novel Differential Protection Technique for Doubly Fed Induction Machines," **IEEE Transactions on Industry Applications**, vol. 55, no. 4, pp. 3697-3706, July-Aug. 2019. DOI: 10.1109/TIA.2019.2914886
- 5- M. E. Zarei, D. Ramírez, C. Véganzones Nicolas and J. Rodriguez Arribas, "Predictive Direct Control of SPMS Generators in Wave Energy Applications" **IEEE Transaction on Power Electronics**, conditionally accepted, 2019

Introduction to Power Converters employed in Renewable Energies Power Generation systems, Control Strategies and Switch Fault-Tolerant Methods

2.1. Introduction

Renewable Energy sources such as wind, solar, wave and geothermal, have emerged as new sources to supply the electrical energy demand. These renewable energy sources are more interesting than fossil fuel sources since they are clean, more accessible, and available in many places, naturally replenished and they do not have such a bad impact on the climate and on our planet. The renewable energy sources are usually used for electrical energy generation, heating, and remote and rural power supply. The electrical generation from renewable energies has grown significantly in the recent decades. Twenty two percent of all electricity production in the world is provided by different types of renewable energies [1]-[2]. Due to advances in technology, which lead to decreasing costs, and government incentives for clean energy, power generation systems using renewable energy sources are becoming one of the most important technologies employed to produce electrical energy.

In section 2.2 of this chapter, wind and wave renewable energy technologies and their power conversion topologies are briefly described. Later, in section 2.3, the basic concept of different control strategies that have been proposed for the three phase voltage source converters used with renewable energy generators are explained. Section 2.4 of this chapter will present the fault-tolerant methods for the three phase grid-connected power converters and their control methods. Finally, the summary and conclusion of this chapter are presented.

2.2. Renewable Power Generation and Power Converters

This section includes a brief review of the main electrical power generation technologies employed in wind energy and wave oscillating water column (OWC) power systems. The characteristics and operation of these renewable power plants are discussed and the schematics and the topologies of the power converters and the electrical machines that are used in these kinds of power generations are presented. Moreover, the dynamic models of the electrical generators employed in these installations that will be used in the following chapters, are expressed in this section.

2.2.1. Wind energy

Wind power is one of the most important renewable energy source that supplies considerable electrical energy in the world. Fig. 2.1 shows the capacity of the installed wind turbine systems from the beginning of the 21st century until the end of 2020 [3]. The wind power capacity in 2016 was 477GW and it is predicted to reach up to 792 GW at the end of 2020. Among the different types of renewable energies, only the wind power was supplying 55% of the renewable power generation in the world in 2015. This means that 3.7% of the electricity production in the world was produced by wind power generation in 2015 [4]-[6]. In the Western Europe, the penetration of wind power generation is very high if compared with the rest of the world. For example, the wind power penetration in Denmark, Portugal, Ireland, Spain, Germany are 42%, 23.2 %, 23%, 18%, and 13.3%, respectively, which indicates that these countries have obtained a high level in the world using wind power generation. Other countries like the USA, China, and Brazil also achieved to supply 4.7%, 3.3% and 3% of the power generation by wind power. An interesting case that happened in Denmark on November 2013, was that, for the first time in the world, the wind power generation was higher than the total power consumption in the country, although this case is now quite usual to happen in Denmark [7].

Fig. 2.1. World wind power generation capacity installed from the beginning of the 21st century until the end of 2020 [3].

Moreover, the power rating and the size of the wind turbine systems have been increased significantly from 1980 until now. To give an illustration, the power rating of the wind turbines is shown in Fig. 2.2 from the beginning of 1980 [3]. In this figure, the rotational speed and the usage of power electronics are also included. As it can be seen in this figure, from the beginning of 2000 the size of the wind turbines has been increased notably due to the development of the power electronics allowing the technology of the nowadays variable speed wind turbines. For example, the typical power rating of onshore and offshore wind turbines in Europe was 2.7 MW and 4.2 MW in 2015, while the biggest wind turbine in the market is a 8MW wind turbine [8].

Fig. 2.2. Power rating of the wind turbines from the beginning of 1980 until now with the development of power electronics [3].

Nowadays, many big manufacturers are producing wind turbines between 4 MW to 6 MW and it is predicted that more high rated power wind turbines will be installed soon. For example, 10 MW wind turbines will be installed in the next decade in order to decrease the cost of producing electricity [9].

China has nowadays the largest market for wind turbines, where 30 GW of wind power were installed in 2015. Also, the USA and European Union markets in 2015 installed about 8.6 GW and 12.8 GW, respectively. Therefore, 82% of the total market of wind power plants is in these countries [4]. The top different wind turbine manufacturers in the world with respect to the size and power rating are shown in Table 2.1. As it can be seen, China has a first position in the market while its turbine ratings are between 1.5-3 MW. The main rating of the turbines for all manufacturers is between 1.5 MW to 3 MW while the higher ratings of wind turbines that exist are between 5 MW to 8 MW.

Table 2.1. Different wind turbine manufacturers and their market shares.

Rank	Manufacturer	Global Market shares	Rotor diameter	Power range
1	Goldwind (China)	12.5 %	70-121 m	1.5-3 MW
2	Vestas (Denmark)	11.8 %	90-136 / 164 m	1.8-3.45 / 8 MW
3	GE Wind (USA)	9.5 %	83-137 / 150 m	1.7-3.8 / 6 MW
4	Siemens Wind (Denmark/Germany)	8.0 %	101-142 / 154 m	2.3-4 / 6-8 MW
5	Gamesa (Spain)	5.4 %	80-132 / 132 m	2-3.3 / 5 MW
6	Enercon (Germany)	5.0 %	44-141 m	0.8-4.2 MW

2.2.1.1. Wind turbine system (WTS) topologies and power conversion systems

In order to convert the wind energy to electrical energy, different components such as turbine blades, gearbox (in the direct-drive system it is not necessary), generator and power converters, as well as transformers, are needed [10]. The schematic of converting the wind energy to electrical energy and injecting the extracted power to the grid in a modern WTS is shown in Fig. 2.3.

Fig. 2.3. Different steps for converting the wind energy to electrical energy [3].

The WTSs can be classified into two groups according to the operation speed and the generator type or the aerodynamic aspects such as pitch control [11]. Therefore, WTSs are classified as fixed speed and variable speed wind turbine systems.

The fixed speed and the partial variable speed wind turbines with variable rotor resistance, are the two common classical wind turbines that are classified as fixed speed wind turbines. In the fixed speed wind turbine, a squirrel cage induction machine is employed as generator, and it is directly connected to the grid without using any power converter. In this topology, capacitor banks are needed to provide the reactive power that is demanded by the induction generator. Moreover, the fluctuations in the wind will be transferred to the electrical grid with this topology. In partial variable speed wind turbines, a wound rotor induction generator is used, and a controllable resistance is connected in series with the rotor windings in order to have some variable speed feature. The partial variable speed is achieved in this topology by using the resistance. However, it is also needed to use a capacitor bank to provide the reactive power consumed by the induction machine.

The development of the power electronic equipment leads to introduce two new wind turbine systems that allowed the system to operate at variable speed. These two systems are Doubly Fed Induction Generators (DFIGs), with partial-scale power converters, and Permanent Magnet Synchronous Generators (PMSGs) or squirrel cage induction generator (SCIG), both with full-scale power converters. The topology of the DFIG wind turbine was dominating in the last decade in the wind turbine industry. However, the PMSG wind turbine is now gaining much attention since the generator is isolated from the grid by the power converters and it is completely controllable [4]. In the following sections, these two types of wind turbine systems are described in more detail.

2.2.1.1.1. DFIG with Partial scale back to back converter

A typical schematic of a DFIG with partial scale back to back converter is shown in Fig. 2.4. This wind turbine system was widely installed in the industry from the beginning of year 2000 until now, being one of the most interesting wind turbine technologies. In this concept, the turbine is connected to the generator via a gearbox and the stator winding is connected to the grid directly as can be seen in Fig. 2.4. The rotor windings are connected to the grid through a back to back converter. In this type of WTS, the power converters rating is approximately 30% of the generator power rating [13]-[15]. The current and the frequency of the rotor is controlled by the rotor side converter in order to extract the maximum energy from the wind and, as a result, the wind turbine will operate at different speeds. Moreover, the mechanical losses are reduced with this topology. The partial-scale power converter makes the system be cost-effective since it can operate in a wide range of speeds. However, the need of using a gearbox and slip rings in the rotor circuit of the generator, together with its vulnerability against grid disturbances, are the main drawbacks of this topology [16]-[18]. Usually, for this type of wind turbines, two power converters, which are usually two-level voltage source converters, are connected to each other in a back to back configuration through a DC link as shown in Fig. 2.4. As a result, a four-quadrant operation and smooth connection can be achieved by these power converters easily. Furthermore, it has a higher efficiency compared to the fixed or partially variable wind turbines.

Fig. 2.4. Typical schematic of the DFIG based wind turbine with partial scale power converters[3].

In order to control the DFIG, the dynamic model of the generator should be considered. The dynamic model of the DFIG in the stationary frame ($\alpha\beta$ frame) and in the synchronous frame (dq frame) are shown in Fig. 2.5 and Fig. 2.6, respectively. In these two figures, $v_{\alpha s}$, $i_{\alpha s}$, $\psi_{\alpha s}$ and $v_{\beta s}$, $i_{\beta s}$, $\psi_{\beta s}$ are the stator voltages, stator currents and stator fluxes in α and β axes, respectively. $v_{\alpha r}$, $i_{\alpha r}$, $\psi_{\alpha r}$ and $v_{\beta r}$, $i_{\beta r}$, $\psi_{\beta r}$ are the rotor voltages, rotor currents and rotor fluxes in α and β axes, respectively. Moreover, v_{ds} , i_{ds} , ψ_{ds} and v_{qs} , i_{qs} , ψ_{qs} are the stator voltages, stator currents and stator fluxes in d and q axes, respectively, while v_{dr} , i_{dr} , ψ_{dr} and v_{qr} , i_{qr} , ψ_{qr} are the rotor voltages, rotor currents and rotor fluxes in d and q axes, respectively. R_s and R_r are the stator and rotor resistances, $L_{\sigma s}$, $L_{\sigma r}$ are the stator and rotor leakage inductances and L_m is the magnetizing inductance. ω_m , ω_s and $\omega_r = (\omega_s - \omega_m)$ are the electrical rotor speed, synchronous speed, and slip speed, respectively.

Fig. 2.5. The dynamic model of the DFIG in the stationary frame ($\alpha\beta$ frame).

Fig. 2.6. The dynamic model of the DFIG in the synchronous frame (dq frame).

2.2.1.1.2. PMSG and asynchronous generator with full scale back to back converter

Another variable speed wind turbine concept is shown in Fig. 2.7. In this concept, the stator winding of the generator is connected to the grid through a back to back converter and as a result, the generator is isolated from the grid. In this topology, the generator can be PMSG or SCIG.

However, the PMSG is more interesting than the induction machine due to its higher efficiency. In this wind turbine system, the active and reactive power are completely controlled by the power converters. The turbine can also be a direct driven multipole generator where the gearbox can be discarded. Thus, this topology is simpler, since the gearbox can be removed, the speed and the powers are fully controlled, and it is robust under grid disturbances. However, it should be noted that since the power converters ratings are the same as the generator rated power, the power converter of this system is more expensive than that in the DFIG based wind turbine. Hence, the power losses in the power converters are also

higher than those in DFIG. Furthermore, if the voltage level of the system is low, less than 1 kV, and the power rating is high, the design of the cables will be a challenging aspect [3]. Hence, using two parallel back to back converters or multilevel converters are suggested to overcome the mentioned problems. It is proposed to use the configuration of the multicell converters as a solution for high-power rating wind turbines especially for the systems above 3MW [19]-[20]. Two typical schematics of this multicell converters are shown in Fig. 2.8 [3]. As can be seen, one schematic is for a multi winding generator and the other one is for a three phase winding generator. As a result, in these schematics, the power rating of the converters has been reduced.

Fig. 2.7. PMSG or asynchronous generator with full scale back to back converter wind turbine.

Fig. 2.8. Two typical examples of Multicell converters for PMSG wind turbines: (a) for multi winding PMSG wind turbine; (b) Three phase winding PMSG wind turbine [3].

The dynamic model of the PMSG in the synchronous frame is illustrated in Fig. 2.9. Where V_d , V_q and I_d and I_q are the stator voltages and currents in d and q axes. R_s , L_d , and L_q are the stator resistance and the stator inductance in d and q axes, respectively. λ_m and ω symbols represent the permanent magnet flux and electrical rotor speed, respectively. According to the dynamic model of the PMSG, the generator is fully controlled, and the maximum energy is normally achieved by the machine side converter of this wind turbine.

Fig. 2.9. Dynamic model of the PMSG in synchronous frame.

2.2.1.2. Control strategies for variable speed wind turbine systems

Both DFIG and PMSG variable speed wind turbines consist of two voltage source power converters, as shown in Fig. 2.4 and Fig. 2.7. The converter that is directly connected to the grid is called grid side converter (GSC) and the other one that controls the generator is called machine side convert (MSC), for PMSG, or rotor side converter (RSC), for DFIG based wind turbines.

The control of these power converters should consider different subsystems such as electrical and mechanical parts (blades, generator, and the converters). The following control structures are necessary for the DFIG based wind turbines or PMSG with full-scale power converters.

Firstly, the input and output power of the system should be controlled and managed. On one hand, the input power of the system, mechanical power, can be controlled and limited by adjusting the pitch angle of the blades and, on the other hand, the injected power to the grid should be controlled according to the standards and the received commands that are provided by the system operator. Other control functions, such as the maximization of the power, improving the power quality of the grid, reserving active and reactive power, can be achieved by the power converters of the WTS. For example, the MSC, which is controlling the generator, can be adjusted in order to absorb the maximum energy from the turbine. Moreover, the control also should consider the low voltage ride-through operation working perfectly during normal and abnormal grid voltage, according to the grid codes.

Fig. 2.10. Different layers of control for the power converters of WTS.

The control system of a wind turbine can be divided into three layers, as can be seen in Fig. 2.10. The basic control loops of WTS, such as current and voltage controls, dc bus regulators, Phase-locked loop (PLL), are primary controls and should be very quick [21]-[23]. During abnormal grid voltage, such as unbalanced grid voltage or grid faults, an integrated control is needed for the MSC, GSC, choppers or crowbars and pitch angle of the blades [22]-[24]. These integrated controls are the secondary controls. Furthermore, the system operators can demand some other extra control requirements, such as increasing the power inertia, reserving some amount of power, or improving the quality of the injected power, during the operation of the WTS. Hence, the control system that is designed for the WTS should be prepared to satisfy all these requirements.

2.2.2. Wave energy: Oscillating Water Column Systems

There are many different physical phenomena such as waves, tides, ocean and sea currents etc, that can be used to extract energy from the sea [25]. Many different devices and patents have been presented in order to extract sea energy and convert it to the electricity. However, most of them, when tested under actual conditions, did not show good performance and were discarded. Although many simulation and laboratory results have been presented showing the effectiveness of wave energy conversion systems, the real prototypes that are installed in the sea usually present many problems [26]. As a result, the conversion of the wave energy into electricity is still a challenging area and many studies are taken place to solve the problems.

Among different wave energy converter systems, one of the most interesting systems is the oscillating water column (OWC), since it can be designed in different power rating ranges, it has suitable efficiency and it does not have any bad impact on the environment [27]-[28]. Furthermore, another advantage of this technology is that the generator and the mechanical parts of this system are not inside the seawater which leads to having a more reliable system [29]. This system was started to develop in 1974 with powerful generators. However, this system was expensive, and the reliability was low. In recent decades, the development of power electronics and the development of new efficient generators made the OWC power plants to be a very promising technology to provide a small part of the electricity demand. Fig. 2.11 shows a typical schematic of the OWC system. In this type of power plant, the pressure exerted by the waves inside a chamber produces an airflow that drives an air turbine connected to an electric generator. Despite the alternating airflow, most of the power take-off (PTO) systems used with OWC devices maintain a continuous direction of rotation, for instance, using a Wells turbine. The OWC can be installed onshore and offshore. However, the onshore is preferred due to its easier maintenance, robust structure, and easier grid connection.

Fig. 2.11. Schematic of an OWC system.

2.2.2.1. The electrical part of the OWC system

The main objective of the OWC is to generate an airflow in order to convert the wave energy to electrical energy like in the wind generators. Hence, wind turbine concepts can also be utilized in OWC systems. However, it should be noted that the OWC system will confront a harsh environment due to the storms and sea weather, and the mechanical stresses are higher in this application. Therefore, when the generator and the converters are designed, these kinds of issues should be considered. For example, it has been suggested to use direct-drive generators to avoid the use of a gearbox.

Most of the development in the OWC until now was concentrated on the mechanical part of the PTO system. This part has been developed sufficiently and the wave energy can be effectively converted to pneumatic energy by the PTO mechanism. Hence, the researchers are now developing the electrical part of the PTO system of OWC in order to optimize the whole system and to improve the global efficiency [30].

Nowadays, the wind turbine system is a very advanced technology and the offshore wind turbine technology is facing the same conditions as the fixed OWC power plant can be confronting. As a result, the technology used for wind turbines can be extended to the fixed OWC system. However, the floating OWC system has a rigid mechanical part and not all kind of generators can be applied for this system [31].

Fig. 2.12. The schematic of the electrical part of the first generation of OWC power plant in Vizhinjam [35].

Many different generators such as brushed and brushless induction machines and PMSG were used from the beginning of the OWC development [32]-[33]. The airflow produced by the mechanical part of an OWC has a large variation and it changes widely in a few seconds. As a result, a variable speed generator concept and a fast dynamic control are needed to absorb the maximum energy from the airflow. The electrical model schematic of the first OWC technology that was used in Vizhinjam OWC power plant is shown in Fig. 2.12 [34]. A three phase PMSG, a rectifier, a DC motor, and a three phase induction machine were used to convert the airflow to electrical energy. Later, the DFIG with partial-scale converters and the PMSG/SCIG with full-scale converters were used in OWC power plants. The schematic of a DFIG and PMSG with back to back converters for OWC system is the same as those used in wind turbine generations, Fig. 2.4 and Fig. 2.7. A variable speed DFIG featuring vector controlled back to back converters is used in a 296 kW OWC power plant in Mutriku, Spain, to obtain the maximum energy from the waves [34]. It should be noted that by considering the conditions of the OWC system, the PMSG with full-scale converters are

more interesting than the DFIG. The PMSG could be used without gearbox and it is more robust when facing the sea conditions.

In summary, the three phase voltage source converters are used extensively in wind turbines, OWC power plants and other renewable power plants like solar photovoltaic systems. Therefore, in the following section, the different control strategies that have been proposed for these converters are presented.

2.3. Control strategies for the three phase voltage source power converters

2.3.1. Introduction

The control strategies for the power converters started to develop in parallel with the semiconductor improvements, even before the development of the control platforms [35].

In order to have a controllable switch very close to the ideal switches, the fully controllable semiconductors like power MOSFETs and IGBTs have been introduced and developed since the 1960s. In those days, the pulses for semiconductor switches were generated in analog control platforms using the hysteresis controller strategy or modulation stages. Ten years later, more advanced controllers have been presented to provide a higher performance of the system where huge calculations and complex techniques were needed. However, implementing these controllers in analog circuits was difficult due to the low computational capability of the analog systems.

To implement control strategies, an adequate control platform is needed. This control platform is an important part since the control strategy should be loaded and run inside it. In 1983, the digital signal processor (DSP), a digital control platform, was introduced by Texas instruments allowing to easily implement advanced controllers with large computation calculations in control platforms by using data samples in series. From this point, real-time digital control platforms started to develop and to improve significantly in the industry in order to allow the capability of implementing complex controllers to have better performance and higher efficiency [36].

Nowadays, there are many different digital control platforms, such as DSP, microcontrollers, and field programmable gate array (FPGA). The Floating-point and fixed-point DSPs are two common types of Texas Instruments DSPs, such as the TMS320F28335 and the TMS320F2812, where usually the floating points are more expensive. FPGAs are quicker and have a flexible hardware structure which can work at high speeds. Xilinx and Altera are one of the two most common type of FPGA in the market now [36]-[37].

The dSPACEs, rapid prototyping platform, such as dSPACE DS1104 or dSPACE DS1103, are advanced DSPs microprocessors that allow implementing the control directly into the DSP from a Matlab Simulink software. These powerful control platforms are suitable for research purposes when a fast implementation is necessary to test a new strategy in an experimental laboratory bench.

In the following section, different control techniques (classical and advanced controls) for the two-level three-phase power converters are briefly explained and finally, the advantages and disadvantages of these control strategies are summarized in this section.

2.3.2. Classical control strategies for the power converters

In this part, two traditional and well-known control techniques for power converters are briefly presented. These control techniques can be divided into two main methods which are hysteresis and linear controllers. These control techniques have been traditionally implemented to the MSC, RSC, and GSC of the PMSGs and DFIGs [13], [38]-[39]. These controls are mainly described as a current controller in this section. However, these classic controllers can be easily extended to power controllers or torque and flux controllers.

2.3.2.1. Hysteresis controller

The hysteresis controller is a nonlinear controller in which the main objective is to keep the controlled variables errors inside a hysteresis bound. In order to achieve this, the switch states of the power converter should be changed according to the hysteresis comparators signs [40]. Fig. 2.13 shows a basic hysteresis controller in “abc” frame for a two-level three phase voltage source converter. In this figure, the measured currents (the measured grid currents or the machine currents) are compared to the reference currents and then the errors go to the hysteresis comparators. The states of each switch will be changed by the output flag of this comparators. When the error in each phase is higher than the upper limit, $\delta/2$, the upper switch will turn on and the lower switch will turn off to bring the error of the respected phase back inside the hysteresis band. In contrast, when the error is lower than the lower limit, $-\delta/2$, the upper switch will turn off and the lower switch will turn on in this case. As a result, the current error of each phase will be forced to remain inside the hysteresis band. Obviously, by decreasing the bandwidth of the hysteresis controller, the current error will be decreased, but, in return, the switching frequency will be increased.

Fig. 2.13. Schematic of the hysteresis current controller for a two-level three phase voltage source converter.

As can be seen in Fig. 2.13, this method is very simple, robust and could be easily implemented in the microprocessors. This technique has been implemented in the industry for many years, when analog platforms were used to control the power converters [41]. This

method has a fast dynamic response time and it is robust to parameter variations due to the inherent control structure.

This method has been developed to control directly active and reactive power, such as in direct power control (DPC) based on a lookup table [42], or to control directly torque and flux [43], or to control directly currents in dq frame, such as in direct current controls (DCC) [44]. A block diagram of the direct power control (DPC) for the grid-connected converter is shown in Fig. 2.14. In this method, the power errors pass through hysteresis controllers. According to the output of these hysteresis controllers and the sector where the grid voltage is located, a voltage vector will be selected by a lookup table and later on it will be implemented by the converter. Direct torque control (DTC) for electrical machines is very similar to the DPC method. However, the state of the switches is selected according to the output of the hysteresis controller of the torque and flux errors, and the rotor or stator sector positions. These methods are also easy to be carried out by the control platforms.

However, it should be noticed that the switching frequency of all these methods are variable and is dependent to the hysteresis band and the parameters of the machine or the grid as well as the operating points. These are very important drawbacks of this method which can cause many problems. Some of these problems are: a. It can lead to having a resonances problem [45], b. The switching power losses are higher than usual, c. The current will have a wide range of different harmonics especially the low order harmonics which are difficult to be removed by the filters. Moreover, a very high sampling frequency is needed to achieve a good performance of the system. Many types of research have been done to solve the variable switching frequency problem. It was also proposed to use a delta modulation where the values of the errors were compared at a fixed switching time. However, all the presented methods lost the simplicity of the structure. Moreover, in this method, the constraints cannot be included in the control system [2].

Fig. 2.14. Block diagram of the DPC method for a grid-connected converter.

2.3.2.2. Linear Control

The linear control is a very well-known control structure in the industry and has been widely implemented in different applications such as wind energy, wave energy, and electrical drives. This method has very interesting features such as fixed switching frequency and low current THD.

Fig. 2.15. Shows the schematic of the linear current controller that consists of a PI controller and a modulator used for generating the switching pulses [45]. The modulation stage for producing the switching signals in this type of control system, makes it possible to control the converter linearly while the power converter structure is nonlinear. Thanks to the modulation stage, a fixed switching frequency is achieved with this method. In this case, the proportional integral (PI) controllers can be utilized in order to control the system.

Fig. 2.15. Schematic of a linear current controller.

As can be seen in Fig. 2.15, the current error, that is, the error between the measured value and the reference value, goes to a PI controller. The output of the PI controller will produce the reference voltage that passes to the modulation stage to produce the switching pulses of the power converter. As a result, the performance of the linear controller is depended on the PI values and the modulation technique. In the modulation stage, the switching pulses are generated according to a carrier waveform, the DC link voltage, and the reference signal. There are two main modulation strategies to generate the pulses from the reference voltage, which are the pulse width modulation (PWM), and the space vector modulation (SVM) techniques. These modulation techniques include a carrier AC waveform which leads to having a fixed switching frequency. The SVM is more efficient than the PWM due to the lower harmonic values that are generated and a higher maximum modulation index. However, it should be noted that the SVM is more complicated to be implemented than the PWM, especially for multilevel converters.

The PI controllers have been widely used to control electrical drivers and renewable power generation system. The field oriented control (FOC) is a linear control for electrical machines which controls the flux and the torque while using decoupling terms. A typical block diagram of this method is shown in Fig. 2.16. In this method, the machine flux is oriented along the d axis so its magnitude is regulated in the d axis, while the speed/torque is regulated in the q axis. In this method, the synchronous frame (dq frame) is used to control the variables with two or four PI controllers. Finally, the decoupled terms are added to the

control system to generate the voltage reference before the outputs are sent to the modulation stage.

Fig. 2.16. Schematic of a field oriented control for electrical machines.

The voltage oriented control (VOC) or vector control (VC) is another linear control that has been implemented in many power converters of renewable generators [38]. In this strategy, the active and reactive powers, or the currents that are representing the active and reactive powers, are controlled independently. The typical VOC control for a grid-connected converter is illustrated in Fig. 2.17. In this technique, the grid voltage is used to be along one of the d-q axes while the two control variables are controlled in a synchronous frame with two or four PI controllers. Like in the FOC control, the decoupling terms are added to the control system to control the active and reactive powers independently.

Fig. 2.17. Schematic of the voltage oriented control for a grid-connected converter.

2.3.3. Advanced control strategies for the power converters

In addition to classical methods, many advanced controls have been proposed by the researchers. In this part, some of these important advanced control methods are presented.

2.3.3.1. Sliding Mode Control

One of the important nonlinear control methods which are developed for the power converters is the sliding mode control (SMC), which belongs to the variable structure control family [46]. In this method, the control uses discontinuous signals to change the dynamics of the power converters by making the system to slide into a particular surface in the state space of the system while the control variables are kept into a predefined surface [47]. Fig. 2.18 shows the basic concept of the SMC method for a two-level three phase voltage source converter. As can be seen, similar to the linear control, a modulation stage is needed to generate the pulses. However, in this method instead of using a PI controller, a SMC control is used to generate the reference voltage while considering the nonlinear nature of the power converter and discrete time control. The SMC method also depends on the parameters of the system and the power converter. Hence, by changing the values of the parameters the performance can be changed. The controller structure of a SMC will be adjusted according to the law sets that have been introduced in the control system, in order to have a stable and fast dynamic response during different operation points and disturbances.

Fig. 2.18. Current sliding mode control structure for controlling the two-level three phase voltage source converter.

Although the SMC method has good features such as well and fast dynamic response and fixed switching frequency during different operation points, it suffers from very complicated mathematical equations and harmful chattering phenomena [48]. Consequently, the SMC method is not implemented widely in industrial applications.

2.3.3.2. Intelligent Control method

The intelligent control technology is another type of advanced control systems that can be divided into different strategies such as Artificial Neural Networks, Genetic Algorithms, Expert and Planning Systems, Intelligent and Autonomous Control, Bayesian probability, Fuzzy control, etc [49]. In this section, the basic concept of a fuzzy control and artificial neural networks used to regulate the current variables, are presented briefly.

2.3.3.2.1. Fuzzy Logic Control

The block diagram of the nonlinear fuzzy logic controller (FLC) for regulating the current is shown in Fig. 2.19. The reference voltage is generated according to the current error, the derivative of the current error, and the FLC control. In this strategy, the expert knowledge and humans experience are implemented in the system to control the power converters, by a set of “if-then” control rules. The FLC method is completely robust during parameters variations and it is a very interesting control system to be used when the exact parameters of the system and the power converters are unclear.

Fig. 2.19. Basic block diagram of a fuzzy logic control for a two level three-phase voltage source converter.

The FLC strategy can have a better dynamic performance than the conventional PI Controllers and can be used during a wide range of operation points. However, it should be noted that accurate control rules and complete heuristic construction knowledge are essential parts of this method in order to have a good performance.

2.3.3.2.2. Artificial Neural Networks Based Control

Artificial neural networks are an attempt to simulate the human thinking process which is achieved by the neural networks of the humans. This method has a high degree of parallelism with human way of thinking and it has a learning capability which makes it different from other control strategies. The basic structure of this method is shown in Fig. 2.20. In this strategy, the current error will pass through a scaling factor and later on it will enter the artificial neural networks modulator to generate the pulses [50]. The modulator will generate the signals for the switches according to the neurons network and their corresponding weights. In order to obtain good dynamic performance in this method, the neurons will adjust and modify the weightings factors during the training process. With this method, a fixed switching frequency is achieved. Nevertheless, in order to make the controller work in a real practical system, it should pass the offline training process before being implemented.

Fig. 2.20. Schematic of an artificial neural networks controller for a two-level three phase voltage source converter.

The combination of fuzzy logic control and artificial neural networks have also been reported by researchers in the literature [51]. This combination shows that the performance of the system can be improved even more by this strategy. The major advantage of the intelligent controllers is that they do not need the system parameters and that they are robust during the variation of the system parameters. However, the designer of this controller needs to have a very accurate knowledge of the power converter and the different operation points to design the controller perfectly. Consequently, this type of controllers is not used as the power converters controllers in high power conversion systems [52].

2.3.3.3. Predictive Control Strategy

The basic idea of the predictive control was firstly presented in the 1950s to be applied for slow systems [53]. Then, it was used in the oil and chemical industry to solve some problems and later on it was spread out in these industries [54]. By introducing the microprocessors, especially the DSPs, the predictive controls started to develop in power electronic converters. The first attempting of applying predictive control in the power electronics industry goes back to the 1980s [55]-[56]. Later on, many different predictive controls were reported for the power electronic converters. Hence, the name of predictive control presents a wide range of controllers such as deadbeat predictive control, hysteresis based predictive control, trajectory-based predictive control and model based predictive control [57]. However, all these controls have a common basis, which is predicting the behavior of the system for the next sampling time according to the model of the system and, depending on this prediction and on the defined objectives for the system, the switching pulses will be generated.

These predictive controls can be divided into two main groups which are [57]-[58]:

A- Predictive control with modulation stage

B- Predictive control without modulation stage

The deadbeat predictive control, the generalized predictive control, and the continuous control set model predictive controls, they all fitted to the first category of predictive control, while the hysteresis based model predictive control, the direct model

predictive control, and the finite-control-set model predictive control, are classified as the second type of predictive controls [62]-[63].

2.3.3.3.1. Deadbeat predictive control (predictive control with Modulation stage)

The deadbeat predictive control was presented in 1980 as the first type of predictive controls for power converters and now is a very popular strategy among various types of predictive controls [57], [59]-[60]. The block diagram of this type of predictive control is shown in Fig. 2.21. In each sampling period, the reference voltage is estimated according to the load and the converter model, in order to have a zero-reference tacking in the next sampling instant. In this strategy, the future performance of the system is predicted, and the optimal voltage is generated. Later on, this reference voltage will be employed by the modulation stage. In this method the dynamic response is fast, and the switching frequency is fixed due to the use of SVM or PWM modulation [60]. Nevertheless, the deadbeat predictive control is very sensitive to the system parameters variations.

Fig. 2.21. Overall block diagram of the deadbeat predictive control.

The deadbeat control is very simple to be implemented and does not need any cost function and optimization process or any tuning like the classic PI controllers. Nonetheless, the measurement noises, the delays in the control and the perturbation have a negative influence in this strategy which can lead to instability problems. Moreover, if the parameters are not correctly estimated, there will be an error in tracking the references during the steady-state. Many investigations have been done to remove some of these problems and to make the control be more robust during noises and parameters variations [57].

2.3.3.3.2. Finite control set predictive control (model predictive Control without modulation stage)

The model predictive control (MPC) or finite control set MPC (FSC-MPC) started to be implemented in power electronic applications during the last decade due to the development of fast microprocessors, since this kind of controls need a very fast sampling time. The schematic of this type of controller, model predictive control, is shown in Fig. 2.22. This method presents a different approach for producing the switch pulses without using any PI controller and modulations. Unlike the other controllers that try to decrease the current error in the upcoming sample times, the model predictive control implements a strategy to

eliminate the error before the next sampling time. As a result, at the beginning of the next sampling time, the error will be removed by the MPC [61]. Many different types of MPC have been presented and all of them consists of the following parts:

- 1- Estimation of references at $k+1$ instant according to the k instant reference values.
- 2- Predict the variables at $k+1$ instant according to the measurement values at k instant.
- 3- Using a cost function for the predictions in order to produce the switching pulses.

These three parts can be seen in Fig. 2.22. Since in two-level converters the number of the switches are finite, the control variables at $k+1$ instant can be calculated for the different switching states. The typical cost function for a current controller can be shown as:

$$g(k) = |i^*(k+1) - i(k+1)| \quad (2.1)$$

Where i is the current and superscript $*$ indicates the reference value. According to this cost function and the values that can have this cost function for different switching states, the switching state that minimizes this cost function will be selected in each switching period and will be employed by the power converter. The MPC has a good reference tracking in the steady-state as well as a fast dynamic response. The MPC has more advantages compared to the deadbeat control. The MPC is more flexible than the deadbeat control since the cost function can be modified and other objectives, with different weighting factors, can be presented inside the cost function. However, in the deadbeat control, the reference voltage is directly estimated according to the system parameters and the measurement values in each period, without using any cost function or optimization process. It should be noted that the computation burden in the MPC strategy can be increased compared to other control strategies due to the online optimization process. However, the microprocessor computational capacity has been improved significantly in the recent years so the aforementioned issue is not a big concern anymore.

Fig. 2.22. Schematic of the model predictive control for the two-level three phase voltage source converters.

2.3.4. Summary of different control strategies for the power converters

The overall advantages and disadvantages of the mentioned classical and advanced controllers are classified in Table 2.2. The controllers have been compared according to

different aspects, such as the need of modulation stage, dynamic response and fixed or variable switching frequency of the control strategies [2]. The classical linear controllers such as PI controller, vector controls, and FOC controls are still extensively used for controlling the power converters of the renewable power generation systems and electrical drivers due to their advantages that can be seen in Table 2.2. The advanced controllers, as mentioned before, need many complex calculations and many frame transformations which require a powerful microprocessor.

Table 2.2. Summary of the advantages and disadvantages of classical and advanced control methods for the power converters of renewable energies

	Hysteresis	Linear	SMC	Intelligent	DBPC	FCS-MPC
Control Complexity	Low	Medium	High	High	Medium	Low-Medium
Model & Parameters	Not Needed	Needed	Needed	Not Needed	Needed	Needed
Prior Knowledge	Not Needed	Not Needed	Not Needed	Needed	Not Needed	Not Needed
Modulation Stage	Not Needed	Needed	Needed	Needed	Needed	Not Needed
Constraint Inclusion	Not Possible	Not Possible	Possible	Not Possible	Not Possible	Possible
Dynamic Response	Excellent	Average	Good	Good	Good	Excellent
Switching Frequency	Variable	Fixed	Fixed	Fixed	Fixed	Variable

The linear PI control is usually employed in the synchronous rotating frame with a space vector modulation strategy. The hysteresis control is a very easy and simple strategy to be implemented. The hysteresis based predictive control was an attempt to achieve the benefits of the hysteresis control. Among the mentioned controllers, only the intelligent and hysteresis control is robust to the parameter variations, so the system and the converter parameters are not important. However, the intelligent control needs to be designed carefully by an expert that knows the model very well. The constraints can be included in the controls of the power converters in the model predictive and the sliding mode control. The hysteresis and model predictive controls have the best dynamic response, but both of them suffer from the variable switching frequency. Nevertheless, in the cost function of the MPC method, the switching frequency can be considered. The important key for the advanced control strategies to be implemented in the real industry cases is: a. not to be complicated and b. to have better performance compared to the linear controllers. The finite control set MPC strategy seems to be the right advanced method to be more developed since it can satisfy these issues and have a good performance during different conditions.

2.4. Fault-tolerant methods for the power converters and their control strategies

2.4.1. Introduction

As mentioned in the earlier sections, the semiconductor switches, such as IGBTs and MOSFETs, are playing an important role nowadays in extracting power from renewable energy sources and in electrical drivers. Nevertheless, these components, as well as the capacitors in the power converters, are more vulnerable compared to the other parts of the system, which leads to having low reliability of these components [62]-[63]. Since there are not usually any extra switches and capacitors in the power converters, any breakdown or other problem in these components make the whole system to be stopped and shut down. This issue is very dangerous in some applications like the electric vehicles where the converter could stop working while the vehicle is on the road running at high speed. This case would carry a risk in the health of the passengers [64]. In other applications that are used, for example, in hospitals and military equipment, this kind of failure brings other potential risks due to power converters [65]. Consequently, the fault tolerance capability of a system is an important issue for the power converters in order to increase its reliability. Hence, many investigations have been conducted in this area.

When a fault happens in the system, the first thing to do for a system that has a fault tolerance capability is to diagnose the fault in the system [66]. Fast and correct diagnosis of the fault and a later protection of the system will prevent tragic and economic consequences of the fault in a power converter. In order to have this capability, after the detection, the fault should be isolated and later on, the system should be reconfigured depending on the design of the hardware of the system. Moreover, the control of the system should be changed according to the new configuration of the power converter. Many different topologies and solutions have been presented in the literature for the fault-tolerant operation of the power converters for different applications.

According to the hardware redundancy type, the fault-tolerant techniques can be divided into four types [67]:

- 1- Leg level, where the redundant legs are installed in parallel or series to the principal legs [68]-[71]
- 2- Switch level, where the redundant switches are placed in parallel or series, or a dc bus midpoint is used [72-76]
- 3- Module level, such as modular multilevel converters [77]
- 4- System level, where a redundant converter is connected in parallel or series to the main converter [78-79]

Among these four types of fault-tolerant techniques, the switch level solution is easier and cheaper compared to the other solutions. The dc bus midpoint connection or four-switch three phase topology, where the faulty phase is connected to the midpoint of the capacitors, is the cheapest and well-known fault-tolerant solution. In this thesis, this type of topology is selected as the fault-tolerant solution and the corresponding control method is presented.

In the following parts, first, the fault isolation strategies will be briefly presented, later, the four-switch three phase converter topology will be described, and finally, the different control strategies that have been employed for this topology will be discussed.

2.4.2. Fault isolation methods

The important step and the first action that should be carried out when a fault has been detected, is to isolate the fault and to remove/disconnect the faulty switches or faulty legs. If the fault is a short circuit fault, this is a more difficult thing to solve. The isolation strategy makes the damaged and broken switches or legs to be disconnected and isolated from the rest of the system. After this isolation, the new configuration can take place in the system. The designer of the fault tolerance capability of a system should take into account the fast and accurate process needed to change from the pre-fault to the post-fault configuration, as well as the influence of the new topology in the operation of the converter, and, also important, the cost of the structure and the complexity of the system. As a result, according to the application, some of these aspects are more important than others at the design level.

Fig. 2.23(a) to Fig. 2.23(e) show different methods for isolating the faulty switch of a two-level converter. These methods can be extended for the three-level converters or other types of converters. In [80], it is suggested to use two thyristors, two capacitors, and two fuses to isolate the faulty switches, Fig. 2.23(a). In this technique, the two switches or one switch can be isolated from the system when the fault is detected. For example, in Fig. 2.23(a), the upper switch, S1, can be isolated when the lower Thyristor, T2, is triggered. In this case, the upper fuse F1 will blow, and the upper switch will be disconnected from the converter. The duration time of isolating the faulty switch in this strategy depends on the auxiliary capacitor values. Although this method is simple, it suffers from the parasitic inductance that is induced by the fuses and from a high cost, due to using many components. This strategy is simplified in [81], Fig. 2.23(b). In this method, one Triac and two fuses were used to isolate the faulty switch. In this method, if the faulty switch is the lower switch, the upper switch will be blocked and later the Triac will be triggered and the faulty switch will be disconnected by the lower fuse. Nevertheless, this method could not solve the parasitic inductance.

Another method is shown in Fig. 2.23(c) [82]. This strategy was an attempt to remove the parasitic inductance produced by the fuses in the two previous strategies. In this case, only one fuse is used, and it is located in the output of the phase. In this situation, the fuses are in series with the converter inductances and the aforementioned problem will not appear. In this technique, when the faulty switch is located, the other switch will be blocked and, by triggering the Triac, not only the fault switch will be disconnected, but also the whole faulty leg will be isolated from the converter by the fuse. It should be noted that this method is not applicable when both switches fail to work.

In Fig. 2.23(d), using a switch, such as an IGBT, Triac or relay, is suggested for isolating the faulty leg instead of using a fuse. However, this technique is more expensive than the previous methods and the efficiency is lower. Also, in this method, the whole leg will be isolated. By developing the switches and using the fast fuses already present in industry, it is possible to use a very simple method to isolate the fault, as can be seen in Fig. 2.23(e). In this technique, if a short circuit happens in one of the switches, the other switch

will be connected making the fuses to trig and so the leg will be isolated. This is a cost-effective method, but there will be parasitic inductance problems.

Fig. 2.23. Different methods for isolation the faulty switch [67].

2.4.3. Reconfiguration to four-switch three phase converter or DC bus midpoint connection

The four-switch three phase converter (FSTPC) or DC bus midpoint connection is a cost-effective solution for two-level three phase converters. In this case, after the fault is detected in the two-level three phase power converter and the faulty leg is isolated from the rest of the converter, the remain two phases will help the system to keep operating. This configuration is used when one of the switches or one of the legs of the converter has failed.

In [83], three different fault-tolerant schematics are suggested for electrical machine drivers as a switch level solution. These topologies are shown in Fig. 2.24. The first topology, Fig. 2.24(a), is suggested to connect the faulty phase to the dc bus midpoint of the converter by using Triacs after the clearance of the fault. This topology is a very well-known schematic and, as can be seen, is a four-switch three phase converter. One of the advantages of this method is that the converter can still provide the nominal current and the nominal torque after the fault. However, this topology has some limitations as follows:

- 1- The maximum voltage that can be produced by the converter will be decreased to the half if the dc bus voltage is not changed in the new topology.
- 2- In order to be able to change to this configuration, the DC bus midpoint of the converter should be accessible.
- 3- In this method, one of the phase currents will pass through the capacitors, as a result, if the current values are high, the capacity of capacitors should have high values.

Fig. 2.24. Different DC bus midpoint connection topologies as a fault-tolerant solution for electrical machine drives.

The second topology, Fig. 2.24(b), shows that the midpoint of the dc bus can be connected to the neutral point of the electrical machine winding by only using one Triac. In this method, the magnitude of the remained phase currents will be increased by $\sqrt{3}$ and the phases will be shifted as shown in Fig. 2.25 for the post-fault. In this figure, it has been assumed that the faulty phase was phase “a”. This topology is also having some drawbacks as follows:

- 1- The semiconductor switches should be designed to be oversized which increases the cost of the converter.
- 2- The maximum voltage of the line to line voltage will decrease to the half which will lead to having a maximum speed of half of the nominal speed of the machine.
- 3- In order to use this topology, the dc midpoint of the DC bus and the neutral point of the machine should be accessible.
- 4- Oversized DC capacitors are necessary due to passing a higher neutral current.

Fig. 2.25. Diagram of the currents before and after the fault related to Fig. 2.24b [67].

The third topology is shown in Fig. 2.24(c). In this schematic, the midpoint of an extra leg is considered to be connected to the neutral point of the motor by a Triac. It should be noted that in many applications, a four-leg converter is used to drive the electrical machines and the fourth leg is connected to the neutral point of the machine without any Triac in normal operations. As a result, if one of the switches of a phase is broken, the proposed topology can be used to control the motor. This topology has the following advantages:

- 1- There is no need to have access to the midpoint of the dc bus and the capacity of the capacitors is not oversized.

- 2- The maximum voltage that can be produced by this topology can be the nominal rated voltage of the converter.

For the voltage source converters connected to the grid, the four-switch three phase converter idea can be applied, as shown in Fig. 2.26. In this method, the phase angle of the other two phase voltages will be regulated somehow to have balanced line to line voltages. For example, as can be seen in Fig. 2.26(b), if the faulty phase is phase “a”, the phase angles of the two other phases will be shifted as shown in this figure. It should be noted that in this topology the voltage that can be produced by the converter will be reduced to $1/\sqrt{3}$ of the normal operation of the six-switch three phase power converter.

Fig. 2.26. Four-switch three phase grid connected power converter [67].

A similar approach using Triacs can be used for other types of converters, such as five and six leg converters [84], matrix converter [85], and other topologies. However, two-level three phase converters are more common in the industry and the four-switch three phase converter topology seems to be a very interesting topology for the post fault situation since it is a cost-effective topology. Hence, in the following subsection, different control strategies for this topology in literature are presented.

2.4.4. Control strategies for four-switch three phase converters

Different control methods for the four-switch three phase (FSTP) converter have been presented in the literature. In [86] and [87], an easy and feasible vector based PWM method is suggested for the four-switch three phase (FSTP) inverter. The Lyapunov function based nonlinear controller is used to follow the inverter current in “abc” frame and to obtain a decoupled active and reactive power control [87]. Nevertheless, this method needs a PWM control and coordinate transformations.

In order to reduce the unbalanced currents in FSTP inverters and to remove the grid voltage sector location, an adaptive PWM was presented in [88]. The common voltage and the effect of the AC current passing through the DC link capacitors, as well as the current THD, were investigated when the zero voltage was applied to the FSTP rectifier, and a unified space vector modulation was presented [89]. However, sector identification is necessary for this algorithm. In [90], the unbalanced DC link voltage in FSTP rectifiers is eliminated by using an optimized modulation strategy where the capacitor current is minimized. This

strategy was improved by a hybrid space vector modulation, and the current stress in the capacitors was decreased in [91].

Since the MPC is an advanced control method that has many advantages compared to the classical control methods, many researchers tried to extend and develop this control method for the FSTP rectifiers or inverters. This thesis presents an improved model predictive control for the six-switch and four-switch three phase power converters. Consequently, in the following paragraphs, different MPC strategies that have been presented in the literature for FSTP converters, both as rectifiers or as drivers for electrical machines, will be presented.

In [92], predictive direct torque control is presented for a four-switch inverter to control an induction machine. The schematic of the circuit diagram of the converter that is connected to the induction machine is illustrated in Fig. 2.27.

Fig. 2.27. Circuit diagram of the four-switch three phase converter connected to an induction machine [92].

In Fig 2.27, it is assumed that the phase “a” is connected to the midpoint of the dc bus. Also, in this work, the four voltage vectors that can be produced by the four-switch converter are presented and it has also been mentioned that since the phase “a” is passing through the DC bus capacitors, the dc capacitors will have fluctuation according to the machine speed. In this strategy, the DC link voltage offset that can appear in the capacitors of the DC bus is investigated and taken into account in the cost function. In this MPC, the following cost function is proposed for the four-switch converter for driving an induction machine:

$$G_i = \frac{|T_e^* - T_e(k+1)|}{T_{nom}} + \lambda_1 \frac{||\varphi_s^*|| - ||\varphi_s^*(k+1)||}{||\varphi_{nom}||} + \lambda_2 \frac{|V_{dc1}(k+1) - V_{dc2}(k+1)|}{V_{dc}} \quad (2.2)$$

Where T_e^* , $T_e(k+1)$ and T_{nom} are the reference value, estimated value at instant k+1 and nominal value of the electromagnetic torque, respectively. φ_s^* , $\varphi_s^*(k+1)$ and φ_{nom} are the reference value, estimated value at instant k+1 and nominal value of the stator flux, respectively. V_{dc1} , V_{dc2} and V_{dc} are the upper, lower and total capacitor voltages. λ_1 and λ_2 are the weighting factors in this cost function.

In this cost function, the dc offset is presented inside the cost function with a weighting factor in order to suppress it. As a result, the weighting factors for the offset control and the stator flux should be selected.

In this strategy, according to the measured values of the capacitor voltages and the three phase currents and the machine speed, the torque, the stator flux and the capacitor voltages are predicted in every period for each of the four possible voltage vectors. Then, the cost function value is estimated for all these predictions, and the voltage vector that makes the cost function to have a minimum value will be selected and will be employed. It is obvious that since only one voltage vector is applied in this method, the switching frequency will be variable, and it will depend on the switching states of the next periods. The block diagram of this model predictive torque control is shown in Fig. 2.28. As can be seen, according to the current values and the machine speed, the stator and rotor fluxes are estimated and later the stator flux and torque for the next step will be predicted for each of the possible voltage vectors. The proposed model predictive torque control has shown that it is capable to follow the references. However, the values of the weighting factors are very sensitive in this method. It has been shown that for some range of the weighting factors the proposed strategy does not work very well and makes the whole system to be unstable. Moreover, the parameter variations and sensitivities are not reported for this approach.

Fig. 2.28. Block diagram of the proposed model predictive torque control for an induction machine [92].

A model predictive current control for a four-switch rectifier is presented in [93]. The proposed method controls the currents in the stationary frame and suppresses the voltage offset. In this method, the voltage offset was also considered inside the cost function. The cost function in this work is proposed as follows:

$$G_i = (i_{\alpha}^*(k) - i_{\alpha}(k+1))^2 + (i_{\beta}^*(k) - i_{\beta}(k+1))^2 + \lambda_d(V_{dc1}(k+1) - V_{dc2}(k+1)) \quad (2.3)$$

Where i_{α} and i_{β} are the currents in the stationary frame ($\alpha\beta$ frame). V_{dc1} and V_{dc2} are the upper and lower capacitor voltages, respectively. The superscript * denotes the reference value. λ_d is the weighting factor for the differences between the two capacitor voltages.

In this method, the optimal voltage vector that makes the above cost function to have a minimum value is selected every switching period. The results showed that the proposed control can achieve a balanced current while the voltage offset in the capacitors is removed using a reasonable weighting factor for the offset control. The proposed method can track the references very fast and have acceptable performance while a unit power factor can be achieved. However, this method suffers from variable switching frequency and high current THD value, since only one voltage vector can be applied in every switching period.

The two-vector based model predictive current control for the four-switch converter connected to a synchronous reluctance machine is proposed in [94]. In this method, in each of the switching intervals, two voltage vectors are applied in order to emulate the performance of the six-switch converter. This method is robust to machine parameters variation and even the capacitor voltages are not needed. By the combination of two suitable voltage vectors, eight voltage vectors can be achieved in this method, like in the six-switch three phase converters. Hence, one voltage vector will be applied half of each period and another voltage vector the other half. The results of this work showed that the proposed method for the four-switch converter can be used either to control electrical machines, or rectifiers and inverters. Moreover, the authors have compared the result of the two vector-based model predictive control with the one voltage vector-based model predictive control. The results showed that the two suitable vectors have lower ripples in currents compared to the one voltage vector-based MPC. However, the switching frequency of the proposed model predictive control is still variable. Moreover, the voltage offset control used to remove the deviation in the DC capacitors is not reported in this work.

Since the two-voltage vector MPC had promising results, the option of using more voltage vectors in each switching period gained the attention of the researchers. As a result, in order to improve the model predictive control for the four-switch converters, an improved predictive control was presented in [95] where two active vectors and one equivalent zero vector were used. In this strategy, the first active vector is the one that has the lowest value in the presented cost function, and the other active vector is the adjacent vector that also has the lower value of the mentioned cost function. This means that the value of the cost function should be estimated first, for all of the possible four voltage vectors in this method. After that, the duration times of the vectors are estimated. Moreover, two low-pass filters are used in this method to obtain the mean value of the capacitor voltages and according to the differences in these two values, the duration times of V_1 and V_3 (two of the active vectors) are increased or decreased in order to suppress the capacitor voltage deviation. Although it has better current THD and better performance compared to other mentioned predictive controls, it is needed to calculate the cost function value for each of the four voltage vectors before calculating the duration times. Moreover, if the duration time of each of the two active vectors obtains a negative value, or if they are larger than the switching period, they are limited to zero and the switching interval, respectively.

Another MPC method for driving a permanent magnet synchronous machine (PMSM) with a four-switch converter is presented in [96]. In this paper, four sector divisions and a virtual duty cycle are defined in the stationary frame. In this method, one active voltage vector and the determined virtual duty cycle are employed according to the position of the sector. In this method, an offset voltage vector is proposed to remove the DC voltage offset.

The result for this method showed that it has better performance compared to the one voltage vector-based MPC. However, a low pass filter is still required to extract the mean value of the capacitor voltages.

In [97], it has been reported that the voltage offset in the DC link for the FSTP feeding a PMSM deteriorate the SVM technique and cause to have restrictions in controlling the machine resulting in a nonacceptable performance. It has been emphasized in this reference that the DC voltage offset should be removed undoubtedly. As a result, the effect of each voltage vector of the FSTP converter on the voltage offset of the capacitors is investigated and compensation on the duty cycle for each switch is suggested. The block diagram of the proposed voltage offset elimination strategy for this case is presented in Fig. 2.29. In this method, a notch filter is used to extract the voltage offset and later on, by using a PI controller, the duty cycle of the SVM is modified. The results for this method showed that the DC voltage offset could be removed with this strategy used to control the electrical machine. However, extracting the DC voltage offset for low rotor speeds could be difficult with this method.

Fig. 2.29. Block diagram of the proposed voltage offset control for four-switch three phase converter driving a PMSM [97].

It should be noted that all of the methods that have been discussed require to employ low pass filters or notch filters to extract the mean value of each DC capacitor voltages since they have a dc and ac components. Extracting the DC values of the capacitors voltages with low pass filters have many disadvantages, especially for the four switch converters that are driving an electrical machine during low speeds. Consequently, when the machine speed is low, extracting mean values is slow and will have errors. Therefore, a conclusion is that there is a need to use a new method to extract the DC voltage values of the capacitors without using any low pass filters.

2.5. Summary

In this chapter, the power electrical generation group of wind turbine systems and wave oscillating water column systems has been briefly presented and analyzed. The variable speed DFIG with partial-scale power converters and the PMSG with full-scale power converters are the most common type of electrical systems that are used in wind turbine and OWC systems. The power converter of these two systems consist of two converters connected

back to back. One converter is controlling the electrical machine while the other converter is regulating the active power, by controlling the DC link voltage, as well as the reactive power.

The general concepts of different control strategies such as linear and predictive controllers for two-level voltage source converters were presented in this chapter. The advantages and disadvantages of each of the different control strategies were discussed. As a result, the MPC seems to be a very interesting strategy and it is nowadays under development in order to remove the drawbacks.

Later on, fault-tolerant methods and reconfiguration of the electrical circuits for the two-level voltage source converter were presented in this chapter. As it was mentioned, the Four-switch three phase converters are one of the more cost-effective and interesting solutions when a switch fault takes place in a converter. Finally, different control strategies were presented for the four-switch three phase converters and the features of each of them were discussed.

Since DFIG and PMSG with back to back converters are playing an important role in the renewable power generation systems, these topologies have been selected in this thesis and an improved MPC, which has a fixed switching frequency, will be presented for these two systems in the following chapter. Moreover, the four-switch three phase converter has been chosen to be the fault-tolerant configuration of the system when there is a switch fault in the system. As a result, a novel predictive control will be presented in this thesis for the four-switch three phase system.

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Chapter 3

A Robust Model Predictive Direct Control based on Four Voltage Vectors for The Power Converters of DFIG and SPMSG

3.1. Introduction

In this chapter, a new and robust Model Predictive Control (MPC) for the power converters of the DFIG and the SPMSG will be presented. In this strategy four voltage vectors are applied in each switching period. The proposed method has the features of fixed switching frequency, low current THD and fast dynamic response. The DFIG is very vulnerable to the grid voltage unbalance since the stator windings of the machine are directly connected to the grid. As a result, the proposed model predictive control should consider these voltage disturbances for the rotor side converter of the DFIG to reduce the bad effects in the machine and to be able to control the induction generator perfectly. However, in the SPMSG, the machine is isolated from the grid by the full rated power converters. Therefore, it is not required to consider the unbalanced grid voltages for the control of the machine side converter of the SPMSG.

This chapter is divided into three parts as follows:

- 1- Proposed predictive direct power control for the rotor side converter (RSC) of the doubly fed induction generator (DFIG)
- 2- Proposed model predictive direct current control for the machine side converter (MSC) of a surface permanent magnet synchronous generator (SPMSG)
- 3- Proposed model predictive direct power control for the grid side converter (GSC) of both the DFIG and the SPMSG.

In the first part, section 3.2 of this chapter, the proposed predictive direct power control for the rotor side converter of a DFIG is presented. The calculation of stator active and reactive power slopes for the DFIG is presented both for ideal and unbalanced grid voltage. Later, the proposed voltage vector sequences and the calculation of the duration times are presented again for ideal and disturbed grid voltage. The negative duration time of the active vectors are analyzed and a new algorithm to select the correct voltage vector sequences for the RSC of a DFIG is proposed and demonstrated in this part. Finally, a different control strategy to overcome the problems during grid voltage unbalance is presented for the RSC.

In the second part of this chapter, section 3.3, the proposed model predictive current control for the machine side converter of a SPMSG is explained. The current slopes in synchronous frame and the voltage vector sequences for the MSC are shown in this part.

Finally, the proposed strategy is explained with the help of a block diagram in order to control the stator current in synchronous speed.

In the last part, section 3.4, the model predictive control for the GSC is presented, where the GSC active and reactive power slopes in stationary frame for balanced grid voltage are expressed. Later on, the duration time of the proposed active vectors during one complete cycle, when the GSC is absorbing or injecting power, is investigated and a new algorithm to recognize the correct voltage sector required by the GSC, is proposed.

3.2. Proposed model predictive direct power control for the rotor side converter of doubly fed induction generator based on four voltage vectors during balanced and unbalanced grid voltage

3.2.1. Introduction

The DFIGs are extremely vulnerable to grid disturbances due to the direct connection of the generator to the grid. As a result, if the DFIG controller does not take into account voltage disturbances, stator and rotor currents will be distorted, and active and reactive powers as well as the electromagnetic torque will be subjected to considerable disturbances. However, the DFIG should tolerate these types of grid voltage situations and should demonstrate an acceptable performance according to power grid codes [1]-[2].

The most widely used control strategy for DFIGs is the field oriented control (FOC) method because it is easy to implement and very reliable. However, classic FOC methods for DFIG cannot handle grid disturbances. Therefore, many modifications have been proposed for this technique to improve wind turbine dynamic performance. The first improvement was introduced in [3]-[4], where two proportional integral (PI) regulators in the positive and negative synchronous frames were used. In order to avoid the use of two reference frames, a resonant controller was added to the PI (R+PI) in [5]-[6]. This technique has been also extended for the case of a simultaneously existing negative sequence and fifth and seventh harmonic voltages, where PI and dual frequency resonant (PI+DFR) controllers were applied [7]-[9]. These improvements are completely dependent on machine parameters, PI controller and Resonance controllers, sequences and harmonic voltages. Therefore, any inaccurate estimation of these values leads to the estimation of an incorrect rotor current reference.

Another well-known method that is easy to implement for controlling the rotor side converter of DFIG is either the direct power or direct torque control (DPC/DTC) strategy. When a voltage imbalance occurs on the stator winding, smooth stator active power and electromagnetic torque [9]-[10] or balanced and sinusoidal stator current [11]-[12] have been achieved by conventional DPC/DTC techniques. Although the structure of this technique is very simple and the extraction of the negative sequence voltage is eliminated, total harmonic distortion (THD) of the current is high due to variable switching frequency. To solve the non-constant switching frequency problem, an SVM-DPC strategy with constant switching frequency has been introduced for unbalanced grid voltage conditions [13]. However, a notch filter, a PI and a resonant controller have been added to the system.

In order to avoid the mentioned problems in FOC and DPC strategies for DFIG, non-linear control methods, such as the sliding mode (SM) control and predictive controls, which have a constant switching frequency and fast dynamic response, have been introduced by researchers in recent years. Initially, predictive direct power control (PDPC) and predictive direct torque control (PDTC) of DFIG were introduced in [14]-[15], where two active vectors and one zero vector were applied during a switching period according to lookup tables and stator active and reactive power values. The first vector of the voltage sequences was selected according to the rotor flux location and the sign of hysteresis comparators. The PDPC vector selection strategy has been simplified in [16]-[17]. However, mentioned PDPC strategies have not taken into account the negative sequence voltage, and performance during voltage disturbance was not addressed. Moreover, the estimated duration times of vectors when they show unexpected values were not studied in depth in these earlier papers. For example in [14], it is suggested to use only the first active vector in a period if the estimated duration time of active vectors became negative. In [18]-[19], an improved PDPC has been presented for unbalanced voltage. In this strategy, only one voltage vector was selected during a switching period. In each period, the active and reactive power slopes for the eight possible vectors are estimated according to two equations. The vector resulting in a lower value of the cost function will be selected. Although this strategy eliminates the duration time estimation and shows better performance compared to conventional DPC, the current THD is high and the switching frequency is still variable.

In order to solve the aforementioned drawbacks related to PDPC during voltage disturbance, this thesis presents an improved PDPC for a DFIG rotor side converter (RSC) where four voltage vectors are applied during each switching period. This strategy has low current THD and small active power and electromagnetic torque ripple. The proposed strategy not only considers the negative value of estimated duration times, but also uses these signs to select the appropriate voltage vector sequences. In addition, three control targets could be easily achieved during unbalanced voltage conditions.

At the beginning of this section, the stator active and reactive power slopes for DFIG are explained for the balanced and unbalanced grid voltages. Later, the voltage vector sequences are presented and the estimation of duration time for each voltage vector of the sequence will be described. Then, the impact of selecting appropriate and inappropriate sequence on duration times are investigated for balanced and unbalanced grid voltage. After that, the proposed model predictive power control will be presented.

3.2.2. Estimation of Stator Active and Reactive Power Slopes

The space vectors of stator and rotor voltage and stator and rotor flux in the stationary frame under ideal and non-ideal stator voltage can be expressed as [20]

$$\vec{V}_s = R_s \vec{I}_s + \frac{d\vec{\varphi}_s}{dt} \quad (3.1)$$

$$\vec{V}_r = R_r \vec{I}_r + \frac{d\vec{\varphi}_r}{dt} - j\omega_m \vec{\varphi}_r \quad (3.2)$$

$$\vec{\varphi}_s = L_s \vec{I}_s + L_m \vec{I}_r \quad (3.3)$$

$$\vec{\varphi}_r = L_m \vec{I}_s + L_r \vec{I}_r \quad (3.4)$$

Where $V_s, V_r, I_s, I_r, \varphi_s, \varphi_r$ are the stator and rotor voltages, stator and rotor currents and stator and rotor fluxes, respectively. The R_s, R_r are the stator and rotor resistances, respectively. The L_s, L_r and L_m are the stator and rotor self inductance and mutual inductance, respectively. The ω_m is rotor angular frequency. According to (3.3) and (3.4), the stator current and rotor flux can be written as

$$\vec{I}_s = \lambda \{L_r \vec{\varphi}_s - L_m \vec{\varphi}_r\} \quad (3.5)$$

$$\vec{\varphi}_r = \frac{1}{L_m} \left\{ L_r \vec{\varphi}_s - \frac{1}{\lambda} \vec{I}_s \right\} \quad (3.6)$$

Where $\lambda = 1/(L_r L_s - L_m^2)$. Stator active and reactive powers can be obtained by

$$P_s = 1.5 \operatorname{Re} \{ \vec{V}_s \cdot \vec{I}_s^\wedge \} \quad (3.7)$$

$$Q_s = 1.5 \operatorname{Im} \{ \vec{V}_s \cdot \vec{I}_s^\wedge \} \quad (3.8)$$

Where P_s and Q_s are the stator active and reactive powers, respectively. The superscript $^\wedge$ denotes conjugated value. According to (3.7) and (3.8), stator active and reactive power derivatives can be described as

$$\frac{dP_s}{dt} = 1.5 \operatorname{Re} \left\{ \frac{d\vec{V}_s}{dt} \cdot \vec{I}_s^\wedge + \frac{d\vec{I}_s^\wedge}{dt} \cdot \vec{V}_s \right\} \quad (3.9)$$

$$\frac{dQ_s}{dt} = 1.5 \operatorname{Im} \left\{ \frac{d\vec{V}_s}{dt} \cdot \vec{I}_s^\wedge + \frac{d\vec{I}_s^\wedge}{dt} \cdot \vec{V}_s \right\} \quad (3.10)$$

As can be observed, stator active and reactive power derivatives are related to stator current and stator voltage derivatives in the stationary frame. According to (3.1), (3.2) and (3.5) and neglecting stator and rotor resistance, the stator current derivative can be shown as

$$\frac{d\vec{I}_s^\wedge}{dt} = \lambda [L_r \vec{V}_s^\wedge - L_m [\vec{V}_r^\wedge - j\omega_m \vec{\varphi}_r^\wedge]] \quad (3.11)$$

By substituting (3.6) into (3.11), the stator current derivative can be rewritten as

$$\frac{d\vec{I}_s^\wedge}{dt} = \lambda [L_r \vec{V}_s^\wedge - L_m \vec{V}_r^\wedge + j\omega_m L_r \vec{\varphi}_s^\wedge] - j\omega_m \vec{I}_s^\wedge \quad (3.12)$$

The stator current slope in the stationary frame can be estimated by (3.12). The stator voltage under unbalanced voltage can be described as follows

$$\vec{V}_s = \vec{V}_s^+ + \vec{V}_s^- = |v_s^+| e^{j(\omega_s t + \alpha_+)} + |v_s^-| e^{j(-\omega_s t + \alpha_-)} \quad (3.13)$$

Where ω_s is synchronous angular frequency and superscripts + and – denote positive and negative sequences, respectively. The α_+ and α_- are the angle of positive and negative sequences of stator voltage, respectively. The stator voltage derivative under unbalanced grid voltage can be illustrated as

$$\frac{d\vec{V}_s}{dt} = j\omega_s(\vec{V}_s^+ - \vec{V}_s^-) = j\omega_s\vec{V}_s - 2j\omega_s\vec{V}_s^- \quad (3.14)$$

The precise stator voltage slope value can be estimated by (3.14). This equation shows that stator voltage variations depend on the positive and negative sequence voltage. Consequently, according to (3.12) and (3.14), stator active and reactive power slopes yield as follows

$$\begin{aligned} \frac{dP_s}{dt} = & -1.5\omega_s \text{Im}[\vec{V}_s \cdot \vec{I}_s^\wedge] + 3\omega_s \text{Im}[\vec{V}_s^- \cdot \vec{I}_s^\wedge] \\ & + 1.5\lambda [L_r |\vec{V}_s|^2 - L_m \text{Re}[\vec{V}_s \cdot \vec{V}_r^\wedge] - \omega_m L_r \text{Im}[\vec{V}_s \cdot \vec{\varphi}_s^\wedge]] \\ & + 1.5\omega_m \text{Im}[\vec{V}_s \cdot \vec{I}_s^\wedge] \end{aligned} \quad (3.15)$$

$$\begin{aligned} \frac{dQ_s}{dt} = & 1.5\omega_s \text{Re}[\vec{V}_s \cdot \vec{I}_s^\wedge] - 3\omega_s \text{Re}[\vec{V}_s^- \cdot \vec{I}_s^\wedge] \\ & + 1.5\lambda [-L_m (\text{Im}[\vec{V}_s \cdot \vec{V}_r^\wedge] + \omega_m L_r \text{Re}[\vec{V}_s \cdot \vec{\varphi}_s^\wedge])] \\ & - 1.5\omega_m \text{Re}[\vec{V}_s \cdot \vec{I}_s^\wedge] \end{aligned} \quad (3.16)$$

As a result, active and reactive power slopes could be calculated by (3.15) and (3.16) during unbalanced stator voltage. According to grid codes, the allowed negative sequence stator voltage value is much smaller than the stator voltage (V_s). As a consequence, the second term of equation (3.14) could be neglected in order to avoid extracting the negative sequence voltage. Thus, by neglecting the second term of (3.14), the power slopes could be obtained by

$$\begin{aligned} \frac{dP_s}{dt} = & -(\omega_s - \omega_m)Q_s + 1.5\lambda L_r |V_s|^2 - 1.5\lambda L_m \text{Re}\{\vec{V}_s \cdot \vec{V}_r^\wedge\} \\ & - 1.5\lambda L_r \omega_m \text{Im}\{\vec{V}_s \cdot \vec{\varphi}_s^\wedge\} \end{aligned} \quad (3.17)$$

$$\frac{dQ_s}{dt} = (\omega_s - \omega_m)P_s - 1.5\lambda L_m \text{Im}\{\vec{V}_s \cdot \vec{V}_r^\wedge\} + 1.5\lambda L_r \omega_m \text{Re}\{\vec{V}_s \cdot \vec{\varphi}_s^\wedge\} \quad (3.18)$$

The stator active and reactive power derivatives under unbalanced grid voltage can be calculated according to (3.17)-(3.18) with a very small error.

To give an illustration of estimating power slopes using (3.17) and (3.18), active and reactive power slopes of the eight possible rotor voltage vectors under ideal and unbalanced grid voltage are established. Fig. 3.1 shows active and reactive power slopes under balanced voltage situations as a function of rotor flux angle in the rotor frame. In this illustration, it has been assumed that the stator is injecting 1.5 MW active power and zero Mvar reactive power into the grid, and the rotor speed is set to 0.8 PU

Fig. 3.1. Stator active and reactive power derivatives under balanced grid voltage situation as a function of rotor flux location in rotor frame for the eight possible rotor voltage vectors at 0.8 PU rotor speed.

Active and reactive power slopes under unbalanced stator voltage are shown in Fig. 3.2. In this figure, it has been assumed that a 10% voltage dip in one of the stator phases has happened while the DFIG is injecting 1.5 MW and zero Mvar. As shown in this figure, it is obvious that during imbalanced voltage, stator active and reactive power slopes for the eight rotor voltage vectors are distorted compared to the balanced voltage conditions due to the existence of a negative sequence stator voltage and flux.

Fig. 3.2. Stator active and reactive power derivatives under unbalanced grid voltage as a function of rotor flux location in rotor frame for the eight possible rotor voltage vectors at 0.8 PU rotor speed.

Stator flux under unbalanced grid voltage could be expressed as

$$\vec{\varphi}_s = \vec{\varphi}_s^+ + \vec{\varphi}_s^- = |\varphi_s^+|e^{j(\omega_s t + \alpha_+ - \frac{\pi}{2})} + |\varphi_s^-|e^{j(-\omega_s t + \alpha_- + \frac{\pi}{2})} \quad (3.19)$$

Rotor voltage in the stationary frame for converter vectors can be shown by

$$\vec{V}_r = \frac{2}{3} V_{dc} e^{j(\omega_m t + (m-1)\frac{\pi}{3})} \quad (3.20)$$

Where V_{dc} is the DC link voltage and the m is the number of the two-level converter active voltage vectors (between 1 and 6 for active vectors). The rotor voltage value for the zero vectors (V_0 and V_7) are zero. In order to investigate the component of stator active and reactive power slopes under unbalanced grid voltage, (3.13), (3.19) and (3.20) are substituted into (3.17) and (3.18). As a result, the power slopes can be shown by

$$\begin{aligned} \frac{dP_s}{dt} = & k_1 + k_2 \cos((\omega_s - \omega_m)t + \beta_+) + k_3 \cos((\omega_s + \omega_m)t + \beta_-) \\ & + k_4 \cos(2\omega_s t + \delta) \end{aligned} \quad (3.21)$$

$$\begin{aligned} \frac{dQ_s}{dt} = & k_5 + k_2 \sin((\omega_s - \omega_m)t + \beta_+) - k_3 \sin((\omega_s + \omega_m)t + \beta_-) \\ & + k_6 \sin(2\omega_s t + \delta) \end{aligned} \quad (3.22)$$

Where the values of k_1 to k_6 and the β_+ , β_- and δ are expressed in the appendix A. These two equations indicate that when the grid voltage is unbalanced, the power slope component frequency is related to $\omega_s - \omega_m$ (slip speed), $\omega_s + \omega_m$ and twice the grid frequency. In contrast, when grid voltage is balanced, only the first two terms of (3.21) and (3.22) exist. Consequently, as can be seen in Fig. 3.1, when grid voltage is balanced, the slopes have a DC value and sine or cosine waves with slip speed frequency. While under unbalanced grid voltage, the other components are added to the power slopes. As shown in Fig. 3.2, the power slopes contain the mentioned distorted components during unbalanced grid voltage.

In order to evaluate the impact of neglecting the second term of equation (3.14) on power slopes, stator active and reactive power slopes for the rotor voltage vector V_1 are estimated by equations (3.15), (3.16) and (3.17), (3.18) during a 20% voltage dip in one of the stator phases. In addition, stator active and reactive power slopes for this rotor voltage vector are estimated according to the equations proposed in [20-21]. The results are shown in Fig. 3.3. The injected powers and rotor speed are the same as before. By comparing the result of estimated power slopes based on (3.17) and (3.18) with the precise power slopes (equations (3.15), (3.16)), it is obvious that neglecting the second term of (3.14) has a very small impact on estimations. Consequently, in order to prevent the extraction problems involved with negative sequence voltage, the equations of (3.17) and (3.18) are used for calculating the power slopes of selected rotor voltage vectors. On the other hand, according to Fig. 3.3, the estimated power slopes based on [20-21] show considerable error when compared with precise values. Nevertheless, the control strategy proposed in [20-21] could work during unbalanced voltage.

Fig. 3.3. Comparison between estimation of stator active and reactive power slopes based on equations (3.15, 3.16) and (3.17, 3.18) and Ref. [20-21] for rotor voltage V_1 under unbalanced grid voltage.

3.2.3. Duration Times of Vector Sequences

Once the approximate location of a required rotor voltage for a period has been detected, the two nearest voltage vectors are appropriate active vectors to construct the required rotor voltage. Therefore, if the required rotor voltage sector is predicted, voltage vector sequences could be easily determined according to Table 3.1. In this table, V_a , V_b , V_c and V_d are voltage vector sequence in a period. The range and border of each sector are shown in Fig. 3.4. In this figure, the D and Q axes show the rotor frame. For example, assuming that the required rotor voltage is located in sector 1, the required rotor voltage is between 0° and 60° , the four voltage vector sequences will be V_0 , V_1 , V_2 , V_7 , respectively.

Table 3.1. Proposed Voltage vector sequences

Vector sequence	Sector of required rotor voltage (N)					
	1	2	3	4	5	6
V_a	V_0	V_7	V_0	V_7	V_0	V_7
V_b	V_1	V_2	V_3	V_4	V_5	V_6
V_c	V_2	V_3	V_4	V_5	V_6	V_1
V_d	V_7	V_0	V_7	V_0	V_7	V_0

Fig. 3.4. Rotor and stator flux and voltage vectors at steady state in rotor frame (DQ): (a) sub-synchronous speed and (b) hyper-synchronous speed.

When voltage vector sequences are applied in a period, stator active and reactive powers at $k+1$ instant can be shown as

$$P_s(k+1) = P_s(k) + S_{pa}t_a + S_{pb}t_b + S_{pc}t_c + S_{pd}t_d \quad (3.23)$$

$$Q_s(k+1) = Q_s(k) + S_{qa}t_a + S_{qb}t_b + S_{qc}t_c + S_{qd}t_d \quad (3.24)$$

Where S_{pa} , S_{pb} , S_{pc} , S_{pd} and S_{qa} , S_{qb} , S_{qc} , S_{qd} are the active and reactive power slopes of the corresponding voltage vectors, respectively. The duration times of the first, the second, the third and the fourth active vectors are called t_a , t_b , t_c and t_d respectively.

Zero vectors have the same power slopes, which means

$$\begin{aligned} S_{pa} &= S_{pd} \\ S_{qa} &= S_{qd} \end{aligned} \quad (3.25)$$

Errors between stator powers and their references at $k+1$ instant can be shown

$$P_{s-error}(k+1) = P_s(k+1) - P_s^*(k) \quad (3.26)$$

$$Q_{s-error}(k+1) = Q_s(k+1) - Q_s^*(k) \quad (3.27)$$

Where superscript $*$ denotes the reference value. In order to reduce errors, the following standard cost function can be used in this situation

$$F(k+1) = P_{s-error}^2(k+1) + Q_{s-error}^2(k+1) \quad (3.28)$$

In order to minimize this cost function, the derivatives of this function with respect to duration times of active vectors are set to zero. By solving these equations, the duration times

of the vectors could be yielded. The duration times of the vectors in a period can be obtained by [21]

$$t_b = \frac{(P_s(k) - P_s^*(k)) \cdot (S_{qc} - S_{qa}) + (Q_s(k) - Q_s^*(k)) \cdot (S_{pa} - S_{pc})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} + \frac{T_{sw}(S_{qc} \cdot S_{pa} - S_{qa} \cdot S_{pc})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} \quad (3.29)$$

$$t_c = \frac{(P_s(k) - P_s^*(k)) \cdot (S_{qa} - S_{qb}) + (Q_s(k) - Q_s^*(k)) \cdot (S_{pb} - S_{pa})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} + \frac{T_{sw}(S_{qa} \cdot S_{pb} - S_{qb} \cdot S_{pa})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} \quad (3.30)$$

$$t_a = t_d = \frac{T_{sw} - t_b - t_c}{2} \quad (3.31)$$

Where the T_{sw} is the switching period time. The duration times of vectors in a period could be obtained according to the power slope of the selected vectors and power errors. If the selected active vectors are appropriate vectors, the duration of the active vectors should have positive values.

3.2.4. The Impact of Selecting Appropriate and Inappropriate Vector Sequences on Duration Times

The rotor voltage in the synchronous frame could be expressed by

$$\vec{V}_r = R_r \vec{I}_r + \frac{d\vec{\varphi}_r}{dt} + j(\omega_s - \omega_m) \vec{\varphi}_r \quad (3.32)$$

By neglecting rotor resistance, the required rotor voltage in the k^{th} switching period could be shown as

$$\vec{V}_r^*(k) = \frac{\vec{\varphi}_r^*(k) - \vec{\varphi}_r(k)}{T_{sw}} + j(\omega_s - \omega_m) \vec{\varphi}_r^*(k) \quad (3.33)$$

3.2.4.1. In steady state

In steady state, the rotor flux derivative is much smaller than the second term of Eq. (3.33). In this situation, the first term of (3.33) could be neglected when there is a significant difference between rotor speed and synchronous speed. Therefore, rotor voltage could be shown by

$$\vec{V}_r = j(\omega_s - \omega_m) \vec{\varphi}_r \quad (3.34)$$

According to (3.34), the required rotor voltage is almost 90 degrees ahead of rotor flux under sub-synchronous speeds and almost 90 degrees behind rotor flux during hyper-synchronous speeds. For example, as can be seen in Fig. 3.4, if the rotor flux is located between 270° and 330° , the required rotor voltage under sub-synchronous and hyper-synchronous speed will be located in sector 1 and sector 4, respectively.

In order to see the influence of selecting appropriate and inappropriate active vectors on duration times (t_b and t_c) during steady state, V_1 and V_2 are selected as the two active vectors existing in one period, and the duration times of these two vectors are estimated by (3.29) and (3.30) for one complete rotor flux rotation. Figs. 3.5(a) and 3.5(b) show the duration time of V_1 and V_2 according to the rotor flux position in steady state and under sub-synchronous and hyper-synchronous speed, respectively.

According to Fig. 3.5(a), only when rotor flux is located between $[270^\circ, 330^\circ]$, the duration times of V_1 and V_2 are simultaneously positive, which means that the required rotor voltage is located in sector 1. During other rotor flux positions, one of the active vector durations, or even both, are negative. When rotor flux is located at $[150, 210]$ and $[210, 270]$, according to (3.34) the required rotor voltage is located in sectors 5 and 6, respectively. However, according to Fig. 3.5(a), if V_1 and V_2 are selected to be applied in these locations, the duration time of V_2 (t_c) will be negative while the duration time of V_1 (t_b) will be positive. On the other hand, If V_1 and V_2 are selected to be applied when rotor flux is located at $[-30, 30]$ and $[30, 90]$ (according to (3.34), the required rotor voltage is located in sectors 2 and 3), the duration time of V_2 (t_c) and V_1 (t_b) will be positive and negative, respectively. Finally, when the required rotor voltage is located in sector 4, rotor flux is located at $[90, 150]$, the duration time of both V_1 and V_2 (t_b, t_c) will be negative if these vectors are selected.

Fig. 3.5. Duration times of V_1 and V_2 as two selected active vectors in a period versus rotor flux position at steady state: (a) during sub-synchronous speed and (b) during hyper-synchronous speed.

The effect of selecting appropriate and inappropriate active vectors on duration times during hyper-synchronous speed is the same as during sub-synchronous speed. For example, as can be seen in Fig. 3.5(b), when the required rotor voltage is located in sectors 5 and 6 (rotor flux is located at $[-30, 30]$ and $[30, 90]$, respectively) duration times of V_2 (t_c) and V_1 (t_b) will be negative and positive, respectively. At this speeds, duration times of V_1 and V_2

are both positive when the rotor flux is located between $[90^\circ, 150^\circ]$ (rotor voltage is located in sector 1).

Duration times of V_2 and V_3 as the two selected active vectors in a period for one complete rotor flux rotation under sub-synchronous and hyper-synchronous speeds are shown in Figs. 3.6(a) and 3.6(b), respectively. It can be concluded from Figs. 3.5 and 3.6 that when both selected active vectors $[V_N, V_{N+1}]$ have positive duration times, the selected vectors are appropriate vectors and the required rotor voltage is located in sector N . However, if duration times of first active vector (V_N) and second active vector (V_{N+1}) have negative and positive values, respectively, this means that the required rotor voltage is located in sector $N+1$ or $N+2$, and the appropriate active vectors are $[V_{N+1}, V_{N+2}]$ or $[V_{N+2}, V_{N+3}]$. In contrast, if the duration times of first active vector (V_N) and second active vector (V_{N+1}) have positive and negative values, respectively, this means that the required rotor voltage is located in sector $N-1$ or $N-2$, and the appropriate active vectors are $[V_{N-1}, V_N]$ or $[V_{N-2}, V_{N-1}]$. Finally, if both selected active vectors $[V_N, V_{N+1}]$ have negative duration times, the required rotor voltage is located in sector $N-3$ and the appropriate active vectors are $[V_{N-3}, V_{N-2}]$. Consequently, the appropriate rotor voltage sector in a period versus the sign of the duration time of two selected active vectors $[V_N, V_{N+1}]$ is summarized in Table 3.2.

Fig. 3.6. Duration times of V_2 and V_3 as two selected active vectors in a period versus rotor flux position at steady state: (a) during sub-synchronous speed and (b) during hyper-synchronous speed.

Table 3.2. Appropriate rotor voltage sector in a period versus the sign of duration time of two selected active vectors $[V_N, V_{N+1}]$

	$t_{v_{N+1}} > 0$	$t_{v_{N+1}} < 0$
$t_{v_N} > 0$	N	$N - 1$ or $N - 2$
$t_{v_N} < 0$	$N + 1$ or $N + 2$	$N \pm 3$

3.2.4.2. In transient state or near synchronous speed

In transient state or near synchronous speed, rotor flux variation in a period is considerable and should not be neglected in (3.33). In this situation, the error between stator powers and their references plays an important role in determining the required rotor voltage.

In order to investigate the effect of selecting appropriate and inappropriate voltage vectors on duration times during the transient state, the stator active power reference has been increased by 0.1 PU at the beginning of a control period. Therefore, stator active power in this switching period should be increased. The duration times of three different voltage vector sequences in this situation, i.e., $[V_1, V_2]$, $[V_2, V_3]$ and $[V_3, V_4]$, versus rotor flux position under sub-synchronous speed are shown in Fig. 3.7. According to Fig. 3.7(a), duration times of V_1 and V_2 are simultaneously positive only when rotor flux is between A and B (highlighted section). Therefore, while rotor flux is between A and B, the required rotor voltage for this active power increment is located in sector 1. According to Fig. 3.7, when rotor flux is located at $[B, C]$ and $[C, D]$, rotor voltage is in sectors 2 and 3, respectively. The duration time of V_1 is negative and the duration time of V_2 is positive when rotor flux is between $[B, D]$. On the other hand, duration times of V_3 and V_4 are positive and negative, respectively, when rotor flux is between $[A, C]$. As can be seen in Fig. 3.7, the required rotor voltage is located in sector 1 or 2 in this case. Consequently, the effect of selecting appropriate or inappropriate vector sequences on duration times during the transient state is the same as in the steady state.

Fig. 3.7. Duration times of two selected active vectors in a period versus rotor flux position under sub-synchronous speed at transient state: (a) duration times of $[V_1, V_2]$, (b) duration times of $[V_2, V_3]$ and (c) duration times of $[V_3, V_4]$.

It should be noted that during the transient state, the total duration time of active vectors could be longer than the switching period time (T_{sw}). This indicates that in order to reach the reference value in one period, a larger rotor voltage is needed, which could not be generated by the RSC. Hence, the maximum required rotor voltage value that can be generated by the RSC is the one that should be applied in a period. Thus, if calculated duration times are both positive and the total duration time is longer than the period time ($t_b + t_c > T_{sw}$), duration times should be corrected as follows

$$t'_b = \frac{t_b}{t_b + t_c} T_{sw}, t'_c = \frac{t_c}{t_b + t_c} T_{sw}, t'_d = t'_a = 0 \quad (3.35)$$

Where t'_a, t'_b, t'_c and t'_d are the corrected values of the vector sequences duration times in this situation.

3.2.4.3. Under an unbalanced voltage situation

The stator flux will contain negative sequence when an unbalanced voltage occurs in stator windings. Therefore, rotor flux and voltage might have positive and negative sequence according to required power references. In this subsection, positive and negative duration time values for three different vector sequences are investigated under a 20% voltage dip in one of the phases. In this situation, power references were set to inject constant active and reactive powers, and the speed was set to 0.8 PU. The results are shown in Fig. 3.8. According to this figure, when rotor flux is located at [270, 330], in some areas, duration times of V_1 and V_2 are both positive, which indicates that the appropriate vectors are [V_1, V_2]. These areas are highlighted in Fig. 3.8(a). Therefore, when rotor flux is located at [A, B], [C, D], [E, 330] and [-30, F], the appropriate vectors are V_1 and V_2 . However, when rotor flux is located at [270, 330], in other areas, the appropriate vectors are [V_2, V_3] and [V_6, V_1]. For example, when rotor flux is at [B, C], duration times of V_2 and V_1 are negative and positive, respectively. According to Fig. 3.8(c), duration times of V_6 and V_1 in this area are positive, which indicates that the required rotor voltage is located in sector 6. Moreover, according to Fig. 3.8(a) when rotor flux is at [D, E], duration times of V_2 and V_1 are positive and negative, respectively. According to Fig. 3.8(b), the appropriate vectors in this area are [V_2, V_3].

When comparing Fig. 3.5(a) to Fig. 3.8(a), it is clear that the rotor flux position is not a good solution for selecting vectors during unbalanced grid voltage. Although under balanced grid voltage, the rotor flux position can be used to determine the appropriate vectors, it cannot be used during unbalanced grid voltage. Nevertheless, according to Fig. 3.8, under an unbalanced voltage situation, the effect of selecting inappropriate vectors on duration times is the same as in a normal voltage situation. Subsequently, Table 3.2 is correct and can also be used under unbalanced grid voltage to determine the appropriate voltage vector sequences.

Fig. 3.8. Duration times of two selected active vectors in a period versus rotor flux position under unbalanced grid voltage and sub-synchronous speed: (a) duration times of $[V_1, V_2]$, (b) duration times of $[V_2, V_3]$ and (c) duration times of $[V_6, V_1]$.

3.2.5. Proposed Vector Sequence Selections and Overall Block Diagram of the proposed predictive control

In the previous section, the impact of selecting appropriate and inappropriate voltage vector sequences on duration times of selected active vectors has been investigated. According to the duration time signs of selected voltage vector sequences, voltage vectors could be selected. The block diagram of the proposed PDPC for DFIG is shown in Fig. 3.9. At the beginning of each switching period, it is assumed that the required rotor voltage sector (N) is the same as in the previous switching period. Therefore, in a first step, the voltage vector sequences are chosen using this assumption.

Stator active and reactive power slopes are estimated for these vector sequences according to (3.17) and (3.18). Then, according to the error of active and reactive powers as compared to their references, and estimated power slopes, duration times of rotor voltage vectors are calculated. If duration times of both selected active vectors are positive, it means that the selected active vectors are appropriate vectors, and they will be applied by the RSC according to the calculated duration time of each vector. However, the duration time of one of the selected active vectors or even both of the selected active vectors could be negative. In this situation, three cases would be possible:

Fig. 3.9. Proposed predictive direct power control (PDPC) Block diagram.

3.2.5.1. $t_b < 0$ and $t_c > 0$

These duration time signs indicate that the appropriate voltage vector sequences are located in sector $N+1$ or $N+2$. Therefore, in this case, it will first be assumed that the appropriate voltage vector sequences are located in sector $N+1$, and the duration times of these new vectors are estimated. If the duration times of both new active vectors are positive, the assumption is correct and the required rotor voltage is located in sector $N+1$. However, if the required rotor voltage is located in sector $N+2$, the duration time of the new first active vector (V_{N+1}) will be negative while the duration of the new second active vector (V_{N+2}) will be positive. In this new situation, the voltage vector sequences of the next sector ($N+2$) will be selected, and the duration times of these new vectors will be calculated. Finally, these new vectors are applied.

3.2.5.2. $t_b > 0$ and $t_c < 0$

In this case, the voltage vector sequences are located in sector $N-1$ or $N-2$. Therefore, it will be assumed that the appropriate voltage vector sequences are located in sector $N-1$. The duration times of new vectors (V_{N-1}, V_N) are estimated, and if the estimated durations of both are positive, it indicates that the required rotor voltage is located in sector $N-1$. However, the required rotor voltage may be located in sector $N-2$. Therefore, in this case, the duration time of the new first active vector (V_{N-1}) will be positive, while the duration of the new second active vector (V_N) will be negative. In this situation, the voltage vector sequences of the sector $N-2$ will be selected.

3.2.5.3. $t_b < 0$ and $t_c < 0$

In this situation, the appropriate voltage vectors are definitely located in sector N-3. Therefore, the voltage vector sequences are selected according to sector N-3, and the duration time of these new vectors, which will definitely be positive, will be estimated according to the power slope of these new vectors.

It should be noted that, as mentioned before, if both the selected active vectors are positive and the sum of both durations is longer than the switching period, duration times will be corrected according to (3.35).

3.2.6. Applying Different Control Targets Under Unbalanced Grid Voltage

Under unbalanced grid voltage, three different control targets could be achieved by adjusting stator active and reactive power references, which are:

3.2.6.1. Constant active and reactive powers (i.e., removing powers pulsations)

In order to fulfill this target stator active and reactive power references should be set to only constant DC values. Therefore, the power references are

$$P_s^* = P_{const} \quad (3.36)$$

$$Q_s^* = Q_{const} \quad (3.37)$$

where P_{const} and Q_{const} are DC values, which P_{const} is obtained from the MPP curve.

3.2.6.2. Constant electromagnetic torque and stator reactive power (i.e., removing electromagnetic torque and stator reactive power pulsations)

To reach to this target, stator power references should be set as follows [12]

$$P_s^* = P_{const} - \left(\frac{T_{em}\omega_s}{p} - P_s \right) \quad (3.38)$$

$$Q_s^* = Q_{const} \quad (3.39)$$

Where T_{em} , P_s and p are measured electromagnetic torque and measured stator active power and the pole pairs, respectively. The stator reactive power in this situation should be set to a constant DC value. Electromagnetic torque could be calculated by

$$T_{em} = \frac{3}{2}p \operatorname{Im}\{\vec{I}_s \cdot \vec{\varphi}_s^*\} = \frac{3}{2}p(\varphi_{s\alpha}I_{s\beta} - \varphi_{s\beta}I_{s\alpha}) \quad (3.40)$$

Where α and β are the stationary axes.

3.2.6.3. Sinusoidal and balanced stator currents

Sinusoidal and balanced stator currents can be achieved under unbalanced grid voltage operation when stator active and reactive power references are as follows [20]:

$$P_s^* = P_{const} + \frac{3}{2} (V_{s\alpha}^- I_{s\alpha}^+ + V_{s\beta}^- I_{s\beta}^+) \quad (3.41)$$

$$Q_s^* = Q_{const} + \frac{3}{2} (V_{s\beta}^- I_{s\alpha}^+ - V_{s\alpha}^- I_{s\beta}^+) \quad (3.42)$$

In this control target, the negative sequence voltage in the stationary frame is required to obtain stator active and reactive power references.

It should be noted that the goal of control target one is to remove active and reactive power pulsation and inject smooth active and reactive power into the grid, while the goal of target three is to have balanced heating on three phase stator windings and inject balanced and sinusoidal current into the grid. Mechanical stress on the turbine system and gearbox damage will be reduced when control target two is applied. Among these three different control targets, it is more important to apply target two due to the high cost of repairing the mechanical part of a wind turbine.

Fig. 3.10. Block diagram of generating power references for the three control targets under unbalanced grid voltage.

Fig. 3.10 shows the block diagram of generating the power references for these three control targets under unbalanced grid voltage. Control target two could be fulfilled by estimating electromagnetic torque and knowing the stator active and reactive power DC values. In order to fulfill control target three, negative sequence stator voltage and positive sequence stator current are used to generate active and reactive power references. Therefore, the extraction of negative sequence stator voltage and positive sequence stator current is required for target three.

3.3. Proposed Model Predictive Direct Current control based on four voltage vectors for the machine side converter of SPMSG

3.3.1. Introduction

Nowadays, the surface permanent magnet synchronous machine (SPMSM) connected to the back to back converter are employed in many applications such as wind turbine and oscillating water column systems due to the excellent features of this machine. The common control strategy for this machine is field oriented control (FOC), or vector control (VC), which has low torque ripples and has fixed switching frequency. This control contains inner and outer loops. The outer loop usually controls the torque and flux while the inner loop controls the stator currents. However, the control provides a slow response and any perturbation in the system or in the grid could deteriorate the performance of the controller [22], [23]. Another control strategy, which is also very common in electrical drivers, is direct torque control (DTC). The performance of the conventional DTC, based on a switching table, is related to the sample time of the microcontroller [24]. As a result, the ripple of the torque and other variables is high if a high sampling time is not applied, in spite of the fact that DTC method is fast and has a robust structure compared to the FOC control [25].

In recent years, different model based predictive controls (MPCs) for voltage source converters have been introduced and applied to renewable energy sources and electrical drives, especially for PMSM drivers. In this control, the model of the system is considered, and according to a cost function and the control objectives, an optimal voltage vector is chosen to drive the PMSM [26]. With this technique, the phase locked loop (PLL) and the space vector modulation code are unnecessary. In classical MPC, only one voltage vector is applied in every switching interval [27]; as a result, a high ripple appears in the control variables. The switching frequency of this method, similarly to DTC, is variable. Additionally, the results look alike to those obtained with the DTC method [28].

Many alternative strategies have been introduced to overcome these disadvantages such as the duty cycle optimization based on two vector MPC (TWMPC) [29]-[30] where a nonzero, and a zero vector are applied in each interval. Also, a reduced ripple based on TWMPC [31]-[32], a multi-vector predictive control [33]-[34] and predictive power control based on four vectors [35] were also presented and comprise the short horizon MPC. A long horizon direct torque MPC for PMSM was suggested in [36]-[37] where an intelligent calculation method is needed. Although this method features promising performances, the implementation in a real system is difficult nowadays.

As aforementioned, an effective and easy implementation of MPC featuring fixed switching frequency and low variable ripple is highly demanded by the industry, especially in electrical generation.

This section presents a model predictive direct current control (MPDCC) based on four voltage vectors for SPMS Generators. The proposed strategy features fixed switching frequency, fast dynamic response and low ripple in the control variables, what make it very suitable to make OWC power plants and wind turbines using SMPMG track the maximum efficiency point in every situation.

The section 3.3 is organized as follows. In Section 3.3.2 and section 3.3.3 the current slopes and the selection of voltage vector sequences of the proposed model predictive direct current control (MPDCC) for the SPMSG is presented and the calculation of the optimal duration time of vectors are explained. In the section 3.3.4, the overall proposed model predictive control for the SPMSG is given. It should be noted that the OWC emulator is used to check the performance of the proposed MPDCC. Hence, the OWC based power plant emulator model is briefly described in section 3.3.5 of this thesis.

3.3.2. Current slopes and vector sequences

The SPMSG dynamic equations in rotor rotating reference frame, d-q frame, can be presented as [38]

$$u_{sd} = R_s i_{sd} + \frac{d}{dt} \lambda_{sd} - \omega_m \lambda_{sq} \quad (3.43)$$

$$u_{sq} = R_s i_{sq} + \frac{d}{dt} \lambda_{sq} + \omega_m \lambda_{sd} \quad (3.44)$$

where u_s , i_s , λ_s , ω_m and R_s are stator voltage vector, stator current vector, stator flux, electrical rotor speed and stator resistance, respectively. The stator flux of SPMSG in the d-q frame is

$$\lambda_{sd} = L_s i_{sd} + \lambda_f \quad (3.45)$$

$$\lambda_{sq} = L_s i_{sq} \quad (3.46)$$

where λ_f and L_s represent permanent magnet flux and stator inductance, respectively. The torque equation can be expressed as

$$T_e = \frac{3}{2} p (\lambda_{sd} i_{sq} - \lambda_{sq} i_{sd}) \quad (3.47)$$

Where T_e and p are the electromagnetic torque and pole pairs. The derivative of current in rotor reference frame can be obtained according to (3.43)-(3.46) as

$$\frac{d}{dt} i_{sd} = \frac{1}{L_s} (-R_s i_{sd} + \omega_m L_s i_{sq} + u_{sd}) \quad (3.48)$$

$$\frac{d}{dt} i_{sq} = \frac{1}{L_s} (-R_s i_{sq} - \omega_m L_s i_{sd} - \omega_m \lambda_f + u_{sq}) \quad (3.49)$$

As a result, the stator current derivatives of SPMSG in the rotor frame can be obtained from the machine parameters, rotor speed, permanent magnet flux, currents and the MSC voltage in the d-q frame.

The MSC can produce eight possible voltage vectors, expressed in the stationary reference frame α - β , according to the state of the six switches. These vectors are shown in Fig. 3.11. In this figure, six sectors are defined by these vectors, and the stator and machine fluxes are shown as an example. Two of these eight vectors are the zero vectors (V_0, V_7) and the rest are the active vectors.

Fig. 3.11. Eight possible voltage vectors of the MSC.

The current slopes in the d-q axis for these eight possible voltage vectors (V_0 - V_7) are derived as

$$S_{i_{sd}}|_{u_{sdi}} = \frac{1}{L_s} (-R_s i_{sd} + \omega_r L_s i_{sq} + u_{sdi}) \quad (3.50)$$

$$S_{i_{sq}}|_{u_{sqi}} = \frac{1}{L_s} (-R_s i_{sq} - \omega_r L_s i_{sd} - \omega_r \lambda_f + u_{sqi}) \quad (3.51)$$

where i and S_i denote the index of the MSC voltage vector and current slope, respectively. The current slopes for these eight possible voltage vectors for an 8.7 kW SPMSG along one cycle are shown in Fig. 3.12. The current slopes for the zero vectors are constant while those for the active vectors are changing sinusoidally.

It is assumed that in the proposed control, four voltage vectors are applied in each switching interval, in which two of them are the zero vectors, and the two remaining vectors are the active vectors. The voltage vector sequence in every period is specified according the corresponding stator voltage sector. These vector sequences for each stator voltage sector are defined in Table 3.3. The variation of stator current in the d-q frame at the end of a switching interval can be expressed as

$$i_{sd}(k+1) = i_{sd}(k) + S_{da}t_a + S_{db}t_b + S_{dc}t_c + S_{dd}t_d \quad (3.52)$$

$$i_{sq}(k+1) = i_{sq}(k) + S_{qa}t_a + S_{qb}t_b + S_{qc}t_c + S_{qd}t_d \quad (3.53)$$

where S_{da} , S_{db} , S_{dc} and S_{dd} are the current slopes in d axis for the respective voltage vector sequence. S_{qa} , S_{qb} , S_{qc} and S_{qd} represent the current slopes in q axis for the selected vector sequence. The t_a , t_b , t_c and t_d are the duration times of the corresponding voltage vectors.

Fig. 3.12. The current slopes for the voltage vectors along one complete cycle: (a) in d axis; (b) in q axis.

Table 3.3. Stator voltage vector sequences

Vector sequence	Stator voltage Sector (N)					
	1	2	3	4	5	6
Va	V0	V7	V0	V7	V0	V7
Vb	V1	V2	V3	V4	V5	V6
Vc	V2	V3	V4	V5	V6	V1
Vd	V7	V0	V7	V0	V7	V0

In order to minimize the error between the current and the reference value in the rotor frame, the well-known cost function is used as follows [26]

$$F(k+1) = (i_{sq}(k+1) - i_{sq}^*(k))^2 + (i_{sd}(k+1) - i_{sd}^*(k))^2 \quad (3.54)$$

The duration time of each vector in an interval can be obtained by minimizing the cost function, which yields to

$$t_b = \frac{(i_{sd}(k) - i_{sd}^*(k)) \cdot (S_{qc} - S_{qa}) + (i_{sq}(k) - i_{sq}^*(k)) \cdot (S_{da} - S_{dc})}{S_{qa}(S_{db} - S_{dc}) + S_{qb}(S_{dc} - S_{da}) + S_{qc}(S_{da} - S_{db})} + \frac{T_{sw}(S_{qc} \cdot S_{da} - S_{qa} \cdot S_{dc})}{S_{qa}(S_{db} - S_{dc}) + S_{qb}(S_{dc} - S_{da}) + S_{qc}(S_{da} - S_{db})} \quad (3.55)$$

$$t_c = \frac{(i_{sd}(k) - i_{sd}^*(k)) \cdot (S_{qa} - S_{qb}) + (i_{sq}(k) - i_{sq}^*(k)) \cdot (S_{db} - S_{da})}{S_{qa}(S_{db} - S_{dc}) + S_{qb}(S_{dc} - S_{da}) + S_{qc}(S_{da} - S_{db})} + \frac{T_{sw}(S_{qa} \cdot S_{db} - S_{qb} \cdot S_{da})}{S_{qa}(S_{db} - S_{dc}) + S_{qb}(S_{dc} - S_{da}) + S_{qc}(S_{da} - S_{db})} \quad (3.56)$$

$$t_a = t_d = \frac{T_{sw} - t_b - t_c}{2} \quad (3.57)$$

where T_{sw} is the switching interval. If the estimated durations of active vectors are positive, and the total value ($t_b + t_c$) is bigger than the switching interval, the duration times should be set to

$$t_b' = \frac{t_b}{t_b + t_c} T_{sw} \quad (3.58)$$

$$t_c' = \frac{t_c}{t_b + t_c} T_{sw} \quad (3.59)$$

$$t_a' = t_d' = 0 \quad (3.60)$$

This case can occur when the stator voltage reference is bigger than the voltage that can be produce by the MSC. When the voltage vector sequence and their durations are calculated, they will be applied like in the space vector modulation (SVM).

3.3.3. Sector identification algorithm

In predictive controls, selecting the appropriate voltage sequence is the vital part of the strategy. In many research papers, it has been reported that certain durations of selected active vectors get negative, and the easiest solution was to remove that voltage vector from the switching interval [14]-[15],[21]. This solution makes the switching frequency variable and increases the ripple. To overcome this problem, many predictive controls suggest using two vectors, which consist of one zero vector (V_0, V_7) and one active vector (V_1 - V_6). The active vector of these predictive controls is achieved from the conventional one voltage based predictive controls [39].

These negative values of the duration times occur when the sector that proposes the voltage vector sequence is not selected correctly. In this part, the negative values of the vector durations are investigated, and a stator voltage sector identification algorithm is proposed for the SPMSG.

In Fig. 3.13, the duration times of the active vectors for the six possible voltage vector sequences during one complete rotation cycle according to the required stator voltage sector are shown. This figure presents the duration times of active vectors during steady state and transients. It can be seen in Fig. 3.13 that in each sector, the duration times of the corresponding active vectors are both positive, while in other situations, at least one of the active vectors is negative. For instance, the durations of the V_1 and V_2 are positive only when they are applied in sector 1, or durations of V_2 and V_3 are positive only when they are applied in sector 2. However, in other sectors, if these voltage vectors are chosen, the duration time of one of the vectors is negative at least. For example, as shown in Fig. 3.13(a), the durations of V_1 gets negative in sector 2 if V_1 and V_2 are chosen as the two selected active vectors.

When the duration times of the selected voltage vector sequence get negative, some information can be obtained from this situation that can help to identify the correct sector. When observing the six cases shown in Fig. 3.13, these negative duration times can be classified as three types:

3.3.3.1. $t_b < 0$ and $t_c > 0$

When the first active vector duration presents a negative value and the second active vector has a positive value, this indicates that the correct stator voltage sector is one or two sectors ahead. To give an example, assume that the stator voltage is located in sector 2, the V_2 and V_3 are the correct vectors. In this sector, if the V_6 and V_1 are selected, the duration of the V_6 presents a negative value while the V_1 obtains a positive value according to Fig. 3.13(f). If duration times of V_1 and V_2 are calculated when the required stator voltage is in sector 2, the duration time of the first active vector will have negative value, and the second vector will be positive, as can be observed from Fig. 3.13(a).

3.3.3.2. $t_b < 0$ and $t_c < 0$

The correct stator voltage sector is located exactly three sectors ahead from the initial assumed sector, when both durations become negative. For instance, if the duration times of V_1 and V_2 are calculated when the required stator voltage is located in sector 4, both of the times will have negative values like it is shown in Fig. 3.13(a). This reveals that the correct vectors are V_4 and V_5 and the required stator voltage is located in sector 4.

Consequently, in this subsection, the negative duration times of vectors were analyzed. It can be concluded that if the durations of the selected active vectors get negative, the sector should be changed according to the respective case type.

Fig. 3.13. Active vector durations in one complete rotation cycle according to the required stator voltage sector: (a) t_{v1} , t_{v2} ; (b) t_{v2} , t_{v3} ; (c) t_{v3} , t_{v4} ; (d) t_{v4} , t_{v5} ; (e) t_{v5} , t_{v6} ; (f) t_{v6} , t_{v1} .

3.3.4. Overall proposed MPDCC of SPMPG

The block diagram of the proposed predictive control is illustrated in Fig. 3.14. In each control interval, the voltage vector sequence is chosen according to the previous prediction of the stator voltage sector in the previous interval. The current slopes in the d-q frame for this vector sequence are evaluated from the speed, the permanent magnet flux of the machine and the currents. After that, the durations of the vector sequence are obtained according to these slopes and the measured currents and their references in the d-q frame. If the times of the active vectors are positive, this vector sequence is applied by the electronic converter.

However, one or two of the active vector durations could become negative and, in that case, the sector location should be modified according to the proposed sector identification algorithm. Hence, new voltage vector sequence will be selected, and the duration times of these vectors will be determined. If at the first place, both vector duration times are negative, the sector will be changed to three forward sectors, making the duration times for the new vectors become positive, and so the new vectors will be applied by the converter. However, if only one of the active vector times is negative at the first place, the sector changes to one sector forward or one sector backward depending which duration time got negative. Nonetheless, these new vector durations might get negative again. If they get negative again, this makes obvious that the correct sector is located in two sectors ahead or two sectors behind the first sector assumption. In other words, in the worst case the durations can be calculated three times and in the last time, the duration times of the active vectors will be positive indubitably. This last case could happen when a big sudden change in the position of the stator voltage is required. Finally, after the correct vector sequences are estimated, the vectors will be applied by the electronic converter. It should be noted that, if the total duration times of the active vectors are bigger than the switching interval, the duration times of the vectors are modified according to the (3.58)-(3.60).

Fig. 3.14. Block diagram of proposed MPDCC for the SPMPG.

3.3.5. OWC based power plant emulator

One of the most well-known wave energy converters is the one based on an oscillating water column (OWC) normally installed onshore, preferably on rocky shores. This type of converter comprises a chamber where the wave is captured, and an air turbine, normally a Wells turbine, that features constant direction of rotation for both direction of the air flow through its blades [40-42]. The schematic of this OWC power plant was shown in Fig. 2.11, in chapter 2.

Waves enter into the OWC chamber forcing the air upwards and downwards, making the Wells turbine to spin and converting wave energy into mechanical energy. Then, the turbine shaft drives an electric generator that converts the mechanical energy into electrical energy.

The strategy chosen in this thesis is to make the OWC power plant work at the maximum efficiency point through the control of the rotary speed of the Wells turbine and the electric generator. The highly variable shaft speed is controlled using the proposed model predictive control developed in this thesis.

The OWC device behavior has been reproduced using the following mathematical models:

3.3.5.1. Irregular wave model

The irregular wave model represents the state of the free surface of the sea by its energy spectrum, $S(\omega)$. This spectrum is obtained from oceanographic buoys measurements and it provides information about the energy content for the frequencies that compose the real wave. The $S(\omega)$ can also be used to reproduce the sea state keeping the same spectral characteristics (significant height, peak period, etc.) as the original real waves.

There are many spectra that describe the state of the sea where the OWC is located: Pierson-Moskowitz [43]-[46], JONSWAP [40-42], [47], TMA [43], [48], and Bretschneider [49].

Using the wave spectrum, $S(\omega)$ provided by these models, it is possible to generate an irregular wave represented by the addition of n regular waves, according to the expression

$$\eta(x, t) = \sum_{i=1}^n \zeta_i \cdot \cos(\omega_i t - k_i x + \varepsilon_i) \quad (3.61)$$

Where the amplitude of each harmonic is

$$\zeta_i = \sqrt{2 \cdot S(\omega_i) \cdot \Delta\omega_i} \quad (3.62)$$

Every wave presents a randomly calculated phase and all the phases have the same probability of occurrence, expressed by a constant probability density in the interval $(-\pi, \pi)$.

The Pierson-Moskowitz [47] model was chosen in this thesis, which can be shown by

$$S(\omega) = \frac{5}{16} \cdot H_s^2 \cdot \omega_p^4 \cdot \omega^{-5} \exp \left[-\frac{5}{4} \cdot \left(\frac{\omega}{\omega_p} \right)^{-4} \right] \quad \zeta_i = \sqrt{2 \cdot S(\omega_i) \cdot \Delta \omega_i} \quad (3.63)$$

where $\omega = 2\pi/T$ is the angular frequency of the wave, H_s is the significant wave height, T_s is the significant wave period and $\omega_p = 2\pi/T_p$ is the angular spectral peak frequency. Fig. 3.15 shows the real wave profile generated for the tests carried out in this thesis.

Fig. 3.15. The real wave profile used is made up of 2400 samples $H_s=1$ and $T_s=10$.

3.3.5.2. Chamber model

Detailed modeling of the chamber in an OWC device, results in complex mathematical models [50]-[53]. The OWC device equations programed in the emulator represent a simplified chamber model [54]. As a consequence, the hydrodynamics inside the chamber is not fully described, but what it is more relevant from the electrical point of view, the electrical power delivered to the grid does not lose the multi frequency spectrum of the wave resource.

The air speed in the duct of the turbine can be described by the following expression (see Table 3.4 and Table 3.5 in Appendix for the nomenclature, and Table 3.5 for the turbine parameters):

$$V_x = \left(\frac{A_c}{A_d} \right) \cdot \frac{\partial h_c}{\partial t} \quad (3.64)$$

The torque coefficient, C_t , depends on the characteristic curve of each particular turbine and is a function of the air flow through the turbine, Φ :

$$\Phi = \frac{V_x}{r_t \cdot \omega_t} \quad (3.65)$$

Table 3.4. Chamber model

V_x	Air speed in the duct (m/s)
h_c	Height of the free water surface inside the chamber (m)
A_c	Chamber cross section (m ²)
A_d	Duct cross section (m ²)

Table 3.5. Turbine model

T_t	Shaft torque (N.m)	r_t	Turbine radius (m).
ϕ	Flow coefficient (non-dimensional)	V_x	Air speed in duct (m/s).
C_t	Torque coefficient (non-dimensional)	ω_t	Angular velocity of rotor system (rad/s)
K	Constant of the turbine (kg/m)		

Table 3.6. Wells Turbine parameters

	Maximum C_t before stall	for $\phi = 0.311$
Maximum	Air density	$\rho = 1.2 \text{ kg/m}^2$
efficiency flow	Height of the blades	$b=0.045 \text{ m}$
$\phi_{max\text{ef}} = 0.17$	Blade chord length	$l=0.090 \text{ m}$
	Number of turbine blades	$n=8$

3.3.5.3. Wells turbine torque emulation

The Wells turbine torque can be obtained as

$$T_t = C_t \cdot K \cdot \text{Radius} \cdot [V_x^2 + (\text{Radius} \cdot \omega_t)^2] \quad (3.66)$$

Fig. 3.16 shows, in solid line, the characteristic curve ($C_t - \Phi$) of the turbine model that has been represented in the prototype [55]-[56] and the efficiency curve in dashed line.

Fig. 3.16. C_t and efficiency as a function of the flow coefficient.

For every wave and V_x , the calculated Wells turbine torque is reproduced using a separately excited DC motor, Fig. 3.17. However, for every calculated motor torque, the resulting rotary speed depends also on the opposite torque. At this point is where the proposed MPDCC method system performance becomes very important since it controls the opposite torque created by the SPMSG to make the power plant to work at the maximum efficiency point in every moment.

Fig. 3.17. Control block of the OWC device emulator.

The torque control of the SPMSG uses the Φ for the maximum efficiency obtained from Fig. 3.16. For this given Φ , the corresponding rotary speed, ω_t is calculated for every wave and V_x in every control cycle and used as reference speed in the proposed MPDCC control system, Fig. 3.14.

3.3.5.4. Maximum efficiency speed reference for the MPDCC

The speed reference used by the MPDCC block, Fig. 3.14, is the Wells turbine maximum efficiency speed. This speed is calculated according to the maximum efficiency flow coefficient represented in the turbine efficiency curve [56]–[57], Fig. 3.18.

Fig. 3.18. Wells turbine efficiency as a function of the flow coefficient.

In every program cycle, the microcontroller calculates the airspeed from the wave height inside the chamber, (3.64). Then, using the $\Phi_{max\ ef}$ obtained from Fig. 3.18, and the turbine radius, the maximum efficiency speed is calculated by using (3.65). The block diagram of Fig. 3.19 shows this procedure.

Fig. 3.19. Calculation of the maximum efficiency speed.

3.4. Proposed Model Predictive Direct Power Control for the grid side converter (GSC) of DFIG and SPMSG

3.4.1. Introduction

As mentioned in the previous sections, model predictive control (MPC) started to be used more than thirty years ago and it is considered one of the most important advances in control processes. In the case of grid-connected converters for electric power generation, the MPC is a very promising approach, where a discrete model is used to predict the behavior of the system within the prediction horizon [40]. For MPC with reference tracking, the reduced computation time available in the microcontroller program cycle, normally limits the number of implementable prediction horizons to one [58]-[59]. The MPC strategy obtains the converter output or switching states that minimizes the reference tracking error using a cost function. This way, the control action can be applied by the converter without requiring a modulation stage, although sometimes it can lead to carrying out several iterations until a correct solution is obtained. Compared with the direct power control (DPC) strategy for the grid-connected converters, the vector selected from the MPC strategy is more accurate and effective in reducing power errors, while a similar or even better dynamic response is obtained [60].

Recently, model predictive DPC (MPDPC or MPPC) has been proposed as a powerful alternative to conventional DPC, because it is more accurate and effective in the vector selection [61]-[62]. By selecting the best voltage vector minimizing a cost function which consist of power errors, a better steady-state performance, in terms of power ripples, and current harmonics, can be obtained [63].

For example, an improved MPPC with duty cycle control for grid-connected power converters, which is able to further reduce the power ripples in conventional MPC, is proposed in [64]. With this approach, only one prediction is required to select the best voltage vector, and its duration is determined based on the principle of error minimization of both active and reactive power. The MPC has been lately used in a number of different topics such as improving the quality of the power injected to the grid, implementing current source

control in the converters, implementing decentralized control of VSC, etc. In [65], current ripple is reduced using active power filters whose modulator is based on a cost function ratio for different output vectors and a single loop controller. Also, reference [66] proposes an alternative MPC implementation intended to reduce the current ripple of a two-level three-phase grid-connected rectifier, by integrating virtual vectors in the conventional MPC structure and synthesizing them from a discrete space vector modulation approach. Current control for a three-phase single-stage grid-tied PV converter, which controls the active and reactive power exchanged with the grid, is proposed in [67]. It carries out a two-step prediction of the future value of the grid current. Also, an MPC for dual-mode inverters in distributed generation applications is proposed in [68]. In this case, the converter operates as a current source for the ac grid and a voltage source for the load, respectively.

In [21], a three-voltage vector based MPC control is proposed for the grid-connected converters. In this method, 12 sectors are defined for the grid voltage and the voltage vector sequences are selected according to the grid voltage sector. In this method, it has been suggested to remove a voltage vector when the duration time of that vector has a negative value.

This chapter proposes an improved model predictive control based on four voltage vectors to be implemented on the GSC of DFIG and SPMSG systems, that can absorb and inject active power. In this method, without using the grid voltage sector, the correct vectors and their duration times can be selected and applied by the converter.

3.4.2. Grid Side Converter Active and Reactive Power Slopes

The schematic of the grid side converter (GSC) of DFIG or SPMSG is shown in Fig. 3.20. As can be seen in this figure, the GSC is connected to the grid by an inductance and a resistance filter. Since the influence of the grid disturbance for the GSC is not severe like in the induction machine, it has been assumed that the grid voltage is balanced in this section.

Fig. 3.20. Schematic of the grid side converter (GSC) of the DFIG and SPMSG.

According to this Fig. 3.20, the GSC voltage vector in the stationary frame, $\alpha\beta$ frame, can be expressed as

$$\vec{v}_{GSC} = \vec{u}_g + L_{fg} \frac{d\vec{i}_g}{dt} + R_{fg} \vec{i}_g \quad (3.67)$$

Where v_{GSC} , u_g , i_g , L_{fg} and R_{fg} are the grid side converter voltage, the grid voltage, the grid current, the grid side inductance, and the resistance, respectively. The grid side converter active and reactive power in the stationary frame can be estimated as

$$P_g = 1.5(u_{g\alpha}i_{g\alpha} + u_{g\beta}i_{g\beta}) \quad (3.68)$$

$$Q_g = 1.5(u_{g\beta}i_{g\alpha} - u_{g\alpha}i_{g\beta}) \quad (3.69)$$

Where P_g , Q_g are the GSC active and reactive power, respectively. The subscript α and β are the stationary axes. The active and reactive powers of the GSC can be estimated by (3.68) and (3.69) at every switching period. According to (3.68) and (3.69), the derivatives of the GSC active and reactive power can be calculated by

$$\frac{dP_g}{dt} = 1.5\left(\frac{du_{g\alpha}}{dt}i_{g\alpha} + \frac{di_{g\alpha}}{dt}u_{g\alpha} + \frac{du_{g\beta}}{dt}i_{g\beta} + \frac{di_{g\beta}}{dt}u_{g\beta}\right) \quad (3.70)$$

$$\frac{dQ_g}{dt} = 1.5\left(\frac{du_{g\beta}}{dt}i_{g\alpha} + \frac{di_{g\alpha}}{dt}u_{g\beta} - \frac{du_{g\alpha}}{dt}i_{g\beta} - \frac{di_{g\beta}}{dt}u_{g\alpha}\right) \quad (3.71)$$

Consequently, these derivatives can be estimated in the stationary frame according to the derivatives of the grid voltage and currents in the stationary frame. The grid voltage in a balanced system can be represented as

$$u_{g\alpha} = |\vec{u}_g| \sin(\omega_s t) \quad (3.72)$$

$$u_{g\beta} = -|\vec{u}_g| \cos(\omega_s t) \quad (3.73)$$

Where the ω_s is the grid angular frequency. According to the (3.72)-(3.73), the grid voltage slope in the stationary frame yields

$$\frac{du_{g\alpha}}{dt} = \omega_s |\vec{u}_g| \cos(\omega_s t) = -\omega_s u_{g\beta} \quad (3.74)$$

$$\frac{du_{g\beta}}{dt} = \omega_s |\vec{u}_g| \sin(\omega_s t) = \omega_s u_{g\alpha} \quad (3.75)$$

As a result, the grid voltage slopes can be represented by the grid voltage and the synchronous speed according to the (3.74)-(3.75). According to (3.67), the derivative of the grid current can be estimated as

$$\frac{di_{g\alpha}}{dt} = \frac{(v_{GSC\alpha} - u_{g\alpha} - R_{fg}i_{g\alpha})}{L_{fg}} \quad (3.76)$$

$$\frac{di_{g\beta}}{dt} = \frac{(v_{GSC\beta} - u_{g\beta} - R_{fg}i_{g\beta})}{L_{fg}} \quad (3.77)$$

The GSC active and reactive power slopes can be represented by substituting (3.74)-(3.77) into (3.70) and (3.71). As a result, the GSC active and reactive power derivatives can be derived as

$$\frac{dP_g}{dt} = \frac{1.5}{L_{fg}} [(u_{g\alpha}v_{GSC\alpha} + u_{g\beta}v_{GSC\beta}) - (u_{g\alpha}^2 + u_{g\beta}^2)] - \frac{R_s}{L_{fg}}P_g - \omega_s Q_g \quad (3.78)$$

$$\frac{dQ_g}{dt} = \frac{1.5}{L_{fg}} (u_{g\alpha}v_{GSC\beta} - u_{g\beta}v_{GSC\alpha}) - \frac{R_s}{L_{fg}}Q_g + \omega_s P_g \quad (3.79)$$

In conclusion, the active and reactive power derivatives of the GSC for all the possible voltage vectors of the GSC can be estimated according to (3.78) and (3.79), respectively.

3.4.3. Proposed vector sequences and duration times of the vector sequences

In order to select the appropriate voltage vector sequences in each switching period, the location of the required GSC voltage should be detected. When the location is determined, the two nearest voltage vectors are the appropriate active vectors to construct the required GSC voltage. Table 3.7 proposes the voltage vector sequences for the GSC converter according to the sector of the predicted GSC voltage. In this table, V_a , V_b , V_c , and V_d are voltage vector sequences in a period. The ranges of the sectors are shown in Fig. 3.21. For example, when the GSC voltage reference is located in sector 1, the appropriate voltage vectors in order to build the reference voltage are the V_0 , V_1 , V_2 , and V_7 . Fig. 3.21 shows an example position of the GSC and the grid voltage as well as the eight possible voltage vectors in the stationary frame. As can be seen in this figure, when the GSC is injecting active power to the grid, with the converter operating as an inverter, the reference voltage of the GSC is ahead of the grid voltage. On the other hand, when the GSC is absorbing active power, rectifier operation of the converter, the reference voltage of the GSC is behind the grid voltage.

Table 3.7. Proposed Voltage vector sequences for GSC

Vector sequence	Sector of required GSC voltage (N)					
	1	2	3	4	5	6
V_a	V_0	V_7	V_0	V_7	V_0	V_7
V_b	V_1	V_2	V_3	V_4	V_5	V_6
V_c	V_2	V_3	V_4	V_5	V_6	V_1
V_d	V_7	V_0	V_7	V_0	V_7	V_0

Fig. 3.21. GSC and grid voltage vectors and the eight possible voltage vectors in the stationary frame ($\alpha\beta$): (a) the GSC is working as an inverter (b) the GSC is working as a rectifier.

In this strategy, four voltage vectors are employed in each switching period according to table 3.7, where two of the vectors are the active vectors and the others are the zero vectors. Hence, these voltage vectors will increase or decrease the active and reactive power values in each period according to their power slopes and their duration times. As a result, the GSC active and reactive powers at $k+1$ instant can be shown as

$$P_g(k+1) = P_g(k) + S_{pa}t_a + S_{pb}t_b + S_{pc}t_c + S_{pd}t_d \quad (3.80)$$

$$Q_g(k+1) = Q_g(k) + S_{qa}t_a + S_{qb}t_b + S_{qc}t_c + S_{qd}t_d \quad (3.81)$$

Where S_{pa} , S_{pb} , S_{pc} , S_{pd} and S_{qa} , S_{qb} , S_{qc} , S_{qd} are the active and reactive power slopes of the corresponding voltage vectors, respectively. The duration times of the first, the second, the third and the fourth active vectors are called t_a , t_b , t_c and t_d respectively.

As mentioned before, the zero vectors will have the same power slopes, which means

$$\begin{aligned} S_{pa} &= S_{pd} \\ S_{qa} &= S_{qd} \end{aligned} \quad (3.82)$$

Errors between GSC powers and their references at $k+1$ instant can be shown as

$$P_{g-error}(k+1) = P_g(k+1) - P_g^*(k) \quad (3.83)$$

$$Q_{g-error}(k+1) = Q_g(k+1) - Q_g^*(k) \quad (3.84)$$

Where superscript * denotes the reference value. The following cost function can be used in this situation

$$G(k+1) = P_{g-error}^2(k+1) + Q_{g-error}^2(k+1) \quad (3.85)$$

In order to minimize this cost function, the derivatives of this function with respect to the duration times of the active vectors are set to zero. By solving these equations, the duration times of the vectors in a period could be obtained by

$$t_b = \frac{(P_g(k) - P_g^*(k)) \cdot (S_{qc} - S_{qa}) + (Q_g(k) - Q_g^*(k)) \cdot (S_{pa} - S_{pc})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} + \frac{T_{sw}(S_{qc} \cdot S_{pa} - S_{qa} \cdot S_{pc})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} \quad (3.86)$$

$$t_c = \frac{(P_g(k) - P_g^*(k)) \cdot (S_{qa} - S_{qb}) + (Q_g(k) - Q_g^*(k)) \cdot (S_{pb} - S_{pa})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} + \frac{T_{sw}(S_{qa} \cdot S_{pb} - S_{qb} \cdot S_{pa})}{S_{qa}(S_{pb} - S_{pc}) + S_{qb}(S_{pc} - S_{pa}) + S_{qc}(S_{pa} - S_{pb})} \quad (3.87)$$

$$t_a = t_d = \frac{T_{sw} - t_b - t_c}{2} \quad (3.88)$$

Where the T_{sw} is the switching period time. The duration times of the vectors in a period can be obtained according to the grid active and reactive power slopes of the selected vectors and the power errors. If the selected active vectors are appropriate vectors, the duration of those active vectors should have positive values. As mentioned in the previous section, if the total duration times of the active vectors is bigger than the switching frequency, the vector duration times should be modified according to

$$t_b' = \frac{t_b}{t_b + t_c} T_{sw} \quad (3.89)$$

$$t_c' = \frac{t_c}{t_b + t_c} T_{sw} \quad (3.90)$$

$$t_d' = t_a' = 0 \quad (3.91)$$

Where t_a' , t_b' , t_c' and t_d' are the modified values of the duration times.

3.4.4. Sector identification algorithm

Selecting the correct voltage vector sequences in each period is very important for the GSC. Many investigations have been carried out in order to select the appropriate voltage vectors according to the grid voltage sector for the GSC. In [60],[69]-[70], it has been suggested to use only one voltage vector for DC/AC converters in every switching period in order to avoid negative duration times. In [21], three voltage vectors are selected in each switching period according to the grid voltage position for a DC/AC converter. However, it has been shown that sometimes the duration times could get negative values and those negative duration times were removed from the voltage sequences. Moreover, twelve sectors are defined in this strategy and a big table was presented to select the vector sequences. The proposed method in [21] was proposed only for the rectifier operation of the two-level voltage source converter while the inverter operation was not considered. However, the GSC of a

DFIG is a bidirectional converter and, depending on the generator speed, the GSC can absorb or inject active power.

In this part, the negative duration times of the active vectors of the voltage sequences for the GSC are investigated, and the proposed method to select the correct voltage vector sequence without needing the grid voltage sector is presented. As mentioned before, the negative values of the duration times happen when the GSC sector that proposes the voltage vector sequence is not correctly selected.

Fig. 3.22 shows the duration times of the active vectors for the proposed six voltage vector sequences of a GSC during one complete rotation cycle of the grid voltage, when the GSC is absorbing active power from the grid (rectifier operation). In this figure, the duration times of the active vectors are presented according to the location of the grid voltage (grid voltage sector). Moreover, this figure presents the duration times applicable in transient and steady-state operation. As can be seen in this figure, the grid voltage sector is not a good solution to select the voltage vector sequences. For example, when the grid voltage is located at the beginning of sector 1, the duration times of V_1 and V_2 are positive and negative, respectively. The location where the duration times of both V_1 and V_2 are positive is highlighted in Fig. 3.22(a). For the other voltage vector sequences, the location where both of the active vectors are positive is highlighted in Fig. 3.22.

Fig. 3.23 presents the duration times of the active vectors according to the grid voltage position for the proposed six voltage vector sequences of the GSC during one complete rotation cycle of the grid voltage, when the GSC is injecting active power to the grid (inverter operation). This figure also shows that the sector of the grid voltage is not suitable for selecting the voltage vector sequences. For example, as can be seen in Fig 3.23(a) when the grid voltage sector is at the end of the sector 1, the duration time of V_1 and V_2 are negative and positive, respectively. According to Fig. 3.23(b), in this location, the appropriate vectors are V_2 and V_3 . The locations where the proposed six voltage vector sequences have positive duration times when the GSC is injecting active power are highlighted in Fig 3.23.

According to the results that have been obtained in Fig. 3.22 and 3.23, when the duration times of the selected voltage vector sequence get negative, some information can be obtained from this situation that can help identifying the correct sector. These negative duration times for the GSC when it is absorbing or injecting active power can be classified as three types:

3.4.4.1. $t_b < 0$ and $t_c > 0$

This case shows that if the first and second active vector durations have negative and positive values, respectively, the correct required GSC voltage sector is one or two sectors ahead. For example, if it is assumed that the GSC voltage is located in sector one and the voltage vector sequence of this sector is selected and the estimated duration times of V_1 and V_2 obtain negative and positive values, the correct GSC voltage sector could be sector 2 or 3, which means that the correct voltage vector sequence could be (V_2, V_3) or (V_3, V_4) , as can be seen in Fig. 3.22 and Fig. 3.23.

Fig. 3.22. Active vectors duration times of the six voltage vector sequences in one complete rotation cycle according to the grid voltage sector when the GSC is operating as rectifier: (a) t_{v1} , t_{v2} ; (b) t_{v2} , t_{v3} ; (c) t_{v3} , t_{v4} ; (d) t_{v4} , t_{v5} ; (e) t_{v5} , t_{v6} ; (f) t_{v6} , t_{v1} .

Fig. 3.23. Active vectors duration times of the six voltage vector sequences in one complete rotation cycle according to the grid voltage sector when the GSC is operating as inverter: (a) t_{v1} , t_{v2} ; (b) t_{v2} , t_{v3} ; (c) t_{v3} , t_{v4} ; (d) t_{v4} , t_{v5} ; (e) t_{v5} , t_{v6} ; (f) t_{v6} , t_{v1} .

3.4.4.2. $t_b > 0$ and $t_c < 0$

If the first and second active vector durations have positive and negative values, respectively, the correct required GSC voltage sector is one or two sectors behind. Hence, a new voltage vector sequence should be selected, and its duration times should be estimated. For example, if it is assumed that the GSC voltage is located in sector one and the voltage vector sequence of this sector is selected and the estimated duration times of V_1 and V_2 obtain positive and negative values, the correct GSC voltage sector could be sector 6 or 5. As can be seen in Fig. 3.22 and Fig. 3.23, the correct voltage vector sequence for this example should be (V_2, V_3) or (V_3, V_4) .

3.4.4.3. $t_b < 0$ and $t_c < 0$

In this case, when both duration times of the active vectors have negative duration times, the correct required GSC voltage sector is located three sectors ahead. For instance, if both duration times of V_1 and V_2 get negative values, the correct GSC sector is sector 4, and the appropriate active vectors are V_4 and V_5 . This example can be seen in Fig. 3.22 and 3.23.

Consequently, the negative duration times of active vectors for the GSC of DFIG and SPMSG have been analyzed. It can be concluded that if the durations of the selected active vectors get negative, the required GSC voltage sector should be changed according to the respective case type. It should be noted that these cases are the same when the GSC is operating both in inverter mode and rectifier mode.

3.4.5. Overall proposed Model Predictive Power Control for the GSC

The overall block diagram of the proposed model predictive power control for the GSC of DFIG and SPMSG systems is shown in Fig. 3.24. As can be seen in this figure, the GSC active and reactive powers are calculated in each switching period. The active power reference of the GSC is estimated according to the PI controller output of the DC link voltage error. The GSC reactive power reference is selected according to the desired value. In this strategy, the voltage vector sequence is chosen in each period according to the previous prediction of the grid side converter voltage sector. The GSC active and reactive power derivatives are calculated for this vector sequence according to the grid angular frequency, the grid voltage in the stationary frame, the GSC estimated active and reactive powers and the filter parameters. The duration times of the vectors are obtained according to these derivatives as well as the measured active and reactive powers, according to equations (3.86)-(3.88). These selected vectors are applied by the GSC, only if the duration times of the active vectors are positive.

However, if one or two of the active vector durations result negative, it means that the selected vector sequence is not correct and the GSC sector should be modified according to the proposed sector identification algorithm. As a result, according to each case, the new voltage vectors will be selected. For example, if both vector duration times are negative, the GSC sector will be changed to the third forward sector and new voltage vectors will be selected according to this new sector, as well as the duration times of these new vectors, which will also be estimated. In this case, both estimated durations will be positive and can be employed with the new vectors by the GSC. If only one of the active vector durations

obtains a negative value in the first place, the GSC sector should change to one sector forward or one sector backward depending on the case, as shown in Fig. 3.24. Nonetheless, these new vector durations might get a negative value again. If they get negative again, this makes obvious that the correct sector is located in two sectors ahead or two sectors behind the first sector assumption.

Finally, when the estimated duration times of the selected active vectors are positive, the vectors will be applied by the electronic converter. If the total duration times of the active vectors is bigger than the switching interval ($t_b + t_c > T_{sw}$), the duration times of the vectors should be modified according to equations (3.89)-(3.91).

Fig. 3.24. Block diagram of the proposed model predictive power control for the GSC of DFIG and SPMSG.

3.5. Summary and Conclusions

In this chapter, the proposed model based predictive control based on four voltage vectors has been presented for the RSC of the DFIG, the MSC of the SPMSG and the GSC of both the DFIG and the SPMSG. The proposed strategy has a fixed switching frequency and fast dynamic response features.

The proposed predictive control requires to estimate the slopes when controlling the variables in order to estimate the duration times of the selected vectors. The estimation of the stator active and reactive power slopes for DFIG has been presented, both for balanced and unbalanced grid voltages. The GSC active and reactive power slopes as well as the current slopes in the synchronous frame for the MSC of SPMSG were also presented in this chapter. The appropriate voltage vector sequences in each period for this strategy are recognized according to the estimated duration times of the selected vectors. A negative value in the estimated duration time of the selected vectors has been considered in this strategy and it has been proven that this information helps when selecting the correct voltage vectors.

In the proposed MPC strategy for the RSC, the estimation of the rotor flux location is not needed and the voltage vector sequences are selected according to a simple algorithm. The proposed model predictive control for the RSC that has been presented in this chapter can easily follow the power references under normal and abnormal voltage conditions, even if the power references contain AC terms. Three different techniques were presented during unbalanced grid voltage without using any notch filter, in order to decrease the effect of the negative sequence on the induction machine. Without any additional controller, PI Controller, or resonant controller, the proposed technique can obtain smooth stator active and reactive powers, or smooth electromagnetic torque, or can inject sinusoidal and balanced currents into the grid, when unbalanced voltage appears in the DFIG stator windings.

The proposed model predictive control is easy to be implemented in a microcontroller and its computing time means not a bigger burden than the amount of time typically spent in classical vector control. The control can effectively predict the appropriate stator voltage sector and the voltage vector sequences while controlling the currents in dq frame, keeping the switching frequency fixed as well. The magnetic field and torque control, carried out by means of the predictive calculation of the d-q components of the stator current, are very fast whereas the ripple in the stator currents would be very low.

In the model predictive control that was presented for the GSC, the correct voltage vector sequences are selected without needing to know the grid voltage sector. The proposed strategy can follow the GSC power references when the GSC is operating as a rectifier and as an inverter. This bidirectional operation of the proposed strategy for the GSC is essential for the DFIG.

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Switch Fault Tolerance capability of the model predictive control for the Power converters of DFIG and SPMSG

4.1. Introduction

The maintenance of offshore wind turbines and wave sea systems is difficult and present high cost due to the limited access and bad weather conditions. Hence, any fault in the electrical generation group might cause to disconnect from the grid the overall generation power unit for several weeks or months, which causes a huge potential loss of power generation [1]. As a result, wind turbines, especially the offshore ones, and wave sea systems, should contribute to extract energy and inject power to the grid even under certain fault conditions, to make the system more cost-effective.

The Fault tolerance topologies for a two-level three phase power converter were presented in chapter 2. As it was mentioned in that chapter, among many different causes of electrical failures, the breakdown of capacitors and semiconductor switches are the more frequent, and have drawn the attention of researchers have proposed solutions trying to improve the reliability of wind generators [2]-[4]. There are different fault-tolerant methods for power electronic converters such as using multilevel converters [19]-[22][5]-[8], redundant parallel [9] and series switches installation [10]-[11], in which the extra switches are placed in parallel and series with the principal switches, and the four-switch three phase topology, whereas the midpoint of the capacitors is connected to the faulty phase or connected to the neutral point of the motor or the transformer [12]-[14].

As mentioned in chapter 2, the four-switch three phase converter (FSTP) is a very interesting solution for the switch fault in the two-level converters. In this topology, the faulty phase will be disconnected from the faulty leg and will be connected to the midpoint of the DC link capacitors in order to continue working as a four-switch converter, providing three phase currents. Therefore, this topology was selected in this thesis as a solution during the switch fault of the two-level converters and the research has been conducted in this line.

Since the converter configuration will be changed after transferring to the post-fault topology, the control of the power converter should also be changed and improved according to the new situation. As a result, the Model Predictive Control (MPC) that was presented in chapter 3 should be changed and adapted to the new power converter topology. The MPC strategy for the four-switch three phase converters (FSTPCs) should be designed carefully to work properly. Consequently, this chapter will present the model predictive control strategies for the FSTP machine side converter (MSC) of SPMSG and for the FSTP grid side converter (GSC). This chapter will present the following subjects:

- 1- Model Predictive Current Control for the machine side converter of SPMSG when fed by a four-switch three phase converter.
- 2- Model Predictive Direct Power Control Strategy for Four-Switch Three Phase Operation of the Grid Side Converter.

This chapter is organized as follows:

- In section 4.2 of this chapter, the stator current slopes and the four possible voltage vectors in a four-switch three phase converter are explained. Later, two voltage vector sequences consisting of three voltage vectors and two proposed sectors are defined. The sector identification algorithm and the offset voltage control for the FSTP MSC are also presented in this section. In section 4.3, the active and reactive power slopes and the vector sequences for a FSTP GSC are presented. The effect of the voltage vector sequences on the duration time of the vectors are also investigated in this section and afterward, the overall model predictive power control with capacitor deviation control for the GSC of DFIG/PMSG is described.
- Finally, in section 4.4, the summary and conclusions of this chapter are drawn.

4.2. Model Predictive Current Control for the machine side converter of SPMSG when fed by four-switch three phase converter

4.2.1. Introduction

An important matter that should be considered in wind turbines or wave energy conversions is that the system should continue working during some hardware faults such as the breakdown of a switch, in order to prevent losing potential power from the waves or winds or to keep the control of the buoy in the case of floating OWCs to assure the buoy survival under storm conditions.

The four-switch three phase converter (FSTPC) is one of the fault-tolerance solutions that can be applied when one of the six switches or one of the converter legs is damaged. In this case, the midpoint of the DC link should be connected to the faulty phase, while the rest four switches continue working [15]-[16]. Although the converter can work with only four switches, unbalanced currents and high power ripples could appear in the system if the control is not properly designed for this fault situation [17].

FSTPC working as rectifiers, as shunt active filters or feeding PMS motors, have been investigated during recent years [18]-[21]. In [21], a comprehensive torque ripple reduction is presented for FSTPC feeding a PMSM, where a preferred space vector modulation (SVM) was proposed. The torque ripple that is induced by the voltage drop of the converter switches was eliminated by a compensation modulation strategy in the case of PMSM [22]. A hybrid SVM technique to reduce the torque ripple of a PMSM by using equivalent zero vectors is presented in [23]. This strategy determines the voltage sector location from the stator current position. In these last methods, the scope was focused on the modulation, while the current control strategy was not reported.

A predictive direct torque control for four-switch three phase (FSTP) inverter was introduced to drive an induction machine in [24]. In this strategy, only one voltage vector that minimizes a cost function was employed at every switching period. Although the voltage deviation suppression was also included in the cost function using a weighting factor, the value of the corresponding factor had a big influence on the control, to the extent that the performance of the machine was deteriorated. Besides, the switching frequency was variable, and the torque ripple was high. A model predictive current control (MPCC) for PMSM is presented in [25], introducing a four sectors division and a virtual duty cycle. According to the sector, one active voltage vector and the determined virtual duty cycle are employed. The effect of each voltage vector on the voltage deviation is studied in [26]. In order to remove the voltage offsets, it is necessary to extract the mean value of the capacitors which is usually done by using notch or low pass filters [23], [26]-[27]. This method was adopted in most of the research works to overcome the voltage offset. However, in this technique when the speed of the machine is low, the results are not good, and it presents a big delay when extracting the mean value.

To the best knowledge, FSTP connected to SPMSG for wave energy applications has not been reported. This is an important case since the inherent irregular waves can make the dc voltage offset to have big values if a fast offset compensation strategy is not adopted, especially during the low rotational speeds. In this case, one of the capacitor voltages could reach zero volts, while the entire DC link voltage is applied to the other capacitor, which could carry the breakdown of this capacitor. Additionally, a big offset in the capacitor voltages will prevent a good performance of the system. Consequently, due to the aforementioned reasons the proposed control for the FSTPC of the SPMSG and for the capacitor voltage offset, should have a fast dynamic response.

In this section, a simple and effective MPCC for the four-switch three phase machine side converter of the SPMSG is presented. It is capable of performing highly dynamic speed control and features a fixed switching frequency. In this method, only two sector divisions are defined, and three voltage vectors are applied at every switching period without using the stator voltage sector location. Moreover, a fast response offset control is introduced in an outer layer of the predictive control without including it in the cost function. The capacitor voltage deviations are quickly suppressed without using any filter. It altogether makes the system capable of working properly despite having to deal with large variations in rotor speed.

The proposed MPCC with a fixed switching frequency feature for the FSTPC connected to SPMSG is presented in section 4.2.2. In this section, the proposed three voltage vector sequences and the voltage offset control are explained while the effects of the vectors on the duration times are investigated.

4.2.2. Proposed MPCC for the SPMSG when fed by FSTPC

The proposed fault-tolerance topology for the MSC of a SPMSG is shown in Fig. 4.1. In this topology when a semiconductor open-circuit or short-circuit fault is detected and located, the faulty leg will be isolated and then by triggering the corresponding anti-parallel thyristors or triacs the faulty phase will be connected to the midpoint of the DC bus as shown

in Fig. 4.1. The semiconductor open-circuit and short-circuit faults can be diagnosed by different methods such as modified normalized dc current, current deviation, Wavelet, fuzzy, and neural network-based methods, di/dt feedback control, and vector composition of output voltage methods [28].

Fig. 4.1. Fault-tolerance topology schematic for the machine side converter of a SPMSG.

In this thesis, it is assumed that the faulty leg, which is phase “a”, was detected and isolated and the phase “a” is connected to the midpoint of the dc capacitors. As a result, the structure of the proposed FSTPC for the SPMSG is shown in Fig. 4.2.

Fig. 4.2. Structure of a FSTPC connected to a SPMSG.

4.2.2.1. Current slopes

The dynamic equations for a SPMSG in rotor rotating reference frame, can be presented as [20]

$$u_{sd} = R_s i_{sd} + \frac{d}{dt} \lambda_{sd} - \omega_m \lambda_{sq} \quad (4.1)$$

$$u_{sq} = R_s i_{sq} + \frac{d}{dt} \lambda_{sq} + \omega_m \lambda_{sd} \quad (4.2)$$

Where u_s , i_s , λ_s , ω_m and R_s are the stator voltage, the stator current, the stator flux, the electrical rotor speed and the stator resistance, respectively. The SPMSG stator flux in the rotor reference frame can be express as

$$\lambda_{sd} = L_s i_{sd} + \lambda_f \quad (4.3)$$

$$\lambda_{sq} = L_s i_{sq} \quad (4.4)$$

where λ_f and L_s represent the permanent magnets flux and the stator inductance, respectively. The current slopes in dq frame can be obtained according to (4.1)-(4.4) as

$$\frac{d}{dt} i_{sd} = \frac{1}{L_s} (-R_s i_{sd} + \omega_m L_s i_{sq} + u_{sd}) \quad (4.5)$$

$$\frac{d}{dt} i_{sq} = \frac{1}{L_s} (-R_s i_{sq} - \omega_m L_s i_{sd} - \omega_m \lambda_f + u_{sq}) \quad (4.6)$$

As a result, the stator current derivatives can be derived according to the machine parameters, the speed, the permanent magnet flux, the currents and the applied stator voltage.

4.2.2.2. Four-switch three phase converter voltage vectors

Four possible combinations of the switch states can be taking place in the FSTPC. When the faulty phase is the phase “a”, Fig. 4.2, the converter voltage in the abc frame can be presented as

$$\begin{cases} V_{sa} = \frac{1}{3} v_{c1} (-S_b - S_c) + \frac{1}{3} v_{c2} (2 - S_b - S_c) \\ V_{sb} = \frac{1}{3} v_{c1} (2S_b - S_c) + \frac{1}{3} v_{c2} (2S_b - S_c - 1) \\ V_{sc} = \frac{1}{3} v_{c1} (2S_c - S_b) + \frac{1}{3} v_{c2} (2S_c - S_b - 1) \end{cases} \quad (4.7)$$

Where V_{sa} , V_{sb} , V_{sc} are the converter voltages in abc frame, v_{c1} and v_{c2} are the upper and lower capacitor voltages and S_b and S_c are the switching states of the phase b and c, respectively, which can be 1 or 0. The converter voltage in the abc frame can be transferred to the stationary α - β frame according to the Clarke transformation which can be presented as

$$\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix} = \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} v_{sa} \\ v_{sb} \\ v_{sc} \end{bmatrix} \quad (4.8)$$

As a result, the converter voltage vectors in the α - β frame can be obtained from (4.7) and (4.8). The four possible voltage vectors when the faulty phase is “a”, are shown in Fig. 4.3, and the corresponding values for these four vectors in the stationary and abc frames are shown in this Table 4.1. As can be seen in Table 4.1, the voltage values of V_2 and V_4 in α axis depend on the differences between the upper and lower capacitor voltages. Hence, these two vector phases could change as shown in Fig. 4.3. For example, when the upper capacitor has a bigger voltage than the lower capacitor, V_2 and V_4 will be positioned like the blue lines shown in Fig. 4.3.

The stator current slopes for these four voltage vectors can be calculated according to the corresponding voltage values using (4.5) and (4.6).

Fig. 4.3. Voltage vectors positions in the stationary reference frame for a FSTPC. Red vectors: $v_{c1} = v_{c2}$; green vectors: $v_{c1} < v_{c2}$; blue vectors: $v_{c1} > v_{c2}$.

Table 4.1. The four voltage vector values for the FSTPC in the “ $\alpha\beta$ ” stationary and “abc” frame

S_b	S_c	Vector	V_{sa}	V_{sb}	V_{sc}	V_{sa}	V_{sb}
0	0	V_1	$2v_{c2}/3$	$-v_{c2}/3$	$-v_{c2}/3$	$2v_{c2}/3$	0
1	0	V_2	$(v_{c2} - v_{c1})/3$	$(2v_{c1} + v_{c2})/3$	$-(v_{c1} + 2v_{c2})/3$	$(v_{c2} - v_{c1})/3$	$\sqrt{3}(v_{c1} + v_{c2})/3$
1	1	V_3	$-2v_{c1}/3$	$v_{c1}/3$	$v_{c1}/3$	$-2v_{c1}/3$	0
0	1	V_4	$(v_{c2} - v_{c1})/3$	$-(v_{c1} + 2v_{c2})/3$	$(2v_{c1} + v_{c2})/3$	$(v_{c2} - v_{c1})/3$	$-\sqrt{3}(v_{c1} + v_{c2})/3$

4.2.2.3. Voltage vector sequences and sector division definition

Three vectors among the four possible voltage vectors are employed in every switching period according to the stator voltage sector, to have fixed switching frequency and low current ripple. In this strategy, two stator voltage sectors are defined, as can be seen in Fig. 4.3. If the equivalent stator voltage value in β axis is positive, the stator sector is one (S_1), otherwise, the sector is two (S_2). The proposed voltage vector sequences according to

the corresponding sector are shown in Table 4.2. As can be seen, in each switching period, V_1 and V_3 will be applied. A part of V_1 or V_3 , and the whole opposite vector will produce the equivalent zero vector.

Table 4.2. Voltage vector sequences according to the stator voltage sector

Vector sequence for sector 1:	V_1	V_2	V_3
Vector sequence for sector 2:	V_1	V_4	V_3

The switching pattern of the proposed voltage vector sequences when the equivalent stator voltage is in sector 2 (S_2) is shown in Fig. 4.4. Three voltage vectors are symmetrically employed in each period. As a result, the stator current at the end of period ($k+1$) can be presented as

$$i_{sq}(k+1) = i_{sq}(k) + S_{qa}t_a + S_{qb}t_b + S_{qc}t_c \quad (4.9)$$

$$i_{sd}(k+1) = i_{sd}(k) + S_{da}t_a + S_{db}t_b + S_{dc}t_c \quad (4.10)$$

Fig. 4.4. Proposed MPCC Switching pattern for the FSTPC when the stator voltage is in S_2 .

Where S_{qa} , S_{qb} , S_{qc} and S_{da} , S_{db} , S_{dc} are the current derivatives of the corresponding voltage vectors in q and d axis, respectively. The duration times of the first, the second and the third vector are called t_a , t_b and t_c , respectively.

The error between the measured current and its reference value at the end of a switching period can be minimized by the following cost function

$$G(k+1) = \left(i_{sq}(k+1) - i_{sq}^*(k) \right)^2 + \left(i_{sd}(k+1) - i_{sd}^*(k) \right)^2 \quad (4.11)$$

Where the superscript * represents the reference values. Substituting (4.9) and (4.10) into (4.11) and considering that the sum of the duration times of the voltage vectors are equal to the switching period time, leads to

$$G(k+1) = (i_{sq}(k) + S_{qa}t_a + S_{qb}t_b + S_{qc}(T_s - t_a - t_b) - i_{sq}^*(k))^2 + (i_{sd}(k) + S_{da}t_a + S_{db}t_b + S_{dc}(T_s - t_a - t_b) - i_{sd}^*(k))^2 \quad (4.12)$$

The optimized duration times which minimize the cost function can be achieved by setting to zero the partial derivatives of (4.12) with respect to t_a and t_b . Consequently, the optimized duration time of these three vectors can be obtained as

$$t_a = \frac{(i_{sq}(k) - i_{sq}^*(k)) \cdot (S_{db} - S_{dc}) + (i_{sd}(k) - i_{sd}^*(k)) \cdot (S_{qc} - S_{qb})}{S_{da}(S_{qb} - S_{qc}) + S_{db}(S_{qc} - S_{qa}) + S_{dc}(S_{qa} - S_{qb})} + \frac{T_{sw}(S_{db} \cdot S_{qc} - S_{dc} \cdot S_{qb})}{S_{da}(S_{qb} - S_{qc}) + S_{db}(S_{qc} - S_{qa}) + S_{dc}(S_{qa} - S_{qb})} \quad (4.13)$$

$$t_b = \frac{(i_{sq}(k) - i_{sq}^*(k)) \cdot (S_{dc} - S_{da}) + (i_{sd}(k) - i_{sd}^*(k)) \cdot (S_{qa} - S_{qc})}{S_{da}(S_{qb} - S_{qc}) + S_{db}(S_{qc} - S_{qa}) + S_{dc}(S_{qa} - S_{qb})} + \frac{T_{sw}(S_{dc} \cdot S_{qa} - S_{da} \cdot S_{qc})}{S_{da}(S_{qb} - S_{qc}) + S_{db}(S_{qc} - S_{qa}) + S_{dc}(S_{qa} - S_{qb})} \quad (4.14)$$

$$t_c = T_{sw} - t_a - t_b \quad (4.15)$$

Where T_{sw} is the switching period time.

4.2.2.4. Identification algorithm for the stator voltage sector and correction of the duration times

Fig. 4.5(a) and 4.5(b) show the duration times of the two possible voltage vector sequences during one rotational cycle of the rotor in steady state when the speed is 400 rpm. As can be seen in Fig. 4.5(a), the duration times of the voltage vector sequence of sector 1, duration times of $(V_1-V_2-V_3)$, are all positive from the beginning until $t=0.025$ s. However, from $t=0.025$ s to the end of the cycle, the duration time for V_2 has a negative value. In contrast, according to Fig. 4.5(b), for the other voltage vector sequence $(V_1-V_4-V_3)$, from the beginning of the cycle to $t=0.025$ s, the duration time of V_4 has a negative value and from $t=0.025$ s to the end of the cycle, the duration times of all vectors of this vector sequence are positive values. This case shows that, from $t=0$ s to $t=0.025$ s and from $t=0.025$ s to $t=0.05$ s, the stator voltage reference is located in sector 1 and sector 2, respectively. Consequently, when the duration time of the second vector (t_b) has a negative value in the selected vector sequence, it indicates that the correct vector sequence is the other vector sequence.

Another case can arise during transients or when the required stator voltage is higher than that the converter can generate, since in this situation there is no need to generate a zero-voltage vector by the combination of the two opposite vectors V_1 and V_3 . Hence, the duration time of V_1 or V_3 (t_a or t_c) could also become negative as Fig. 4.6(a) and Fig. 4.6(b) show.

Fig. 4. 5. Duration times of the voltage vector sequences at steady state for: (a) V_1 , V_2 and V_3 ; (b) V_1 , V_4 and V_3 .

Fig. 4.6. Duration times of the voltage vector sequences in a transient state for: (a) V_1 , V_2 and V_3 ; (b) V_1 , V_4 and V_3 .

As a result, the vector that has a negative duration time should be removed from the voltage vector sequence and the other two voltage vectors should be normalized and applied by the machine side converter (MSC). For example, if the required voltage is high, and it is located at the left side of sector 1, the correct vectors are V_2 and V_3 and there is no need to apply V_1 . For instance, as it is shown in Fig. 4.6(a), from $t=0$ s to $t=0.013$ s the duration times of V_1 and V_2 are both positive whereas the time for V_3 is negative. This fact shows that the required stator voltage is located at the right side of sector 1, and there is no need to apply

V₃. Another case is from $t = 0.013$ s to $t = 0.025$ s when the duration time of V₁ is negative and the duration times of V₂ and V₃ are positive. In this case, there is no need to apply V₁. As a result, when the estimated duration time of the first or the third voltage vector (t_a or t_c) has a negative value, the corresponding vector is removed, and the two other vectors should be normalized, which means

$$\text{if } t_c \leq 0 \Rightarrow \begin{cases} t_a' = \frac{t_a}{t_a + t_b} T_s \\ t_b' = \frac{t_b}{t_a + t_b} T_s \\ t_c' = 0 \end{cases} \quad \text{if } t_a \leq 0 \Rightarrow \begin{cases} t_a' = 0 \\ t_b' = \frac{t_b}{t_b + t_c} T_s \\ t_c' = \frac{t_c}{t_b + t_c} T_s \end{cases} \quad (4.16)$$

Where t_a' , t_b' and t_c' are the normalized values of the t_a , t_b and t_c , respectively. These duration times of the vectors are applied when the duration time of the first or the third vector in the vector sequence gets negative. In this case, the output voltage will be limited to the maximum voltage of the converter without overmodulation.

As aforementioned, when the duration time of second vector (t_b) is negative, the correct vector sequence is the other one. This case also can be seen in Fig. 4.6(a) and Fig. 4.6(b). As can be seen in this figure, when the duration time of V₂ is negative, the duration time of V₄ is positive and vice versa.

In summary, when the duration time of the second vector (t_b) has a negative value during any situation, transient or steady state, the correct sector is not the one chosen but the other one, and it is necessary to calculate the duration times of the other vector sequence. Moreover, if the duration time of the first or third vector becomes negative, the corresponding vector should be removed, and the duration time of the other two vectors should be normalized according to (4.16).

4.2.2.5. Voltages Offset elimination control

A voltage offset can appear between the upper and the lower capacitors of the DC link due to the AC current flowing through the capacitors. This voltage deviation cause to have unbalanced three phase currents.

According to Fig. 4.2, the stator current that circulates through the capacitors, in phase “a”, can be expressed by the two capacitors voltages as

$$i_{sa} = C \frac{dv_{c1}}{dt} - C \frac{dv_{c2}}{dt} \quad (4.17)$$

Where C is the capacitors value in the DC link and V_{c1} and V_{c2} are upper and lower capacitor voltages respectively. The voltage differences between the upper and the lower capacitors can be derived as

$$v_{c1}(t) - v_{c2}(t) = \frac{1}{C} \int_0^t i_{sa} dt + v_{c1}(0) - v_{c2}(0) \quad (4.18)$$

As a result, the voltage difference is related to the initial value of the capacitors voltage and the faulty phase current. The three phase stator currents will be unbalanced when a voltage offset exists in the capacitors. In this case, the stator currents can be expressed by

$$i_{sa} = I_s \cos(\omega_m t) - \bar{i}_{so} \quad (4.19)$$

$$i_{sb} = I_s \cos(\omega_m t - \frac{2\pi}{3}) + \frac{1}{2} \bar{i}_{so} \quad (4.20)$$

$$i_{sc} = I_s \cos(\omega_m t + \frac{2\pi}{3}) + \frac{1}{2} \bar{i}_{so} \quad (4.21)$$

Where i_{sa} , i_{sb} , i_{sc} are the three phase stator currents, I_s is the amplitude of the AC current and $\bar{i}_{so}$ is the DC component current. By transferring the three phase stator currents to the stationary frame, according to (4.19)-(4.21), the current yields to

$$i_{s\alpha} = I_s \cos(\omega_m t + \alpha) - \bar{i}_{so} \quad (4.22)$$

$$i_{s\beta} = I_s \sin(\omega_m t + \alpha) \quad (4.23)$$

As can be seen, the DC component of the current will only be in α axis. According to (4.18) and (4.19), the capacitor voltage differences can be represented as

$$\begin{aligned} v_{c1}(t) - v_{c2}(t) &= \frac{1}{C\omega_m} I_s \sin(\omega_m t + \alpha) - \frac{\bar{i}_{so}}{C} t - \frac{1}{C\omega_m} I_s \sin(\alpha) + v_{c1}(0) \\ &\quad - v_{c2}(0) \end{aligned} \quad (4.24)$$

In (4.24), only the first term is an AC term in the right side of the equation and the rest of the terms are the DC terms that should have a zero value in order to eliminate the offset. Therefore, the $\bar{i}_{so}$ needs to be regulated to remove the voltage offset.

If $\frac{i_{s\beta}}{C\omega_m}$ is subtracted from both sides in (4.24), and (4.23) is substituted in the right side of (4.24), the result yields to

$$v_{c1}(t) - v_{c2}(t) - \frac{i_{s\beta}}{C\omega_m} = -\frac{1}{C\omega_m} I_s \sin(\alpha) - \frac{\bar{i}_{so}}{C} t + v_{c1}(0) - v_{c2}(0) \quad (4.25)$$

This equation indicates that by subtracting $\frac{i_{s\beta}}{C\omega_m}$ from the capacitor voltage difference, the DC voltage offset can be obtained. Therefore, if the right side of (4.25) is kept equal to zero, the DC voltage offset will be eliminated. As can be seen in (4.25), this goal can be achieved by only controlling the $\bar{i}_{so}$ current. According to (4.22) and (4.23), this DC current only exists in the α axis component of the stator current. Consequently, by regulating the left

side of equation (4.25) with a PI controller, the compensation current for removing the voltage offset in the α axis can be presented as

$$i_{s\alpha-c}^* = -k_p(v_{c1}(t) - v_{c2}(t) - \frac{i_{s\beta}}{C\omega_m}) - k_i \int (v_{c1}(t) - v_{c2}(t) - \frac{i_{s\beta}}{C\omega_m}) dt \quad (4.26)$$

Where the k_p and K_i are the proportional and integral gains of the PI controller, respectively. This compensation current can be transferred to the synchronous frame as

$$i_{sd-c}^* = i_{s\alpha-c}^* \cos(\theta_m) \quad (4.27)$$

$$i_{sq-c}^* = -i_{s\alpha-c}^* \sin(\theta_m) \quad (4.28)$$

Therefore, the compensation current reference in dq frame for eliminating the offset voltage can be obtained from (4.27) and (4.28). The block diagram of the proposed voltage offset elimination control is presented in Fig. 4.7.

Fig. 4.7. Block diagram of the proposed voltage deviation control.

When the voltage offset elimination control is working, the upper and lower capacitor voltages can be expressed by

$$v_{c1}(t) = -v_{c2}(t) = \frac{1}{2} \frac{1}{C\omega_m} I_s \sin(\omega_m t + \alpha) + \frac{V_{dc}}{2} \quad (4.29)$$

The ac term in the capacitor voltage is dependent on the machine speed, the stator current and the capacitance of the capacitor. So, some type of over voltage protection must be provided as a preventive measure, mainly at low speeds.

4.2.2.6. Overall proposed model predictive direct current control with voltage Suppression Control for FSTPC fed SPMSG

The block diagram of the proposed MPCC with voltage suppression control for the FSTPC fed SPMSG is shown in Fig. 4.8. As can be seen, the current references in the synchronous frame consist of two terms, one is the compensatory term to suppress the voltage offset in the capacitors, which is obtained according to Fig. 4.7, and the other term is the reference value which should be chosen in a way that the maximum power could be obtained from the wind or waves.

In each period of this control strategy, three voltage vectors are selected according to Table 4.2. It is assumed that the vector sequence is the same as those in the previous period. The current slopes in the synchronous frame are calculated according to (4.5)-(4.6) using the rotor speed and the stator current for these three voltage vectors. The duration times of this vector sequence are obtained according to the estimated current slopes and the current references in dq frame, according to (4.13)-(4.15). If these estimated times are positive, this vector sequence and their corresponding times will be applied by the converter. But, if the second vector duration time obtains a negative value, the vector sequence should be changed to the other vector sequence, and the corresponding new durations should be calculated, which led to have a positive duration for the second voltage vector in this new sequence. After this step, as it is shown in Fig. 4.8, the first and third vector times will be analyzed to check if they are positive or negative. If one of these vectors is negative, the duration times of these vectors are modified according to (4.16). Finally, these vectors and their durations are applied to the SPMSG by the machine side converter.

Fig. 4.8. Block diagram of the proposed MPCC for FSTPC fed SPMSG.

4.3. Model Predictive Direct Power Control Strategy for Four-Switch Three Phase Operation of Grid Side Converter

4.3.1. Introduction

As it was mentioned in the beginning of this chapter, the DC bus midpoint connection or four-switch three phase (FSTP) converter is an interesting fault tolerant topology. After a fault is detected in one phase, the four intact switches continue to work in the three phase system and the DC bus midpoint is connected to the faulty phase. Although the FSTP system presents low conduction losses (with good capacitors in the DC link), the voltage generated by the converter is reduced by fifty percent, and the capacitor voltage will present fluctuations

during the fault operation if compared to the normal six switches operation [29].

During the fault-tolerant operation, the switching frequency is the same as normal operation, and the currents are practically the same. On the one hand, this topology uses four-switches instead of six switches and from this point of view, the total switching losses should be lower. On the other hand, the DC voltage is bigger than before the fault, which leads to have higher power losses in each switch. However, this increment in switching losses depends on the DC link voltage chosen for normal operation and the DC voltage value for FSTP.

In [30] and [31], an easy and feasible vector based PWM method was suggested for FSTP inverter. The Lyapunov function based nonlinear controller was used to follow the inverter current in abc frame and to obtain a decoupled active and reactive power control [31]. Nevertheless, this method needs a PWM control and coordinate transformations.

In order to reduce the unbalanced currents in FSTP inverters and to remove the grid voltage sector location, an adaptive PWM was presented in [32]. The common voltage and the effect of the passing AC current through the DC link capacitors, as well as the current THD, were investigated when the zero voltage was applied to the FSTP rectifier, and a unified space vector modulation was presented [33]. However, sector identification is necessary in this algorithm. In [34], the unbalanced DC link voltage in FSTP rectifiers is eliminated by using an optimized modulation strategy and the capacitor current is minimized. This strategy was improved by a hybrid space vector modulation, and the current stress in the capacitors was decreased in [35].

A model predictive current control (MPCC) for the FSTP rectifiers was suggested in [36], where only one voltage vector between the four possible voltage vectors is applied in each sector to reduce the current error. In [37], a two vector MPCC was employed for FSTP inverters. Moreover, one voltage vector predictive torque control was presented for driving induction machines and the voltage offsets of capacitors were considered [24]. Although the proposed predictive controls feature a fast dynamic response for FSTP systems, the switching frequency of the converters is variable and the current total harmonic distortion (THD) is high. Moreover, the sample time is also high.

In [17], a predictive control for the four-switches three phases rectifier is presented where two active vectors, and one equivalent zero vector are used. The first active vector is the one that has the lowest value in the presented cost function, and the other active vector is the adjacent vector that has the lower value of the mentioned cost function. After that, the duration times of the vectors are estimated. Moreover, two low pass filters are used in this method to obtain the mean value of the capacitor voltages and according to the differences in these two values, the durations of V_1 and V_3 are increased or decreased in order to suppress the capacitor voltage deviation. Although, it features fixed switching frequency and better performance compared to other mentioned predictive controls, it is needed to calculate the cost function value for each of the four voltage vectors before calculating the duration times. Moreover, if the duration time of each of the two active vectors obtains a negative value, or if they are larger than the switching period, they are limited to zero and T_s , respectively.

In [38], [27], the authors used a vector control strategy for the FSTP GSC of the DFIG and proposed a new space vector pulse width modulation (SVPWM). Although the study

considered the bidirectional operation of the GSC, the proposed technique suffered from the inherent slow dynamic response of vector control, and the experimental result was not reported.

This section proposes a predictive direct power control for the FSTP GSC in which three voltage vectors are employed within switching periods and switching frequency is constant. The proposed strategy can keep the DC link voltage at a fixed value during the sub- and hyper-synchronous speed of DFIG for any operation point while three phase GSC currents are sinusoidal. Moreover, in the proposed strategy, the DC link voltage deviation is effectively and easily suppressed without using any low pass filter. The proposed strategy does not need neither a PLL control nor a grid voltage position detector.

The proposed predictive direct power control for the FSTP grid side converter is presented in section 4.3.2. This section presents the calculations of the power slopes, the proposed voltage vector sequences and the appropriate vectors, as well as the elimination of the DC link voltage offset. Finally, the overall block diagram of the proposed predictive power control method is presented at the end of this section.

4.3.2. Predictive control for FSTP GSC

4.3.2.1. FSTP GSC power slopes

The structure of the FSTP GSC is shown in Fig. 4.9. In this paper, it is assumed that a fault has occurred within the switches of the leg “a” and so the phase “a” is connected at the midpoint of the capacitors.

Fig. 4.9. Structure of FSTP grid side converter of the DFIG/SPMSG.

According to this figure, the FSTP GSC voltage vector in the stationary frame can be represented by

$$\vec{v}_{GSC} = \vec{u}_g + L_{fg} \frac{d\vec{i}_g}{dt} + R_{fg} \vec{i}_g \quad (4.30)$$

Where v_{GSC} , u_g , i_g , L_{fg} and R_{fg} are the grid side converter voltage, the grid voltage, the grid current, the grid side inductance and resistance, respectively. The active and reactive power that can be injected by the GSC can be calculated as

$$P_g = 1.5(u_{g\alpha}i_{g\alpha} + u_{g\beta}i_{g\beta}) \quad (4.31)$$

$$Q_g = 1.5(u_{g\beta}i_{g\alpha} - u_{g\alpha}i_{g\beta}) \quad (4.32)$$

Where P_g, Q_g are the GSC active and reactive power, respectively. The subscript α and β are the stationary axes. The derivatives of the GSC active and reactive power, according to (4.31) and (4.32), can be estimated as

$$\frac{dP_g}{dt} = 1.5\left(\frac{du_{g\alpha}}{dt}i_{g\alpha} + \frac{di_{g\alpha}}{dt}u_{g\alpha} + \frac{du_{g\beta}}{dt}i_{g\beta} + \frac{di_{g\beta}}{dt}u_{g\beta}\right) \quad (4.33)$$

$$\frac{dQ_g}{dt} = 1.5\left(\frac{du_{g\beta}}{dt}i_{g\alpha} + \frac{di_{g\alpha}}{dt}u_{g\beta} - \frac{du_{g\alpha}}{dt}i_{g\beta} - \frac{di_{g\beta}}{dt}u_{g\alpha}\right) \quad (4.34)$$

To calculate the active and reactive power slopes, it is necessary to calculate the derivatives of the grid voltage and current in the stationary reference frame. The grid voltage in a balanced system can be represented as

$$u_{g\alpha} = |\vec{u}_g| \sin(\omega_s t) \quad (4.35)$$

$$u_{g\beta} = -|\vec{u}_g| \cos(\omega_s t) \quad (4.36)$$

Where the ω_s is the grid angular frequency. The derivative of the grid voltage in the stationary frame yields

$$\frac{du_{g\alpha}}{dt} = \omega_s |\vec{u}_g| \cos(\omega_s t) = -\omega_s u_{g\beta} \quad (4.37)$$

$$\frac{du_{g\beta}}{dt} = \omega_s |\vec{u}_g| \sin(\omega_s t) = \omega_s u_{g\alpha} \quad (4.38)$$

Hence, the grid voltage derivatives can be represented by the grid voltage and the synchronous speed. According to (4.30), the derivative of the grid current can be described as

$$\frac{di_{g\alpha}}{dt} = \frac{(v_{GSC\alpha} - u_{g\alpha} - R_{fg}i_{g\alpha})}{L_{fg}} \quad (4.39)$$

$$\frac{di_{g\beta}}{dt} = \frac{(v_{GSC\beta} - u_{g\beta} - R_{fg}i_{g\beta})}{L_{fg}} \quad (4.40)$$

The GSC active and reactive power derivatives can be represented by substituting (4.37)-(4.40) into (4.33) and (4.34). Consequently, the power slopes can be derived as

$$\frac{dP_g}{dt} = \frac{1.5}{L_{fg}} [(u_{g\alpha}v_{GSC\alpha} + u_{g\beta}v_{GSC\beta}) - (u_{g\alpha}^2 + u_{g\beta}^2)] - \frac{R_s}{L_{fg}} P_g - \omega_s Q_g \quad (4.41)$$

$$\frac{dQ_g}{dt} = \frac{1.5}{L_{fg}} (u_{g\alpha} v_{GSC\beta} - u_{g\beta} v_{GSC\alpha}) - \frac{R_s}{L_{fg}} Q_g + \omega_s P_g \quad (4.42)$$

The active and reactive power slopes for the voltage vectors of the GSC can be estimated according to (4.41) and (4.42).

4.3.2.2. FSTP GSC voltage vectors

The FSTP GSC can produce four voltage vectors according to the possible combinations of the four switches states where none of them are zero vectors. According to Fig. 4.9, the phase “a” is connected at the midpoint of the DC capacitors. The four possible voltage vector values in the stationary frame for this FSTP are given in Table 4.3. In this table, S_b and S_c are the switching states of the phase b and c, respectively.

Table 4.3. The four possible voltage vector values for the FSTP GSC in the stationary frame

S_b	S_c	Vector	$V_{GSC\alpha}$	$V_{GSC\beta}$
0	0	V_1	$2v_{c2}/3$	0
1	0	V_2	$(v_{c2} - v_{c1})/3$	$\sqrt{3}(v_{c1} + v_{c2})/3$
1	1	V_3	$-2v_{c1}/3$	0
0	1	V_4	$(v_{c2} - v_{c1})/3$	$-\sqrt{3}(v_{c1} + v_{c2})/3$

These four vectors are active vectors and the voltage value in α axis for the V_2 and V_4 can be zero if the upper and lower capacitor voltages have the same value. However, the ripple in the two capacitors would introduce a variation in module and phase of these vectors. As a result, three cases could be possible for the GSC voltage vectors. The position of the voltage vectors for these three cases is shown in Fig. 4.10.

Fig. 4.10. The position of the voltage vectors in the stationary reference frame: (a) $v_{c1} = v_{c2}$; (b) $v_{c1} < v_{c2}$; (c) $v_{c1} > v_{c2}$.

The active and reactive power slopes when each of these four voltage vectors is employed can be calculated according to (4.41) and (4.42), and the corresponding voltage value of the employed vector that is presented in Table 4.3.

4.3.2.3. Proposed voltage vector sequences for FSTP GSC

In every period of this strategy, three vectors are selected according to the GSC voltage sector. The proposed voltage vector sequences are illustrated in Table 4.4. In this table, the V_a , V_b and V_c are the first, second and third vector in a period. In this strategy, only two sectors are defined. The sector ranges can be seen in Fig. 4.10. The upper side of the β axis is sector one, and the lower side of this axis is sector 2. The sector areas will not change when the position of vectors is changed (V_2 and V_4). In order to build the desired voltage following the reference values, the two nearest voltage vectors are selected, and the other vector is employed to have constant switching frequency and to decrease the power ripple. In other words, since the zero vector does not exist in FSTP, a part of V_1 and V_3 (opposite vectors) produce a zero voltage vector. For the sake of example, suppose that the required GSC voltage for a switching period is located in sector 1 and it is between the vectors V_1 and V_2 . In this case, in order to build the desire voltage, V_1 and V_2 are used as fundamental vectors and the V_3 and a part of V_1 produce a zero vector. In this case, obviously the duration of vector V_3 is smaller than the duration of V_1 .

Table 4.4. Proposed Voltage vector sequences

Vector sequences		V_a	V_b	V_c
sector (N)	1	V_1	V_2	V_3
	2	V_1	V_4	V_3

After applying the proposed voltage vector sequences, the GSC active and reactive powers at the end of the period ($k+1$ instant) can be represented as

$$P_g(k+1) = P_g(k) + S_{Pa}t_a + S_{Pb}t_b + S_{Pc}t_c \quad (4.43)$$

$$Q_g(k+1) = Q_g(k) + S_{Qa}t_a + S_{Qb}t_b + S_{Qc}t_c \quad (4.44)$$

Where S_{Pa} , S_{Pb} , S_{Pc} and S_{Qa} , S_{Qb} , S_{Qc} are active and reactive power slopes of the corresponding voltage vectors. The t_a , t_b , t_c are duration time of corresponding voltage vectors.

The power slopes of the selected vectors can be calculated according to (4.41) and (4.42). In order to decrease the power errors at the end of the period, the following standard cost function is used

$$F(k+1) = (Q_g(k+1) - Q_g^*(k))^2 + (P_g(k+1) - P_g^*(k))^2 \quad (4.45)$$

In order to obtain the optimal duration times of voltage vectors, which minimizes the cost function, the partial derivatives of (4.45) with respect to t_a and t_b are set to zero. By substituting (4.43) and (4.44) into (4.45) and solving the mentioned partial derivatives equations, the optimized duration time of the vectors yields

$$t_a = \frac{(P_g(k) - P_g^*(k)) \cdot (S_{Qb} - S_{Qc}) + (Q_g(k) - Q_g^*(k)) \cdot (S_{Pc} - S_{Pb})}{S_{Qa}(S_{Pb} - S_{Pc}) + S_{Qb}(S_{Pc} - S_{Pa}) + S_{Qc}(S_{Pa} - S_{Pb})} + \frac{T_{sw}(S_{Qb} \cdot S_{Pc} - S_{Qc} \cdot S_{Pb})}{S_{Qa}(S_{Pb} - S_{Pc}) + S_{Qb}(S_{Pc} - S_{Pa}) + S_{Qc}(S_{Pa} - S_{Pb})} \quad (4.46)$$

$$t_b = \frac{(P_g(k) - P_g^*(k)) \cdot (S_{Qc} - S_{Qa}) + (Q_g(k) - Q_g^*(k)) \cdot (S_{Pa} - S_{Pc})}{S_{Qa}(S_{Qb} - S_{Pc}) + S_{Qb}(S_{Pc} - S_{Pa}) + S_{Qc}(S_{Pa} - S_{Pb})} + \frac{T_{sw}(S_{Qc} \cdot S_{Pa} - S_{Qa} \cdot S_{Pc})}{S_{Qa}(S_{Pb} - S_{Pc}) + S_{Qb}(S_{Pc} - S_{Pa}) + S_{Qc}(S_{Pa} - S_{Pb})} \quad (4.47)$$

$$t_c = T_{sw} - t_a - t_b \quad (4.48)$$

Where T_{sw} is the switching period time. When the vector sequences are selected and the duration times are estimated, the vectors will be applied like they appear in Fig. 4.11. In this figure, it is assumed that the required GSC voltage is located in the first sector.

Fig. 4.11. Switching pattern for the IGBTs when the required voltage is in first sector.

4.3.2.4. Effect of the voltage vector sequences on the duration times

In this part, the duration times of the voltage vector sequences will be further investigated for the GSC. The duration time of the voltage vector sequences, for one complete cycle of the grid voltage at steady state, and for both sectors, are shown in Fig. 4.12(a) and 4.12(b). In this case, the required voltage is smaller than the maximum voltage that can be produce by the converter ($VGSC_max = \min(V_{c1}, V_{c2}) / \sqrt{3}$). The duration times of the vectors when the V_1, V_2, V_3 are selected (required GSC voltage is in sector one), are shown in Fig. 4.12(a). In this figure, the duration time of the vectors when $t=0.006$ s to $t=0.016$ s is positive, which means that these three voltage vectors could be applied by the converter. However, the duration time of V_2 (t_b) takes a negative value when t is between $[0, 0.006]$ or $[0.016, 0.0]$. In return, in this area, the duration times of V_1, V_4, V_3 are positive as it can be seen. As a result, these two figures indicate that, when the duration time of V_b gets negative, the appropriate voltage vector sequences are located in the other sector, in which the durations of all vectors will be positive.

The duration time of the first vector and the third vector also could get negative when the required GSC voltage in a period is bigger than that what can be produced by the converter ($VGSC_max$) or in the transient state. The duration times of voltage vector sequences for this case are shown in Fig. 4.13. Because V_1 and V_3 have opposite values, when one of the duration times of these vectors gets a big positive value, the other one gets negative. In this case, the vector corresponding to the negative duration time must be discarded. For example, according to Fig. 4.13(a), when t is between $[0.009, 0.015]$, the durations of V_1 and

Fig. 4.12. Duration times of the voltage vector sequences at steady state: (a) durations of V_1 , V_2 and V_3 ; (b) durations of V_1 , V_4 and V_3 .

Fig. 4.13. Duration times of the voltage vector sequences at transient state: (a) durations of V_1 , V_2 and V_3 ; (b) durations of V_1 , V_4 and V_3 .

V_2 are positive, while the V_3 is negative. Since the total duration times of V_1 and V_2 are bigger than the switching time period, the duration time of V_3 gets negative. This case shows that the required voltage is located between V_1 and V_2 and the required voltage is high. Therefore, there is no need to apply V_3 . In other case, the duration time of V_1 gets negative while the durations of V_2 and V_3 are positive (when t is between $[0.015, 0.02]$ or $[0, 0.001]$). This example also points out that the required voltage does not need V_1 . It should be noted that, in this situation, when the duration time of V_2 (t_b) is negative, the duration time of V_4 is positive and vice versa. As aforementioned, in this case, the voltage sequence should be changed to the other voltage sequence.

Subsequently, it can be concluded from this part that, when the duration time of the V_b (t_b) gets negative, whether t_a and t_c are positive or negative, the required voltage is located in the other sector and the duration times for other voltage vector sequence have to be calculated. When the duration time of t_a or t_c gets negative and the t_b is positive, the voltage vector that gets negative will not be used, and the duration times of the other two vectors should be normalized as follows:

$$\text{if } t_c \leq 0 \Rightarrow \begin{cases} t_a' = \frac{t_a}{t_a + t_b} T_s \\ t_b' = \frac{t_b}{t_a + t_b} T_s \\ t_c' = 0 \end{cases} \quad \text{if } t_a \leq 0 \Rightarrow \begin{cases} t_a' = 0 \\ t_b' = \frac{t_b}{t_b + t_c} T_s \\ t_c' = \frac{t_c}{t_b + t_c} T_s \end{cases} \quad (4.49)$$

Where t_a' , t_b' and t_c' are the modified duration times of corresponding voltage vectors.

4.3.2.5. Voltages Offset Suppression control

Since sinusoidal current passes through the capacitors, a capacitor voltage deviation and a voltage offset could show up in the system. This voltage offset can be produced by the initial values of the grid current in phase "a" or by the unbalanced current that passes through the capacitors.

The dynamic equation for the current which is connected to the capacitor's midpoint can be expressed as

$$i_{ga} = C \frac{dv_{c1}}{dt} - C \frac{dv_{c2}}{dt} \quad (4.50)$$

Where the C , v_{c1} and v_{c2} are the capacitor value, the upper and lower capacitor voltages, respectively. Hence, the capacitor voltage differences can be derived as

$$v_{c1}(t) - v_{c2}(t) = \frac{1}{C} \int_0^t i_{ga} dt + v_{c1}(0) - v_{c2}(0) \quad (4.51)$$

This equation indicates that by controlling the phase "a" current, the deviation of the voltage capacitors could be controlled. The DC link voltage offset can introduce a DC component in the grid currents ($\bar{i}_{go}$). Hence, the three-phase grid currents are

$$i_{ga} = I_g \cos(\omega_s t) - \bar{i}_{go} \quad (4.52)$$

$$i_{gb} = I_g \cos(\omega_s t - \frac{2\pi}{3}) + \frac{1}{2} \bar{i}_{go} \quad (4.53)$$

$$i_{gc} = I_g \cos(\omega_s t + \frac{2\pi}{3}) + \frac{1}{2} \bar{i}_{go} \quad (4.54)$$

Where i_{ga} , i_{gb} , i_{gc} are the three phase GSC currents, I_g is the amplitude of the GSC AC current and $\bar{i}_{go}$ is the DC current in the grid current. By substituting (4.52) in (4.51), the capacitor voltage deviation can be obtained as

$$\begin{aligned}
v_{c1}(t) - v_{c2}(t) &= \frac{1}{C\omega_s} I_g \sin(\omega_s t + \alpha) - \frac{1}{C\omega_s} I_g \sin(\alpha) - \frac{\bar{i}_{go}}{C} t + v_{c1}(0) \\
&\quad - v_{c2}(0)
\end{aligned} \tag{4.55}$$

The first term of this equation is an AC term while the other terms are DC values. The DC current $\bar{i}_{go}$ should be controlled somehow to suppress the total DC terms. According to (4.52)- (4.54), the grid current in the stationary $\alpha\beta$ frame can be expressed as

$$i_{g\alpha} = I_g \cos(\omega_s t + \alpha) - \bar{i}_{go} \tag{4.56}$$

$$i_{g\beta} = I_g \sin(\omega_s t + \alpha) \tag{4.57}$$

If the grid current in β axis is divided by $C\omega_s$ and then subtracted from (26), the result can be obtained as

$$v_{c1}(t) - v_{c2}(t) - \frac{i_{g\beta}}{C\omega_s} = -\frac{1}{C\omega_s} I_g \sin(\alpha) - \frac{\bar{i}_{go}}{C} t + v_{c1}(0) - v_{c2}(0) \tag{4.58}$$

Consequently, the DC terms of the capacitor voltage differences can easily be estimated by (4.58). The DC component current reference ($\bar{i}_{go}^*$) that should be added to the grid current in order to remove the offset, could be obtained by using a PI controller. Hence, the compensation DC current in the stationary frame can be obtained as:

$$i_{g\alpha-c}^* = -k_p * (v_{c1}(t) - v_{c2}(t) - \frac{i_{g\beta}}{C\omega_s}) - k_i * \int (v_{c1}(t) - v_{c2}(t) - \frac{i_{g\beta}}{C\omega_s}) dt \tag{4.59}$$

$$i_{g\beta-c}^* = 0 \tag{4.60}$$

Where the k_p and K_i are the proportional and integral gains of the PI controller, respectively. The active and reactive power reference values for removing the offset can be expressed as

$$P_{g-c}^* = \frac{3}{2} u_{g\alpha} i_{g\alpha-c}^* \tag{4.61}$$

$$Q_{g-c}^* = \frac{3}{2} u_{g\beta} i_{g\alpha-c}^* \tag{4.62}$$

Where P_{g-c}^* and Q_{g-c}^* are the active and reactive power references for removing the offset. The calculation of the compensatory active and reactive power references for removing the DC offset is shown in Fig. 4.14.

Fig. 4.14. Block diagram of the compensatory power reference needed to eliminate the DC voltage offset.

4.3.2.6. Overall proposed predictive direct power control for FSTP GSC

The block diagram of the proposed PDPC method for the FSTP GSC is shown in Fig. 4.15. In the proposed strategy, at the beginning of every period, the three voltage vectors that were selected in the previous period are selected again. The active and reactive power slopes for these three vectors are calculated according to (4.41)-(4.42), as well as the grid active and reactive powers, grid voltage and capacitor voltages. Then, according to the active and reactive powers and their references, as well as the estimated power slopes, the duration times of the vectors are determined according to (4.46)-(4.48). If the duration times of all the selected vectors are positive, they will be generated by the converter. However, if the duration time of the second vector (t_b) is negative, the voltage vector sequences of the other sector will be selected, and duration times for these new vectors are calculated. In this case, the duration time of the new second vector (t_b) will be positive. When the duration time of the second vector becomes positive, the duration times of the first and third vectors should be checked, and if they are not positive, they should be corrected according to (4.49). Finally, the modified vectors according to their times will be applied. It should be noted that the active and reactive power references consist of two terms. One term is the compensatory terms for removing the DC link voltage offset that was shown in Fig. 4.14, and the other one is the usual active and reactive power references of the GSC.

Fig. 4.15. Block diagram of the proposed PDPC for FSTP GSC.

4.4. Summary and Conclusions

In this chapter a model predictive current control (MPCC) with capacitor voltage offset elimination control for a four-switch three phase converter (FSTPC) connected to SPMSG, that brings some benefits over the existing control methods used before, has been presented. The proposed MPCC for FSTPC MSC allows the system to follow the stator current references with great accuracy, whereas the switching frequency of the IGBTs is fixed and low, despite the fact that it is working with only four switches. This method minimizes the current reference tracking error, and its fast response makes it suitable for the OWC and wind turbine systems. Furthermore, a fast capacitor voltage offset suppression control was presented for the MSC and balanced three phase currents were achieved. The proposed capacitor voltage suppression control can remove the voltage offset in the capacitors rapidly and without needing to obtain the mean values of the capacitor voltages using a low-pass filter, which would be difficult at low speeds.

Moreover, a predictive power control for a four-switch three phase (FSTP) grid side converter (GSC) of DFIG/PMSG was presented in this chapter. In this method, three voltage vectors are employed to ensure that the proposed strategy maintains the switching frequency constant while the ripple of the active and reactive powers of the GSC is minimized according to the cost function. Moreover, a new powers compensation algorithm to eliminate the DC voltage deviation in the capacitors without changing the voltage vector sequences, and without using any low pass filters, has been presented.

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Simulation and Experimental Results for the Proposed Model Predictive Direct Control based on Four Voltage Vectors for The Power Converters of DFIG and PMSG

5.1. Introduction

In chapter 3, the proposed model predictive control for the RSC of DFIG, MSC of SPMSG and for the GSC of both DFIG and SPMSG was presented. The estimation of power slopes and current slopes to obtain the duration times was addressed and the algorithm to select the correct four voltage vectors in each switching period, two active vectors and two zero vectors, was reported.

In this chapter, the simulation and the experimental results for the control strategies that were presented in chapter 3, are presented and the results are discussed. Consequently, this chapter is divided into three parts, which are:

- 1- Simulation and experimental results of the proposed model predictive direct power control for the RSC of DFIG based on four voltage vectors during balanced and unbalanced grid voltage.
- 2- Simulation and experimental results of the Proposed Model Predictive Direct Current control based on four voltage vectors for the machine side converter of SPMSG.
- 3- Simulation and experimental results of the Proposed Model Predictive Direct Power Control for the grid side converter (GSC) of DFIG and SPMSG

In the first part, section 5.2, the simulation studies for the proposed method are carried out in Matlab Simulink environment, applied to a 2 MW DFIG under balanced and unbalanced grid voltage, in order to see the dynamic performance of the strategy. Later on, the results of the proposed strategy during balanced grid voltage are compared with those obtained with three other different control strategies. Moreover, the results of this technique under unbalanced grid voltage are compared to two other previous PDPC strategies, in order to show the superiority of the proposed method. During unbalanced grid voltage, the three techniques that can be adopted to overcome the problems of the negative sequence on the induction machine are reported. Finally, the dynamic stability of the proposed method for the RSC of DFIG is tested. In the experimental section, the proposed method is tested with a 5 kW DFIG during balanced grid voltage.

In the second part, section 5.3, the simulations are carried out to check the dynamic performance of the proposed method with a 8.7 kW SPMSG and are compared to two previous predictive controls for the MSC of SPMSG. Moreover, the robustness of the method during parameter variations is tested. Also, the performance of the control is analyzed in simulation using a simple model of an oscillating water column (OWC) power plant. In the experimental section, the dynamic performance and the stator current THD for a 8.7 kW SPMSG prototype are analyzed, and finally, the proposed method is evaluated in an OWC laboratory emulator.

In the third part of this chapter, section 5.3, the performance of the proposed method for the GSC during steady-state and transients is checked both in simulation and in experimental tests, and the results of the proposed method are compared with those from a previous predictive control.

5.2. Simulation and experimental results of the proposed model predictive direct power control for the RSC of DFIG based on four voltage vectors during balanced and unbalanced grid voltage

In order to solve the drawbacks related to model predictive controls during voltage disturbance, this section presents the results of the proposed improved PDPC for a DFIG rotor side converter (RSC) where four voltage vectors are applied during each switching period.

Fig. 5.1. Schematic of examined DFIG in simulation.

5.2.1. Simulation results

As mentioned earlier, the proposed novel improved predictive direct power control (PDPC) strategy for a DFIG RSC has been examined through the simulation of different cases in MATLAB/Simulink environment using a 2 MW DFIG model, under both balanced and unbalanced grid voltage conditions. The examined DFIG schematic is shown in Fig. 5.1 and parameters are listed in Table 5.1. In order to keep the DC link voltage at a constant value, the proposed model predictive control approach was used in the DFIG grid side converter. The DC link voltage and the DC link capacitor are 1200V and 16000 μ F, respectively. The switching frequencies of both converters were set to 4 kHz in the simulation studies. In order

to evaluate and investigate the proposed strategy more effectively, simulation results were carried out in the following five subsections.

Table 5.1. Parameters of simulated DFIG

Parameters of simulated DFIG			
Rated power	2 MW	L_m	3.362 pu
Rated voltage	690 V	DC link nominal voltage	1200 V
R_s	0.0108 pu	Number of pole pairs	2
R_r	0.0121 pu	Lumped inertia constant	0.5
L_{ls}	0.102 pu	Grid frequency	50 Hz
L_{lr}	0.11 pu	N_s/N_r	0.333

5.2.1.1. Operation under balanced grid voltage

In this study, the rotor speed has been increased linearly from 0.8 pu at $t=0.1$ s to 1.2 pu at $t=0.9$ s. The rotor speed variation for this study is shown in Fig. 5.2 (a). Stator active and reactive power references were set to be changed at some points to observe the transient response and dynamic performance during rotor speed changes. At the beginning of the simulation, stator active and reactive power references were set to -0.7 MW and 0 Mvar, respectively (negative values indicate that the DFIG is injecting power into the grid). The stator active power reference was changed from -0.7 MW to -1.5 MW at $t=0.2$ s and then was changed again from -1.5 MW to -1 MW at $t=0.5$ s. Finally, the stator active power reference was set to -1.5 MW at $t=0.9$ s. The stator reactive power reference was kept at -0.5 Mvar from $t=0.3$ s to $t=0.9$ s, and at $t=0.9$ s the stator reactive power reference was changed from -0.5 Mvar to 0 Mvar.

The performance of the proposed PDPC strategy in the mentioned study is shown in Fig. 5.2. According to Fig. 5.2(b), stator active and reactive powers perfectly follow the power references and have a fast response, without overshoots. Rotor and stator currents are shown in Figs. 5.2(c) and 5.2(d). As can be seen, rotor current frequency changes according to rotor speed since it depends on the generator slip.

Fig. 5.2. Simulation results of proposed PDPC during balanced grid voltage: (a) rotor speed (pu), (b) stator active (MW) and reactive power (Mvar), (c) three phase rotor currents (kA), (d) three phase stator currents (kA), (e) rotor voltage sector and (f) durations of first (t_b) and second (t_c) active vectors.

The appropriate rotor voltage sector for each switching period in this simulation is shown in Fig. 5.2 (e). The appropriate sector is the sector in which the corresponding two active vectors have positive duration times. Duration times of the two active vectors applied in each switching period are shown in Fig. 5.2(f). In steady state, the duration times of two active vectors when rotor speed is near synchronous speed decrease, as compared to other rotor speed values. As seen in Fig. 5.2(f), when changes occur in active or reactive power references, the estimated values of the duration times increase in order to produce a larger rotor voltage that allows reaching the new power reference

5.2.1.2. Comparison studies of the proposed control with previous strategies during balanced grid voltage

In order to evaluate the dynamic performance of the proposed predictive control strategy during balanced grid voltage, the proposed method has been compared to the classical vector control (VC) [1], the direct power control (DPC) [2] and a predictive direct power control (PDPC) [3], in which the vectors are selected according to the stator flux. The results of this comparison are shown in Fig. 5.3. In this figure, the stator of the DFIG was injecting -1.6 MW active power and 0 Mvar reactive power to the grid, in the beginning. At $t=0.9$ s, the active power reference is changed to -1 MW. The switching frequency for this study was set to 2 kHz for the proposed model predictive control, the PDPC [3] and the vector control [1]. The hysteresis bands for active and reactive power errors for the DPC method were selected to 0.04 pu and the maximum switching frequency for this method was 7 kHz. Moreover, the machine speed was controlled to be 1.2 pu. As can be seen in Fig. 5.3, the proposed predictive control, the PDPC [3] and the DPC [2] methods have a very fast dynamic response. However, as it was expected, the VC method [1], has a slow dynamic response compared to the other three strategies. The parameters of the PI controllers in the VC method were correctly adjusted to show a good and fast performance in this comparison.

Fig. 5.3. Comparison of stator active power Dynamic response for the proposed improved predictive control, PDPC [3], DPC [2], and VC [1].

The stator current harmonic spectrum before the change step in the active power is shown in Fig. 5.4. for these four control methods. The stator current THD for the proposed method is 1.39 % while for the PDPC [3], DPC [2] and VC [1] are 2.45%, 3.86% and 1.48%, respectively. As can be seen, the proposed method has a considerably better stator current THD compared to the previous PDPC [3] and DPC strategies [2].

Fig. 5.4. Comparison of harmonic spectrum of the stator current: (a) proposed model predictive strategy (b) predictive direct power control [3]; (c) Vector control [1]; (d) direct power control [2].

5.2.1.3. Operation under unbalanced grid voltage

The same power changes and rotor speed variations of section 5.2.1.1 have been used during unbalanced voltage in order to observe the behavior of the proposed PDPC and to compare it with normal voltage operation. The stator voltage applied for this case is shown in Fig. 5.5. As can be seen, a 20% voltage dip in phase “b” has been applied in the stator windings at $t = 0.1$ s.

Fig. 5.5. Stator voltage under unbalanced grid voltage.

In order to check the ability of applying different control targets using the proposed strategy, and to check the ability of injecting AC power terms, the following three control targets are chosen to be used in this simulation study:

- 1) Smooth stator active and reactive power: from $t = 0.1$ s to 0.4s.
- 2) Smooth electromagnetic torque and stator reactive power: from $t = 0.4$ s to 0.7s.
- 3) Injecting sinusoidal and balanced stator currents into the grid: from $t = 0.7$ s to 1s.

Fig. 5.6. Simulation results of proposed PDPC during unbalanced grid voltage with three different control targets: (a) rotor speed (PU), (b) stator active (MW) and reactive power (Mvar), (c) electromagnetic torque (kN.m), (d) three phase rotor currents (kA), (e) three phase stator currents (kA), (f) rotor voltage sector and (g) durations of first (t_b) and second (t_c) active vectors.

The results are shown in Fig. 5.6. As can be seen, the proposed PDPC can easily follow step changes in the power references and can inject oscillation terms of powers in order to have smooth electromagnetic torque or balanced stator currents during unbalanced grid voltage. The performance of the proposed control for each of the three control targets can be describe as follows:

Target 1: As can be seen in Fig. 5.6, during $t=0.1s$ to $t=0.4s$ the proposed strategy injects smooth active and reactive powers according to the power references which were set to DC values. In this control target, the torque is pulsating and stator and rotor currents are unbalanced and distorted. It should be noted that when the changes in the power references occur, active and reactive powers quickly reach the power reference keeping the control target unaltered.

Target 2: The results show that when references were set to have constant DC electromagnetic torque and reactive power, during $t=0.4s$ to $t=0.7s$, stator reactive power and torque values are smooth. In this control target, stator active power presents oscillations with twice the grid frequency in order to present a smooth electromagnetic torque and to decrease mechanical stress.

Target 3: Finally, the proposed PDPC can inject sinusoidal and balanced stator currents into the grid when the power references are set according to this target ($t=0.7s$ to $t=1s$). For clarity and to better show stator current with this control target, stator current near $t=0.9s$ is depicted in Fig. 5.7. As can be seen in this figure, stator currents are balanced and sinusoidal before and after the change in the power references at $t=0.9s$.

The appropriate rotor voltage sector and duration times of the active vectors in each switching period for these three control targets during unbalanced grid voltage, are shown in Figs. 5.6(f) and 5.6(g), respectively. By comparing Figs. 5.6(f) and 5.6(g) with Figs. 5.2(e) and 5.2(f), it can be observed that the appropriate rotor voltage sector (suitable active vectors) and duration times for the three control targets during balanced and unbalanced voltage are different.

Fig. 5.7. Stator current under unbalanced grid voltage around $t=0.9s$.

5.2.1.4. Dynamic stability

Dynamic stability could be tested when the power references are changed. In this case the powers should converge to the desired steady state power values. In subsection 5.2.1.3, during unbalanced voltage, at $t=0.9s$ active and reactive power references have been changed from -1 MW to -1.5 MW and from -0.5 Mvar to 0 Mvar, respectively. For this case, the absolute difference between the powers and their references near $t=0.9s$ are shown in Fig.

5.8. Figs. 5.8(a) and 5.8(b) show that active and reactive powers are converged to the neighborhood of their references even when the power references have AC components. Moreover, it can be seen that the active and reactive powers quickly reached their reference values.

Fig. 5.9 shows active and reactive power variations during unbalanced grid voltage and near $t=0.2s$ of Fig. 5.6. In this case, the active power reference has been changed from -0.7 MW to -1.5 MW, while the reactive power reference was constant at 0 Mvar. Fig. 5.9 clearly indicates that the active power has converged to the neighborhood of -1.5 MW, while stator reactive power was in the neighborhood of 0 Mvar.

Fig. 5.8. Absolute difference between the powers and their references: (a) for stator active power $|P_s - P_{s-ref}|$ (MW) and (b) for stator reactive power $|Q_s - Q_{s-ref}|$ (Mvar).

Fig. 5.9. Convergence of stator active power when active power reference has been changed from -0.7 MW to -1.5 MW.

5.2.1.5. Comparison studies under unbalanced grid voltage

In this subsection, the proposed strategy and two other PDPC methods [3, 4] are examined during a 20% voltage dip in phase “b” while rotor speed was fixed at 0.8 PU. Throughout all the simulation studies, stator active and reactive power references were set to -1.5 MW and 0 Mvar, respectively. The switching period time for these three strategies is set to 250 μ s. Moreover, the mentioned three control power targets in subsection 5.2.1.3 have been adopted to have faire comparison. At $t=0.4s$, the grid voltage has the mentioned negative sequence voltage. Target 1 was adopted from $t=0.4s$ to $t=0.5s$, and target 2 was chosen to be applied from $t= 0.5s$ to $t=0.6s$. Finally, target 3 was applied from $t= 0.6s$ to $t=0.7s$.

Fig. 5.10. Simulation results for the proposed PDPC of DFIG during unbalanced grid voltage with three different control targets at 0.8 PU rotor speed: (a) stator active power (MW), (b) stator reactive power (Mvar), (c) electromagnetic torque (kN.m), (d) three phase stator currents (kA) and (e) three phase rotor currents (kA).

The simulation results for the proposed PDPC are shown in Fig. 5.10. Results for PDPC based on three vectors [3] and PDPC based on one voltage vector for unbalanced voltage conditions [4] are shown in Figs. 5.11 and 5.12, respectively. As can be seen in Fig. 5.10, the proposed PDPC easily follows the power references of the three control targets while the switching frequency is constant. Moreover, the proposed PDPC has less ripple in stator active and reactive powers and low current THD compared to the other two control strategies. Fig. 5.11 shows that the suggested PDPC in [3] cannot work properly under unbalanced grid voltage, and power ripples are significantly high. Moreover, it indicates that applying the first active vector of the vector sequences according to rotor flux position when the duration time becomes negative [3] is not a good solution during unbalanced voltage.

Fig. 5.12 demonstrates that the suggested technique in [4] can obtain the three mentioned control targets while switching frequency is variable. However, the ripple of active and reactive powers and current THD in [4] is greater than that obtained with the proposed strategy presented in this thesis.

Active and reactive power ripples of the proposed PDPC and the two previous PDPC strategies for control target 1 are summarized in Table 5.2. This table shows that stator active and reactive powers of the PDPC based on three vectors [3] have very large ripples, especially in reactive power. In contrast, the proposed PDPC has small ripples compared to the two previous PDPC strategies. As a result, the proposed PDPC is more successful in obtaining smooth stator active and reactive powers.

The values of electromagnetic torque and reactive power ripples of the proposed PDPC and the two previous PDPC strategies for control target 2 are shown in Table 5.3. It can be seen that the PDPC based on one voltage vector [4] achieved better performance compared to ref. [3]. Nevertheless, the ripple of electromagnetic torque and reactive power of the proposed strategy is the lowest. This table clearly demonstrates that the proposed strategy could remove electromagnetic torque pulsations.

Finally, stator current THD and current unbalance value of the three control strategies for target 3 is summarized in Table 5.4. Stator current THD for the proposed strategy is 0.88%, while current THD for the PDPC based on [4] and [3] are 3.94% and 5.14%, respectively. Moreover, stator current of the proposed control is more balanced compared to the two previous strategies.

Fig. 5.11. Simulation results of PDPC based on three voltage vectors [3] during unbalanced grid voltage with three different control targets at 0.8 PU rotor speed: (a) stator active power (MW), (b) stator reactive power (Mvar), (c) electromagnetic torque (kN.m), (d) three phase stator currents (kA) and (e) three phase rotor currents (kA).

Fig. 5.12. Simulation results of PDPC based on selecting one vector during unbalanced grid voltage [4] with three different control targets at 0.8 PU rotor speed: (a) stator active power (MW), (b) stator reactive power (Mvar), (c) electromagnetic torque (kN.m), (d) three phase stator current (kA) and (e) three phase rotor current (kA).

Table 5.2. The stator active and reactive power ripples of the proposed PDPC and two previous PDPC strategies for the control target 1

Target 1	Proposed PDPC	PDPC based on three vectors [3]	PDPC based on one vector [4]
P_s ripple %	1.9	12.86	8.06
Q_s ripple %	2.1	30.27	6.54

Table 5.3. The electromagnetic torque and reactive power ripples of the proposed PDPC and two previous PDPC strategies for the control target 2

Target 2	Proposed PDPC	PDPC based on three vectors [3]	PDPC based on one vector [4]
T_e ripple %	2.03	12.04	7.2
Q_s ripple %	1.99	28.5	6.9

Table 5.4. The stator current THD and stator current unbalance value of the proposed PDPC and two previous PDPC strategies for the control target 3

Target 3	Proposed PDPC	PDPC based on three vectors [3]	PDPC based on one vector [4]
I_s THD %	0.88	5.14	3.94
I_s unbalance %	0.2	5.83	2.4

5.2.2. Experimental results

The performance of the proposed predictive direct power control for RSC of DFIG is evaluated in the laboratory for a 5 kW DFIG during balanced grid voltage. The schematic of the experimental setup is shown in Fig. 5.13. Two back to back Semikron SKS 35F B6U converters are used as RSC and GSC in this setup. The parameters of the experimental prototype are illustrated in Table 5.5.

Fig. 5.13. Schematic of the DFIG setup used for experimental tests.

Table 5.5. Parameters of Experimental DFIG setup

DC link voltage	250 V	Stator leakage inductance	8.6 mH
Rated power	5 kW	Rotor leakage inductance	8.6 mH
Grid voltage	400 V	Turns ratio	2.19
Rated stator voltage	400V	Inertia	0.16 kg.m2
Rated rotor voltage	182 V	Synchronous speed	1000 rpm
Number of pole pairs	3	C	4080 μ F
Stator resistance	0.65 Ω	L_{gs}, L_{fs}	20 mH
Rotor resistance	2.2 Ω	R_{gs}, R_{fs}	1 Ω
Mutual inductance	81 mH		

The test bench with the 5 kW DFIG laboratory prototype is shown in Fig. 5.14. The proposed PDPC strategy was implemented on a dSPACE 1104 to control the DFIG. Two dSPACE 1104 were used to control the GSC and RSC separately. The sampling period and the switching frequency for the GSC and RSC were set to 250 μ s and 4 kHz, respectively. Besides, a 3.3 μ s dead time was selected for the IGBT drivers in the dSPACE code as a software protection to prevent short circuits.

The experimental results were registered and recorded by the dSPACE control desk with a 1 ms sampling rate. Since the wound rotor induction machine was designed to work as a motor, the induction machine should absorb reactive power from the grid in order to avoid overcurrent in the rotor windings. In this case, a part of the magnetization current is supplied by the grid. In this bench, a DC motor has been connected to the DFIG to supply the mechanical torque of the shaft emulating a wind turbine.

Fig. 5.14. Experimental DFIG setup.

With the setup described above, three different tests have been performed in order to check the viability of the proposed controlled method and to validate some of the simulation studies described in section 5.2.1:

1. Firstly, the dynamic response of the proposed method during a change in the stator active power reference was checked. At the beginning of this test, the stator active and reactive powers were set to 1 kW and -1.5 kvar, respectively. The negative value of the reactive power indicates that the electrical machine is absorbing reactive power from the grid. At $t=0.5$ s, the stator active power reference was decreased from 1 kW to 0 kW. The rotor speed at the beginning was 900 rpm. However, after the change in the stator active power, the rotor speed was increased since the rotor speed control was not employed for the DC machine. The result of this test is shown in Fig. 5.15. As can be seen in this figure, the stator active power is following the reference value perfectly while the stator current amplitude is decreased due to the decrease in the stator active power. The rotor currents are sinusoidal and balanced and the rotor frequency is changed according to the rotor speed value.

Fig. 5.15. Dynamic performance of the proposed PDPC for the RSC of DFIG during a change in stator active power reference: (a) rotor speed (rpm); (b) stator voltage (V); (c) stator active and reactive power (kW, kvar) (d) stator current (A); (e) rotor current (A).

2. Secondly, the proposed control method is evaluated during reactive power changes. In this test, the rotor speed was set to 938 rpm and the active and reactive power were set to 0 kW and -1.5 kvar, respectively. At $t=0.2$ s, the reactive power reference was changed to -2 kvar. The result of this test is presented in Fig. 5.16. As can be seen in this figure, the reactive power follows the reference value correctly. Since the absorbed reactive power from the grid is increased in the stator, the reactive power that is needed to be supplied by the rotor is decreased. Hence, the rotor current value is decreased after the stator reactive power change. The rotor current frequency is almost constant due to the rotor speed value.
3. Finally, the PDPC is checked during a variation in the rotor speed. In this test, the rotor speed was increased from 800 rpm to 1075 rpm during 8 seconds, in order to see the performance of the strategy during a change from sub-synchronous to hyper-synchronous speed. The variation of the rotor speed is shown in Fig. 5.17(a). For this test, the stator active and reactive powers were set to be 1 kW and -1.5 kvar, respectively. The results are illustrated in Fig. 5. 17. During the variation of the rotor speed, the proposed strategy is following the active and reactive power references. As can be seen, the rotor current frequency is changed according to the generator slip, reaching a change in its sequence when the speed increases over the synchronous speed, coming from sub-synchronous operation.

The experimental result of this section shows that the proposed PDPC for the RSC is capable to control the DFIG and make it follow the power references at different circumstances.

Fig. 5.16. Dynamic performance of the proposed PDPC for the RSC of DFIG during a change in stator reactive power reference: (a) rotor speed (rpm); (b) stator voltage (V); (c) stator active and reactive power (kW, kvar) (d) stator current (A); (e) rotor current (A).

Fig. 5.17. Dynamic performance of the proposed PDPC for the RSC of DFIG during rotor speed changes: (a) rotor speed (rpm); (b) stator voltage (V); (c) stator active and reactive power (kW, kvar) (d) stator current (A); (e) rotor current (A).

5.3. Simulation and experimental results of the Proposed Model Predictive Direct Current control based on four voltage vectors for the machine side converter of SPMSG

5.3.1. Simulation results

The proposed model predictive direct current control (MPDCC) for SPMSG is evaluated using a model developed in Matlab/Simulink, and the results are presented throughout this section.

Firstly, in section 5.3.1.1, the dynamic performance and current THD of the proposed MPDCC method are checked and compared with two other methods, conventional one voltage vector MPC [5], and two voltage vectors MPC [9].

Secondly, in section 5.3.1.2, the performance of the proposed control is assessed in the OWC power plant emulator using a real wave profile. The emulator of OWC power plant was described in chapter 3. This emulator is simulating the experimental OWC test bench that will be used in section 5.3.2.3 to evaluate the operation of the proposed control. The machine parameters of the simulated system and the laboratory setup are displayed in Table 5.6. The selected switching frequency and the sampling frequency were both 4 kHz.

Table 5.6. SPMSG parameters

DC Link voltage	600 V
Rated power	8.7 kW
Rated voltage	400 V
Number of pole pairs	3
Permanent magnet Flux	1.05 Wb
Stator Inductance	50 mH
Stator resistance	2 Ω

5.3.1.1. Dynamic performance of the proposed MPDCC and comparison with two MPC controls

In this part, the dynamic performance of the proposed MPDCC is presented and is compared with conventional one voltage vector MPC and two voltage vectors MPC methods.

In this dynamic performance study, the generator speed was accelerated from 0 at $t=1$ s to the rated speed in 0.5 seconds, for all the methods. The references of the stator currents in d and q axis were set to 0A and -8A respectively, at the beginning of the simulation. However, at $t=1.2$ s i_{sq}^* was set to -4A, and at $t=1.3$ s i_{sd}^* was changed to 1A.

The simulation results of this study for the proposed MPDCC, for the conventional one voltage vector MPC and for the two voltage vectors MPC methods, are shown in Fig. 5.18, Fig. 5.19 and Fig. 5.20, respectively.

Fig. 5.18. Simulation results for the Proposed MPDCC for SPMSG: (a) generator speed; (b) stator current in the d-q frame; (c) three-phase stator current; (d) electromagnetic torque; (e) stator flux; (f) active vector durations.

All the predictive controls follow the current references correctly and have a fast dynamic response to the sudden changes in the references. However, it is very clear in these figures that the proposed MPDCC has a lower ripple in d-q currents, as well as in the torque and in the stator flux. Moreover, the stator current for the proposed MPDCC looks “cleaner” and has a better sinusoidal shape. In the proposed MPDCC, the duration times of the active vectors are increased when the generator speed is high, as can be seen in Fig. 5.18(f). This is because, at low speeds, the zero vectors have small current slopes. Therefore, in most portions of the period, these vectors are employed in order to minimize the current ripples.

The duration of the active vectors for the proposed control are illustrated in Fig. 5.18(f). It can be observed that the duration times of all the active vectors are positive, which demonstrates that the proposed sector identification algorithm works very well and can find the correct voltage vector sequences quickly.

Among the presented MPC strategies, the conventional one voltage vector MPC has higher ripple in all the variables. The performance of the two-voltage vector MPC is better than the conventional one voltage vector MPC. The torque, stator flux and current ripples in the two voltage vectors MPC are lower than in the one voltage vector MPC. The performance of the two-voltage vector MPC at low speeds is good and it is similar to the proposed MPDCC since a zero vector is applied in each period. However, during higher speeds, the current ripples of the two-voltage vector MPC is increased significantly while the current ripple for the proposed control is lower.

Fig. 5.19. Simulation results for the conventional one voltage MPC for SPMSG: (a) generator speed; (b) stator current in the d-q frame; (c) three-phase stator current; (d) electromagnetic torque; (e) stator flux.

Fig. 5.20. Simulation results for the two voltage vectors MPC for SPMSG: (a) generator speed; (b) stator current in the d-q frame; (c) three-phase stator current; (d) electromagnetic torque; (e) Stator flux.

In order to compare the harmonic spectrum of the proposed MPDCC with the conventional one voltage vector and two voltage vectors MPC control, FFT analyses were carried out at $t=1.62$ s for one cycle. One of the current phases and the harmonic spectrum for the proposed MPDCC, conventional one voltage vector MPC and two voltage vectors MPC controls are presented in Fig. 5.21, Fig. 5.22 and Fig. 5.23, respectively. The total harmonic distortion (THD) of the stator current for the proposed MPDCC is 2.63%, which is much lower than the conventional one voltage vector and two voltage vectors MPC with THD values of 16.06% and 10.23%, respectively.

The results for the conventional MPC show that the stator current contains high low-order harmonics, in consequence, they are difficult to remove by means of filters. In two voltage vectors MPC the low-order harmonics are decreased compared to the conventional MPC. However, still these harmonics are notable. In contrast, for the proposed MPDCC, the dominant harmonics are located around the 4 kHz and 8 kHz and so on.

Fig. 5.21. Harmonic spectrum of proposed MPDCC for SPMSG in simulation at $t=1.62$ s: (a) phase "a" current; (b) Harmonic spectrum.

Fig. 5.22. Harmonic spectrum of conventional one voltage vector MPC for SPMSG in simulation at $t=1.62$ s: (a) Phase "a" current; (b) Harmonic spectrum.

Fig. 5.23. Harmonic spectrum of two vectors MPC for SPMSG in simulation at $t=1.62$ s: (a) Phase "a" current; (b) Harmonic spectrum.

However, even using the same control period for the three cases, the switching frequency of the one voltage vector and the two-voltage vector MPCs are both variable and their average switching frequencies are lower than that of the proposed MPDCC.

5.3.1.2. Robustness to parameter variations

In order to check the robustness of the proposed MPDCC method to parameter variations, it has been assumed, as an example, that the stator inductance is 30% lower than its real value. The same study conditions of Fig. 5.18 were set to check the change in the performance due to the inductance variation. The resulting stator current in dq frame is shown in Fig. 5.24. As can be seen, the proposed method makes the current dq components to follow the references accurately. Moreover, an FFT analysis was carried out at $t=1.62$ s to check the stator current THD and to compare it with that of Fig. 5.21. The stator current THD for this case is shown in Fig. 5.25. As can be seen, the current THD value for this case is almost the same as when the real value is used.

Fig. 5.24. Stator current in dq frame for the proposed MPC control with 30% of stator inductance error.

Fig. 5.25. Harmonic spectrum of proposed MPDCC for SPMSG during the 30% stator inductance error in simulation at $t=1.62$ s: (a) phase “a” current; (b) Harmonic spectrum.

5.3.1.3. Performance of the proposed MPDCC in an OWC power plant

The proposed MPDCC has been also tested in a simulation model of an OWC power plant emulator. The test was done for the irregular wave profile shown in Fig. 3.15 which lasts 45 seconds. In this test i_{sd}^* was set to zero. The results are illustrated in Fig. 5.26. The Wells turbine maximum efficiency speed and the actual speed during the 45 seconds test are shown in Fig. 5.26(a) with blue and red lines, respectively. As can be seen, with the proposed control, the generator follows the maximum efficiency speed perfectly during the test. The d

axis, q axis and three-phase stator current are shown in Fig. 5.26(b), Fig. 5.26(c) and Fig. 5.26(d), respectively. The stator current in q axis varies in order to make the Wells turbine to work at the maximum efficiency point.

(a)

(b)

(c)

(d)

Fig. 5.26. Simulation results for 45 seconds of OWC test: (a) Rotor speed (speed reference (blue line) and actual speed (red line)) (b) stator current in d axis (c) stator current in q axis (d) stator current in abc frame.

A more detailed view of this test during $t=10$ s to $t=12$ s is presented in Fig. 5.27. It can be seen that the proposed control strategy follows the maximum efficiency reference speed with good accuracy, and the stator currents are sinusoidal and balanced during this test.

Fig. 5.27. Detailed view of the simulation results of the OWC during $t=10$ s to $t=12$ s: (a) Rotor speed (speed reference (dash blue line) and actual speed (red line)) (b) stator current in d axis (c) stator current in q axis (d) stator current in abc frame.

5.3.2. Experimental results of the proposed MPDCC

As aforementioned in chapter 2, WECs (especially OWC power plants) are subjected to variations of torque and speed generated by the incoming and outgoing waves. Depending on the sea state, these variations can be sudden and large in amplitude and persistent over time when the sea is rough, or very low when the sea is calm. For this reason, a maximum efficiency-based control of the WEC should provide a precise torque control over a wide range of speed and torque.

Firstly, the laboratory prototype is described. Secondly, the dynamic performance of the proposed MPDCC is evaluated in a laboratory experimental setup during different transient tests and later the results obtained from several experimental tests carried out in the laboratory using a SPMSG in a WEC emulator are presented. The objective of testing the proposed control in a WEC emulator was to verify the ability of the proposed MPDCC to satisfy the high torque requirements of an OWC based power plant.

5.3.2.1. Description of the laboratory prototype

The prototype of the power plant is divided into two different systems: the “OWC device” and the “Electric power generation system”. Fig. 5.28 shows the layout of both parts, along with an external electric power system to which the prototype is connected, and that absorbs the generated electrical power. A photo of the experimental prototype is shown in Fig. 5.29.

Fig. 5.28. Schematic of the test bench used to carry out the experimental tests.

Fig. 5.29. Overview of the experimental test bench.

5.3.2.1.1. OWC device

The prototype part that simulates the OWC device, system 1 in Fig. 5.28, reproduces: the irregular wave, the chamber containing the oscillating water column, and the Wells turbine.

From the irregular wave profile, the control program calculates, in each cycle, the corresponding turbine torque by means of the chamber and the Wells turbine models. Then, this torque is reproduced in real time using a separately excited DC motor. The sea state used in the OWC emulator was characterized by a significant wave height $H_s = 1m$ and an energy period $T_e = 10s$ and the wave profile was generated using the Pierson-Moskowitz model [7],[8].

5.3.2.1.2. PTO and grid connection

The second part of the prototype, system 2 and system 3 in Fig. 5.28, reproduces the entire electrical system of the OWC power plant and the grid connection using real elements. As Fig. 5.29 shows, a SPMSG connected to the DC motor shaft (Wells turbine shaft) and a three-phase converter represent the PTO, which is controlled by the proposed MPDCC, while the grid connection is carried out using a vector-controlled three phase converter.

The dual core microcontroller used to program the MPDCC code was a F28M35x Concerto of Texas Instruments with a switching frequency of 4kHz and a sampling time of 250 μs . All of the waveforms were recorded using the F28M35x itself, except the stator currents, that were recorded using a Tektronix TDS2024C oscilloscope with a sampling time of 100 μs . The computation burden for the proposed control in F28M35x is 122 μs , while for the conventional one voltage vector MPC and two voltage vectors MPC is 95.5 μs and 100 μs , respectively. This indicates that the calculation burden of the proposed method is not high if compared to the two other MPC methods.

5.3.2.2. Dynamic performance of the proposed control

First of all, the electric current waveform quality was checked for several torque levels. Fig. 5.30 shows the generator phase currents corresponding to two different torque levels while rotor speed is around 205 rpm. In both, the waveform is almost ripple-free and is balanced.

Fig. 5.30. Stator currents when i_{sq} reference was set to -4A (top) and to -8A (bottom) whereas the i_{sd} reference was set to zero in both cases.

The harmonic spectrum corresponding to the stator current of Fig. 5.30 bottom, is shown in Fig. 5.31. The resulting THD calculated up to 2 kHz is 4.16% which is reasonably low.

Fig. 5.31. Stator current THD. The fundamental frequency is 10.22Hz and its magnitude is 7.81.

Secondly, the proposed control was assessed when the rotor speed increases. In this case, the rotor speed was increased from 250 rpm to 750 rpm and the d and q axis stator current references were set to 0A and -4A, respectively. The three-phase stator current and the rotor speed are displayed in Fig. 5.32. The stator current of this test was acquired (registered) by the DSP every 2 ms. The stator current frequency increased and followed the references when the rotor speed was increased.

Fig. 5.32. Stator currents and rotor speed during a rotor speed increment test.

Thirdly, the performance of the proposed control was tested using a reference torque step and a reference field step. It must be taken into account that the usual dynamic of a WEC might change within the range between several tenths of seconds to several seconds, not in the order of milliseconds. Fig. 5.33 shows the fast response of the actual i_{sq} following its reference during a step, from $-4A$ to $-6A$, while the SPMSG generates active power. The order of the time response indicates that this control is suitable to cope with the WECs dynamics. As can be seen, the stator current is sinusoidal and balanced during the test.

Fig. 5.33. Torque control of the SPMSG through i_{sq} . Step response when reference changes from $-4A$ to $-6A$. Note how fast the actual i_{sq} tracks the new reference after the step. The i_{sd} reference was set to zero.

The following test checked the proposed predictive control ability to control the magnetic field in the electric generator, what is of paramount importance to obtain a good torque control.

Although the magnetic field has a slow dynamic response, Fig. 5.34 shows how the MPDCC is able to get the new i_{sd} reference in a very short time. In this case, a field weakening was carried out, taking into account not to reach the irreversible demagnetizing point. The magnetic field weakening would allow to increase the rotary speed of the generator surpassing the rated speed in the event of a sudden large wave.

Fig. 5.34. Magnetic field control of the SPMSG through i_{sd} . Step response of the actual i_{sd} when the reference changes for 0A to -1.5A, carrying out a magnetic field weakening. i_{sq} reference was set to -4 A.

5.3.2.3. OWC power plant performance

Once the dynamic performances of the MPDCC were checked, the next step was to implement the control in an OWC-based power plant. In this control, the generator shaft speed is taken to the maximum efficiency point through the torque control.

As the incoming and outgoing waves push the air flow through the chamber, it makes the turbine and generator to rotate with changing speed. Fig. 5.35 shows the magnetic field angle and the electromotive force angle registered in the generator from the starting point when the irregular wave profile is used with the OWC emulator. Note the 90° diphas existing between them and the large rotary speed variations.

Fig. 5.35. Magnetic field angle (blue) and electromotive force angle (red) at the starting point of the test.

The wide range of speed variation makes the control of the generator more difficult. On the one hand, an accurate speed and shaft position measurement over the whole range, even close to zero rpm, becomes complex. So, the MPDCC should be robust to inaccuracies. Also, close to zero speed implies a negligible electromotive force in the generator, what makes the stator current change very fast as the voltage pulses are applied by the VSC. However, as the tests below will show, the robustness and fast dynamic of the proposed MPDCC make the control handle these problems properly.

In the following tests, an irregular wave profile was used in the OWC emulator. This irregular wave leads to a changing maximum efficiency speed which must be tracked by the torque control whereas the field control maintains the magnetic field constant.

Fig. 5.36 shows the reference speed needed to get the maximum efficiency (blue) and the actual shaft speed (red) along the test that obtained in a 4 minutes long test, where the rotary speed follows the speed reference very accurately along the whole test. Note the large speed variations and the irregular speed profile within this test.

Fig. 5.36. Rotary speed reference (blue) vs. actual speed (red) during a 4 minutes long test.

Fig. 5.37 shows a more detailed view of reference and actual speeds from $t=20$ s to $t=80$ s. As can be seen the rotor speed is following the reference value with a small error in some points.

Fig. 5.37. Detailed view of the rotary speed reference (blue) and actual rotary speed (red) during the test.

The actual torque is a fast-changing magnitude, so the torque control has to follow a fast reference given by the q component of the stator current. Fig. 5.38 shows the large changes appearing in the i_{sq} reference and how the actual i_{sq} (red) follows the reference (blue) very accurately meaning that the PTO works approximately at the maximum efficiency at any time.

Fig. 5.38. The torque control generates fast changing reference for i_{sq} while for i_{sd} keeps constant. A detailed view, sampled by the DSP each 0.00250 s, shows the actual i_{sq} and i_{sd} following their references accurately at every moment.

5.4. Simulation and experimental results of the Proposed Model Predictive Direct Power Control for the grid side converter (GSC) of DFIG and SPMSG

5.4.1. Simulation results

In this part, the simulation results for the GSC are presented.

Firstly, in section 5.4.1.1, the proposed strategy for the GSC is tested using Simulink/Matlab with a 2 MW DFIG model, during sub synchronous and hyper-synchronous speeds, in order to see both the invert and rectifier operation of the GSC. The parameters and the schematic of this DFIG is shown in Fig. 5.1 and Table. 5.1 of section 5.2, respectively. The switching frequency and the sampling frequency used was 4 kHz.

Later, in section 5.4.1.2, in order to see the performance of the proposed model predictive control for the GSC of SPMSG, the simulation studies have been carried out for a small size 5 kW three phase voltage source inverter. The parameters and the schematic of this three phase voltage source inverter are shown in Table 5.7 and Fig. 5.39, respectively. This system is simulating the laboratory workbench characteristics that will be employed later, to perform the experimental tests. The switching frequency and the sampling frequency used in this study are both 4 kHz.

Finally, in section 5.4.1.3, the operation of the proposed predictive control method is compared with two other previous controls.

Table 5.7. Parameters of the voltage source inverter

R_f	0.5Ω
L_f	20 mH
e_g (Grid voltage)	127V
V_{dc}	500 V
Grid frequency	50 Hz

Fig. 5.39. Schematic of the three phase voltage source inverter used for both simulation and experimental studies

5.4.1.1. Simulation results for the GSC of DFIG

The performance of the proposed model predictive control for the GSC of DFIG is evaluated in Matlab Simulink during the variation of the rotor speed.

In this case study, the generator was injecting 1.5 MW active power and 0 Mvar reactive power to the grid from the stator windings. At the beginning of this study, the rotor speed was 0.8 PU. At $t = 0.1$ s, the rotor speed was linearly increased and reached 1.2 pu at $t = 0.9$ s. After that, the rotor speed was fixed at 1.2 pu. The variation of the rotor speed is shown in Fig. 5.40(a). Moreover, the GSC reactive power reference is changed from 0 Mvar to 0.1 Mvar at $t = 1.1$ s. A positive value of power indicates that the power is being injected to the grid in this case. The DC link voltage reference was set to 1200 V for this study. The results are shown in Fig. 5.40.

As can be seen in Fig. 5.40, the GSC is absorbing active power from the grid at the beginning. The absorbed GSC active power decreases when the rotor speed increases. At the synchronous speed, the absorbed GSC active power decreases to zero, as can be seen in Fig. 5.40(b). After $t = 0.5$ s, the GSC is injecting active power to the grid. As can be seen in Fig. 5.40(d), the control is capable to maintain the DC link voltage at 1200V. Moreover, the control follows the GSC reactive power reference correctly during the study and, at $t = 1.1$ s when the reactive power reference is changed, the dynamic response to the step change is fast and without any error. The predicted GSC voltage sector and the estimated duration times of the active vectors are shown in Fig. 5.40(e) and Fig. 5.40(f), respectively. As can be seen, the duration times of the active vectors at every switching period is positive showing that the proposed method for the GSC can work perfectly during sub-synchronous and hyper-synchronous speeds (during inverter and rectifier operation of the GSC).

Fig. 5.40. Simulation results of the proposed model predictive control for the GSC of the 2 MW DFIG: (a) rotor speed (PU); (b) GSC current (A); (c) Reactive power (Mvar); (d) DC link voltage (V); (e) predicted GSC voltage sector; (f) durations of first (t_b) and second active vectors (t_c).

5.4.1.2. Simulation results for the GSC of SPMSG

In this part, the simulation studies for a 5 kW three-phase voltage source converter are carried out. In this study, the dynamic performance of the proposed model predictive control during changes in the GSC active and reactive power references is evaluated, always with the converter operating as inverter. In this case study, in the beginning the converter is injecting 1 kW and 0 Mvar reactive power to the grid. At $t=0.3\text{s}$ and $t=0.4\text{s}$, the GSC active and reactive power references are changed to 2 kW and -1 kvar (absorb reactive power) respectively.

Fig. 5.41. Simulation results of the proposed model predictive control for the GSC (three phase voltage source inverter): (a) GSC Active and reactive power (kW, kvar); (b) GSC current (A); (c) predicted GSC voltage sector; (d) durations of first (t_b) and second active vectors (t_c).

The simulation results for this test are presented in Fig. 5.41. As can be seen in Fig. 5.41(a), the active and reactive powers follow the power references correctly and very fast when the power references are changed. Moreover, the GSC current during this test is sinusoidal and balanced according to Fig 5.41(b). The duration times of the selected active vectors are positive in the whole simulation which indicates that the predicted GSC sectors and the selected vectors are correct. In Fig. 5. 41(d), at $t=0.3s$ and $t=0.4s$, the duration times of the active vectors are increased to build a bigger GSC voltage, needed to reach the new power references quickly.

5.4.1.3. Comparative study

The proposed model predictive control is compared with two previous model predictive controls which are the one voltage vector [9] and the two voltage vectors [10]-[11] model predictive controls. In order to have a fair comparison, the same sampling frequency and switching period were selected for the proposed predictive control and the two other previous predictive controls. In this comparison, the GSC is injecting 2 kW active power to the grid and absorbing -1 kvar reactive power from the grid. The GSC current and the current THD for the proposed method, one voltage vector MPC, and two voltage vector MPC are shown in Fig. 5.42(a), Fig. 5.42(b) and Fig. 5.42(c), respectively. As can be seen in these figures, the current of the proposed method is more sinusoidal than that in the other two previous model predictive controls. Moreover, the current THD for the proposed method and one voltage vector and two voltage MPC are 2.05 %, 13.32%, and 7.53 %, respectively. The proposed method has lower values for the low order harmonics while for the two other predictive controls, the low order harmonics are very high compared to the proposed strategy. This study indicates that the proposed method has superior current THD compared to the two previous model predictive controls.

Fig. 5.42. The GSC current and the harmonic spectrum for: (a) proposed model predictive control; (b) one voltage vector MPC [9]; (c) two-voltage vectors MPC [10].

5.4.2. Experimental results

In this part, the experimental results of the proposed model predictive control for the GSC of DFIG and SPMSG are presented.

Firstly, section 5.4.2.1, the result of the GSC of the DFIG is presented. The schematic and the photo of the DFIG laboratory prototype are shown in Fig. 5.13 and Fig. 5.14, respectively. The parameters of the DFIG setup are presented in Table 5.5. The switching frequency and the sampling frequency of the GSC of DFIG are both 4 kHz and the proposed control for the GSC has been implemented in a dSPACE 1104. The results are registered every 1 ms. Some of the tests that have been presented in section 5.2.2, Experimental results for the RSC of DFIG, have been selected to see the performance of the GSC of DFIG during the generator operation.

Later, in section 5.4.2.1, the proposed method is validated using a three phase voltage source inverter as the GSC of SPMSG in order to see the control performance when the converter is operating as an inverter. The proposed strategy was implemented in a double-core F28M35H52C Concerto DSP where the sampling frequency and the switching frequency were set to 7.5 kHz for these laboratory tests. The currents were registered using Fluke i30s AC/DC current clamps and a Tektronix TDS2024C oscilloscope. However, the active and reactive powers were calculated by the microcontroller and were stored in the RAM memory. The three-phase voltage source inverter laboratory setup, which is used to model the GSC of SMPSG, is shown in Fig. 5.43.

Fig. 5.43. The three-phase voltage source inverter laboratory setup.

5.4.2.1. Experimental results for the GSC of DFIG

First of all, the performance of the proposed MPC control for the GSC is evaluated during the same stator active power change that was presented in section 5.2.2, Fig. 5.15. In this test, the stator active power was reduced from 1 kW to 0 kW while the stator reactive power was set to be -1.5 kvar as can be seen in Fig. 5.15. The GSC reactive power reference was set to 0 kvar during this test.

Fig. 5.44. Experimental result of the proposed method for the GSC of DFIG during the same stator active power reference change of Fig. 5.15: (a) transformer secondary voltage; (b) DC link voltage; (c) GSC reactive power (kvar); (d) GSC current.

The GSC performance during this test is presented in Fig. 5.44. The transformer secondary voltage, to which the converter is connected, is shown in Fig. 5.44(a). The GSC current decreases after the reduction in the stator active power. The proposed method is capable of maintain the DC link voltage at 250 V. The DC link voltage increased a little bit after the stator active power change due to the slow dynamic PI controller. Moreover, the reactive power follows the GSC reactive power reference correctly without any transients during the test.

Secondly, the operation of the proposed control for the GSC is validated during the same rotor speed variation of Fig. 5.17. The power generation commands during this test , were set to 1kW and -1.5 kvar for the stator active and reactive powers respectively, along the whole variation of the rotor speed while the GSC reactive power reference was set to 0 kvar. The GSC results for this test are shown in Fig. 5.45. The proposed method maintained the DC link voltage at the reference value perfectly during the test. Additionally, the GSC reactive power was constant during the test, as can be seen in Fig. 5.45(b). The GSC current was decreased when the rotor speed is increased. It should be noted that the power losses for the laboratory induction machine are notable and the power losses are supplied from the rotor side. As a result, in this figure, the GSC is absorbing active power from the grid even during the synchronous speed.

Fig. 5.45. Experimental result of the proposed method for the GSC of DFIG during rotor speed variations of Fig. 5.17: (a) DC link voltage; (b) GSC reactive power; (c) GSC current.

Finally, the dynamic performance of the proposed method was evaluated by changing the GSC reactive power reference. The rotor speed was fixed at 900 rpm and the DC link voltage was set to 250 V while the stator was injecting 0 kW active power and absorbing -1.5 kvar reactive power. The GSC reactive power was changed from -200 var (GSC absorbs reactive power from the grid) to -300 var at $t=0.8$ s.

The results are shown in Fig. 5.46. As can be seen, the proposed model predictive control followed the reactive power reference accurately while the DC link voltage remained constant and both the current phase and amplitude are changed according to the new reactive power reference.

Fig. 5.46. Experimental result of the proposed method for the GSC of DFIG during GSC reactive power reference change: (a) GSC reactive power; (b) DC link voltage; (c) GSC current.

5.4.2.2. Experimental results for the GSC of SMPSG

The first test performed to check the steady-state performance of the active and reactive powers and the quality of the converter currents generated by the GSC using the proposed model predictive control. In this test, the active and reactive powers references were set to 1 kW and -1 kvar, respectively.

The steady-state results for this test are presented in Fig. 5.47. The active and reactive power ripples are low as can be seen in Fig. 5.47. Moreover, the three phase GSC currents

are sinusoidal and balanced. The Fast Fourier Transform corresponding to phase “a” of Fig. 5.47 is illustrated in Fig. 5.48. This figure shows the fundamental harmonic, which is 6.2 A, and other low order harmonics which are comparatively low, indicating that the GSC has a low current THD. The harmonic content obtained in the laboratory conditions shows that the 2nd, 3rd, and 4th order harmonics have a value of 0.03055 A, 0.08815 A, 0.0195A, respectively. The GSC current THD is estimated for this current and is 3.2%, which indicates the good quality of the current. It should be taken into account that, in the laboratory, the grid is provided by an autotransformer whose voltage waveform is not perfectly sinusoidal, and the filter inductances are not totally equal in the three phases.

Fig. 5.47. Steady-state performance of the proposed model predictive control for the GSC when is injecting 1 kW active power and absorbing -1kvar: (a) active and reactive power; (b) three phase GSC current.

Fig. 5.48. Harmonic spectrum of the GSC current when is injecting kW active power and absorbing -1kvar

The following tests show the dynamic response provided by the proposed model predictive control when the power references undergo fast changes.

In the first test, the active power reference is changed following a step, from 1 kW to 2 kW, whereas the reactive power reference is kept to 0 kvar. The result of this test, shown in Fig. 5.49, has a very fast dynamic response to the GSC active power reference change with a good reference tracking. The three phase GSC currents during this active power change are illustrated in Fig. 5.49. This figure also shows the fast dynamic response to the change of active power reference and, as can be seen, the currents changed from one steady-state to the other very quickly. Moreover, the currents are balanced and sinusoidal before and after the transient.

Fig. 5.49. Performance of the GSC during the GSC active power reference change from 1 kW to 2 kW while the reactive power reference is constant: (a) active and reactive power; (b) three phase GSC current.

The following test assesses the reactive power transient response during a change in the GSC reactive power reference. The results of this test are shown in Fig. 5.50(a) and Fig. 5.50(b). The first one shows the fast change experimented by the reactive power when the reference changes in a step from -0.5 kvar to -1.5 kvar, while the active power reference is maintained constant at 0 kW. Fig. 5.50(b) shows the GSC current during this reactive power change. In this test the GSC currents also reach the reference values very quickly and the currents are sinusoidal and balanced during the test.

Fig. 5. 50. Performance of the GSC during the GSC reactive power reference change from -0.5 kvar to -1.5 kvar while the active power reference is constant: (a) active and reactive power; (b) three phase GSC current.

5.5. Summary and Conclusions

In this chapter, simulation and the laboratory experimental studies for the proposed MPC strategies applied to the RSC of DFIG, MSC of SPMSG and GSC were carried out.

The dynamic result for the RSC of DFIG showed that under balanced grid voltage, the proposed MPDPC is capable to follow the power references during sub-synchronous and hyper-synchronous speeds and has fast dynamic response to the power reference changes. Additionally, the proposed MPDPC was compared with a vector control, a direct power control and another predictive direct power control based on three vectors [3], all working under balanced grid voltage. This comparison indicated that the proposed method has a fast dynamic response and lower stator current THD compared to the other mentioned methods. Furthermore, the stability of the proposed MPDPC was checked and the results showed that the system is stable.

In addition, the proposed MPDPC was tested during unbalanced grid voltage. Results showed that three control targets can be achieved easily during unbalanced grid voltage, without extracting negative sequence, or using resonance controllers, or negative synchronous frame. Besides, the results of the proposed method during unbalanced grid voltage were compared to other MPC controls (one voltage vector MPC and three voltage vectors MPC). This comparison revealed that the proposed MPDPC can handle the grid

disturbances while satisfying three different control targets with low power ripples, while the other two strategies cannot work perfectly and present higher ripples.

In section 5.3, the proposed MPDCC for the MSC of SPMSG was assessed by simulation and experimental laboratory tests. The dynamic performance was evaluated and compared to two previous MPC controls. The results showed that the proposed method follows the current references very quickly without any errors and the current ripples are lower than those in the other MPC controls. The comparison showed that the stator current THD for the proposed MPDCC was 2.63%, while for the conventional one voltage vector and two voltage vectors MPC, this value was 16.06% and 10.23%, respectively. Moreover, the proposed MPDCC was tested during a variation of the machine parameters and the results indicated that it is robust to the parameter variations. Finally, the performance of the MPDCC was tested in an OWC power plant system. The results imply that the proposed method is very suitable for controlling the MSC of OWC systems.

Finally, in section 5.4, the performance of the proposed MPDPC for the GSC of DFIG and SPMSG was evaluated through simulation and experimental tests. The results for the GSC of DFIG showed that the proposed method can operate perfectly during the sub-synchronous and hyper-synchronous speeds and the system can follow the reference values correctly. Additionally, the performance of the GSC of SPMSG was tested in transients as well as in steady-state. The outcome of the tests shows that the proposed method has a fast dynamic response with low GSC current THD. Moreover, the current THD of the proposed MPDPC is compared with one voltage vector MPC and two voltage vector MPC methods in the same operating conditions. The GSC current THD for the proposed method, one voltage vector MPC and two voltage vectors MPC was 2.05%, 13.32%, and 7.53%, respectively. This outcome proved the superiority of the proposed MPDPC compared to two other previous MPC methods.

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Simulation and experimental results for the Switch Fault Tolerance capability of the model predictive control for the Power converters of DFIG and SPMSG

6.1. Introduction

The proposed model predictive control for the four-switch three phase (FSTP) machine side converter (MSC) of the SPMSG and for the FSTP grid side converter (GSC) was presented in chapter 4. The idea of using three voltage vectors among the four possible voltage vectors of the FSTP converter (FSTPC) for the proposed control strategy was explained for these two power converters and the capacitor offset control was presented in chapter 4.

In this chapter, the simulation and the experimental results for the proposed methods that were explained in chapter 4, are carried out and the results are discussed. As a result, this chapter is divided into two main parts, which are:

- 4- Simulation and experimental results of Model Predictive Current Control (MPCC) for the machine side converter of SPMSG when fed by four-switch three phase converter.
- 5- Simulation and experimental results of Model Predictive Direct Power Control for Four-Switch Three Phase Operation in the Grid Side Converter.

This chapter is organized as follows:

- In section 6.2, the simulation results of the proposed model predictive control (MPC) for the FSTP MSC of SPMSG are shown. The performance of the capacitors offset control is also evaluated in this part. Later, the dynamic performance of the proposed control is checked in the Simulink environment. Afterwards, the proposed method is compared with previous MPC strategies. In the experimental studies, the steady-state and the transient performance of the proposed method for the MSC will be presented. The dynamic performance under different conditions, such as changing the current references and the speed, will be tested. Finally, at the end of this section, the proposed method is evaluated using an OWC Emulator, in order to check the feasibility of the proposed method during large variations of the rotor speed in a few seconds.

- In section 6.3, the results of the proposed MPC for FSTP GSC of DFIG are illustrated. The simulation studies are carried out in this chapter to show the effectiveness of the DC voltage deviation control and to present the dynamic behavior of the proposed method for the GSC of DFIG. Moreover, the proposed method is compared with the one voltage vector MPC and with the two voltage vector MPC. Furthermore, the results of the FSTP of GSC are compared with the six-switch three phase GSC, in order to observe the differences in the performance between the pre-fault and post-fault configurations. Finally, the dynamic performance under parameter variations are checked in a laboratory experimental DFIG setup, working in sub-synchronous and hyper-synchronous speeds.

6.2. Simulation and experimental results of Model Predictive Current control for the machine side converter of SPMSG when fed by four-switch three phase converter

In this section, the simulation results for a 8.7 kW SPMSG are carried out in order to show the dynamic performance of the proposed model predictive control for four-switch three phase machine side convert (MSC) of SPMSG, and the effectiveness of the voltage offset elimination control. Moreover, some of the many experimental tests intended to evaluate the dynamic response and the steady-state performance for a 8.7 kW SPMSG which is included in section 6.2.2. Finally, the FSTPC and the offset control is tested using an OWC power plant emulator with irregular waves in order to check the performances of the proposed method. It should be noted that the OWC based power plant emulator model and the method to extract the maximum energy from the OWC and the speed reference were explained in chapter 3

6.2.1. Simulation results: comparative analysis and discussion

The simulation studies for a FSTPC feeding a 8.7 kW SPMSG is carried out in Matlab/Simulink environment to validate the proposed MPCC. The machine parameters and capacitor values for the simulation studies as well as the experimental tests are illustrated in Table 6.1. The switching frequency and the sampling period are 4 kHz and 250 μ s, respectively. The grid side converter (GSC) for these tests was controlled to maintain the DC link voltage at 550 V.

Table 6.1. SPMSG parameters

DC Link voltage	550 V	Permanent magnet Flux	1.05 Wb
Rated power	8.7 kW	Stator Inductance	50 mH
Rated voltage	400 V	Stator resistance	2 Ω
Pole pairs	3	C1, C2	1080 μ F

First, the voltage offset control is tested and the results are shown in Fig. 6.1. In this study, the i_{sd} and i_{sq} references were set to 0 A and -4 A, respectively. The rotor speed was fixed at 400 rpm for this case. At the beginning, the voltage offset control was disabled. However, at $t=0.5$ s the compensation currents for removing the offsets were employed. As it can be seen in Fig. 6.1, the offset voltage between the two capacitors before $t=0.5$ s is about 140 V. Additionally, the total DC link voltage has oscillations around the 550 V value when the DC voltage deviation control is disabled. However, when the offset control is enabled, both capacitors mean voltages are set to 275 V and the total DC link voltage oscillation is removed.

Fig. 6.1. Simulation results of capacitor voltages and DC link voltage with and without the offset control.

The stator currents in dq and abc frame, the sector and duration times of the vectors without and with the compensation control are shown in Fig. 6.2 and Fig. 6.3, respectively. Clearly, the currents in dq frame could not follow the reference values in some points when the offset control is not working, which led to have a distorted and unbalanced current in phase “a”. This issue takes place since the upper capacitor (V_{c1}) has a lower voltage when there is not an offset control, meaning that the voltage vector V_3 has a lower value compared to normal situation. Therefore, when the reference voltage is at the end of sector 1 and at the beginning of sector 2, even by employing V_3 for the whole switching period, the required voltage will not be produced. Consequently, at the end of sector 1 and at the beginning of sector 2 cannot follow the references when the offset control is not working. On the contrary, the currents are following the reference values when the proposed compensation control is working and, as a result, the currents are sinusoidal and balanced.

Fig. 6.2. Simulation results without capacitor voltage offset elimination control.

Fig. 6.3. Simulation results with capacitor voltage offset elimination control.

To verify the dynamic response, the generator speed was decreased linearly from 500 rpm to 200 rpm during 1.5 seconds as shown in Fig. 6.4. Furthermore, the i_{sq} and i_{sd} references were changed from -6 A to -3 A at $t=0.7$ s and from -1 A to 1 A at $t=1$ s, respectively. The results are depicted in Fig. 6.4. In this test, the compensation control was applied for the whole test. The results indicate that the currents follow the reference values rapidly during different rotor speeds while the capacitor voltage offset remained at zero for the whole test. Moreover, the three phase currents are balanced due to the offset control.

Fig. 6.4. Simulation results during a rotor speed variation with current step changes.

The proposed model predictive control for FSTPC is compared with two previous predictive controls, one voltage vector [1] and two voltage vector predictive controls [2], for this same converter topology at the same switching frequency. In this test, the rotor speed was 400 rpm and the current references for d and q are set to 0 A and -6 A, respectively. The

three-phase stator currents and current THD for this comparison are shown in Fig. 6.5 and Fig. 6.6, respectively. The results show that the stator current for the proposed method is sinusoidal, balanced and free of ripples when compared to the two previous predictive strategies. The stator current THD for the one voltage vector predictive control, two voltage vector predictive control and the proposed strategy are 10.22%, 5.96% and 3.47%, respectively. Moreover, the low-order harmonics are decreased with the proposed control method while for the two previous control methods, these harmonics are notable. It should be noted that the current THD for three phase six switches converter with the same reference values would be 1.32% in this case (pre-fault situation).

Fig. 6.5. Three-phase stator current: (a) one voltage vector predictive control, (b) two voltage vector predictive control, (c) proposed model predictive control.

Fig. 6.6. Stator current THD: (a) one voltage vector predictive control, (b) two voltage vector predictive control, (c) proposed model predictive control.

6.2.2. Experimental results: analysis and discussion

In this part, the dynamic performance of the proposed method is checked in the laboratory setup. Moreover, the proposed MPCC of a FSTPC for SPMSG was tested in a wave energy application using the experimental setup shown in Fig. 6.7 and Fig. 6.8. This setup consists of: Part 1: OWC Power Plant, which reproduces the waves, OWC models and actual torque developed by a Wells turbine emulated by means of a DC/DC converter and a DC machine; Part 2: Power Take Off, where a FSTPC is connected to the SPMSG and is controlled with the proposed MPCC; and Part 3: Grid connection, which injects into the grid the energy that has been absorbed from the waves. Each of these three parts is controlled using a dual-core F28M35x Concerto MCU of Texas Instruments. The switching frequency and the sampling time for the machine side converter were selected to be 4 kHz and 250 μ s. All of the waveforms were recorded using the F28M35x itself, except the stator currents, that were recorded using a Tektronix TDS2024C oscilloscope with a sampling time of 100 μ s. The DC link voltage was set to 550 V during the experimental tests. The machine parameters and

the capacitor values are illustrated in Table 6.1. The capacitor values for the FSTPC are the same capacitor values that are used in the three phase six switches converters.

Fig. 6.7. Photo of the experimental setup used to reproduce an OWC-based power plant.

Fig. 6.8. Electrical scheme of the setup used to carry out the experimental tests.

Three types of experimental tests were carried out: firstly, a steady state test to check the proper operation of the control system. Secondly, three transient tests to check the performance of the control system during changes of the i_{sd} and i_{sq} references and when facing a rotor speed step. Finally, a transient test reproducing the behavior of an OWC wave energy power plant. In all of these tests the voltage offset elimination control was enabled.

6.2.2.1. Steady state test

In this first test, the rotor speed was set to 450 rpm, as can be seen in Fig. 6.9, by means of the DC motor whereas the values for the i_{sd} and i_{sq} references were set to 0A and -4A, respectively. Fig. 6.10 shows the currents in the dq frame. As can be checked, the currents are following the reference values perfectly. The DC link voltage and the capacitor voltages are shown in Fig. 6.11. As can be noticed in this figure, the DC link voltage is smooth without practically any ripple. Moreover, the capacitor voltages are balanced, and the voltage mean value for each of the capacitors is 275 V, which shows the effectiveness of the proposed voltage offset elimination control. Fig. 6.12 shows the three phase current proving that the proposed control system was able to obtain sinusoidal currents, despite the lack of two IGBTs with respect to the standard two-level converter topology.

Fig. 6.9. Rotor speed during the steady state test.

Fig. 6.10. Values of i_{sd} (red) and i_{sq} (blue) obtained from the phase current measurements.

Fig. 6.11. DC voltages in C_1 (light green) and C_2 (red) and DC link voltage (blue) during the steady state test.

Fig. 6.12. Three phase stator current of the SPMSG during the steady state test.

6.2.2.2. Speed transient test

This test was meant to check the capability of the proposed control to maintain the i_{sd} and i_{sq} currents, and consequently, the torque and magnetic field of the SPMSG, close to their references when the rotary speed faces a step change from 270 rpm to 400 rpm, as shown in Fig. 6.13. The test was carried out by applying a DC voltage-step to the armature winding of the DC motor. Fig. 6.14 shows that the i_{sd} and i_{sq} remained constant despite the speed step. Fig. 6.15 shows the capacitor voltages during this test. The two capacitor voltages and the total DC link voltage remained stable when the rotor speeds up during this test. Finally, Fig. 6.16 shows the three phase stator current during the speed step that took place in the SPMSG. It can be checked clearly that the frequency of the stator current is increased after the step.

Fig. 6.13. SPMSG rotor speed during the speed transient test.

Fig. 6.14. i_{sd} (red) and i_{sq} (blue) currents during the speed transient test. Fig. 6.15. DC voltages in C_1 (light green) and C_2 (red) and DC link voltage (blue) during the speed transient test.

Fig. 6.16. Phase currents at the stator of the PMSM during the speed transient test.

6.2.2.3. i_{sq} transient test

In this test, the SPMSG torque was increased through the i_{sq} reference from -2 A to -6 A whereas the DC motor supply remained unchanged. As a result, the rotor speed decreased as the load torque generated by the SPMSG temporarily exceeded the motor torque provided by the DC motor. The rotor speed is shown in Fig. 6.17. Fig. 6.18 and Fig. 6.19 show the fast dynamic response of the proposed control. As can be seen, the i_{sq} follows the step in the reference with great accuracy and i_{sd} remains to zero at every moment. The capacitor voltages are shown in Fig. 6.20. The DC voltages in C_1 and C_2 experimented a higher fluctuation after the transient due to the higher current demand and the lower speed. As can be checked in this figure, the voltage offset control works perfectly and quickly during the transient test and both capacitors mean voltage value is maintained at 275 V. Finally, Fig. 6.21 shows the sinusoidal currents near the i_{sq} transient test, where the waveforms get adapted to the new amplitude immediately, keeping a low distortion.

Fig. 6.17. SPMSG rotor speed during the i_{sq} transient test.

Fig. 6.18. i_{sq} reference (red) and actual i_{sq} (blue) during the i_{sq} transient test.

Fig. 6.19. i_{sd} reference (red) and actual i_{sd} (blue) during the i_{sq} transient test.

Fig. 6.20. DC voltages in C_1 (light green) and C_2 (red) and DC link voltage (blue) during the i_{sq} transient test.

Fig. 6.21. Phase currents at the stator of the SPMSG around the point where the i_{sq} transient takes place.

6.2.2.4. i_{sd} transient test

In this test, the SPMSG changes from a field weakening operation point to an overexcited point ($i_{sd}=-1\text{A}$ to $i_{sd}=+2\text{A}$), maintaining the torque constant through $i_{sq}=-4\text{A}$. The field weakening was limited to -1A on purpose to avoid the irreversible demagnetization of the magnets.

The results are shown in Fig. 6.22 to Fig. 6.26. The rotor speed is fixed at 300 rpm as can be seen in Fig. 6.22. The i_{sd} and i_{sq} currents for this test are shown in Fig. 6.23 and Fig. 6.24, respectively. The results show that the i_{sd} follows the reference value perfectly in the proposed MPCC while the i_{sq} is kept unchanged. The three phase stator current is illustrated in Fig. 6.25 which shows that the good quality of the sinusoidal current continued after a brief transient. Fig. 6.26 shows that the voltage offset control is working properly during the transient while keeping the mean value of the capacitor voltages at the set point value.

Fig. 6.22. Rotor speed during the i_{sd} transient test. Fig. 6.23. i_{sd} reference (red) and actual i_{sd} (blue) during the i_{sd} transient test. Fig. 6.24. i_{sq} reference (red) and actual i_{sq} (blue) during the i_{sd} transient test.

Fig. 6.25. Phase currents at the stator of the SPMSG around the point where the transient takes place.

Fig. 6.26. DC voltages in C_1 (light green) and C_2 (red) and DC link voltage (blue) during the i_{sd} transient test.

6.2.2.5. Tests using OWC Emulator

The final test was intended to study the proposed control system to be used in an OWC based power plant. The lack of two IGBTs could mean a reduction of the control capability of the electronic converter, forcing to disconnect it from the electrical grid, leading to a reduction of profits.

The test was carried out for the wave profile showed in Fig. 3.15 in chapter 3 and lasted 4 minutes. The 6 microcontrollers that comprise the emulator take care of calculating the air speed in the duct of the OWC, the Wells turbine torque, the Wells turbine maximum efficiency speed to be imposed by the PTO, etc., throughout the entire test, at every control cycle.

In Fig. 6.27, the Wells turbine maximum efficiency speed and the actual speed are shown in blue and red lines, respectively, throughout the 240 seconds of the test. This figure shows that the rotary speed of the generator follows the reference speed with great accuracy with the proposed MPCC during the 4 minutes that lasted the test, in the entire range of speed containing fast and slow reference speed changes. Fig. 6.28 provides a more detailed view throughout the first 40 seconds of the test.

Another critical issue is to check the performance of the voltage offset elimination control during this test since, if it is not fast and accurate enough, it could lead to the breakdown of the capacitors. Fig. 6.29 shows the capacitor voltages and the DC link voltage of the first 40 seconds of the test. It can be seen that the proposed method was capable of eliminating the voltage offset of the capacitors during the huge variations of speed, which indicates the effectiveness of the proposed MPCC and the offset elimination control.

Fig. 6.27. Speed reference (blue) and actual speed (red) of the SPMSG along the 4 minutes test.

Fig. 6.28. Detailed view of the speed reference (blue) and actual speed (red) of the SPMSG along the first 40s of the test.

Fig. 6.29. DC voltages in C_1 (light green) and C_2 (red) and DC link voltage (blue) during the test.

6.3. Simulation and Experimental results of Model Predictive Direct Power Control Strategy for Four-Switch Three Phase Operation of Grid Side Converter

In this section, the performance of the proposed model predictive control for the FSTP grid side converter is evaluated in Matlab Simulink and afterwards, it is checked in a laboratory prototype. The GSC of DFIG was selected to test the performance since it is bidirectional converter and the GSC can absorb or inject active power during sub-synchronous and hyper-synchronous speeds. The simulation and experimental results for the FSTP GSC of a 5 kW DFIG are shown in section 6.3.1 and 6.3.2, respectively.

6.3.1. Simulation results: comparative analysis and discussion

In this section, the performance of the proposed predictive control for the FSTP GSC of DFIG was tested in Matlab/Simulink. The schematic of the simulated system and the experimental setup is shown in Fig. 6.30. Since the Semikron SKS 35F B6U converters were used in the experimental setup, the midpoint of the capacitors was not available. Therefore, two series capacitors (C_1 , C_2) were connected in parallel with the inner Semikron capacitor (C_3) and the midpoint of the new capacitors (C_1 , C_2) was connected to the phase “a”. However, in simulation studies, the C_3 was not considered. The parameters of the system are summarized in Table 6.2. The sampling time and the switching frequency for the simulation studies were 250 μ s and 4 kHz, respectively.

Fig. 6.30. Schematic of DFIG setup.

Table 6.2. Parameters of Experimental DFIG setup

DC link voltage	450 V	Stator leakage inductance	8.6 mH
Rated power	5 kW	Rotor leakage inductance	8.6 mH
Grid voltage	400 V	Turns ratio	2.19
Rated stator voltage	400V	Inertia	0.16 kg.m ²
Rated rotor voltage	182 V	C ₁ , C ₂	1080 μ F
Number of pole pairs	3	C ₃	4080 μ F
Stator resistance	0.65 Ω	L _{gs} , L _{fs}	20 mH
Rotor resistance	2.2 Ω	R _{gs} , R _{fs}	1 Ω
Mutual inductance	81 mH		

First of all, the performance of the voltage offset suppression control is tested. In this test, the voltage offset control was applied from the beginning. At $t=0.1$ s the compensation power reference for the offset control was removed from the control. In this case, the generator was injecting 3 kW and 0 var into the grid through the stator. The generator speed was fixed at 800 rpm during the entire simulation tests. The results are represented in Fig. 6.31. As it can be seen, at $t=0.1$ s, the capacitor voltages start to deviate from 225 V and ultimately, there is a 70 V deviation between the two capacitors voltages at steady state. Moreover, the overall DC link voltage starts having small oscillations when the control is removed. The sector location and duration time of each vector is also presented in Fig. 6.31, showing that the proposed control can perfectly determine the proper three voltage vectors and that the corresponding times of each vector are positive and are within the switching period.

Fig. 6.31. Simulation results when the voltage offset control is removed.

A more detailed view of the GSC currents with and without the offset control is shown in Fig. 6.32. In this figure, the currents are balanced when the offset control is applied. However, it is clear that when the offset control is removed the currents are not balanced anymore.

Fig. 6.33 shows the compensation current reference ($\bar{i}_{go}$) when the offset control is enabled in Fig. 6.31. As it is expected, this current is very close to zero in steady state. In

order to produce the compensation power references, this DC current is multiplied by the grid voltage in the stationary frame. As a result, the compensation power references have a very small oscillating term with 50 Hz frequency. This very small ripple is shown in Fig. 6.31. By increasing the sampling time, the ripple of the compensation current reference will decrease, and this power ripple will also decrease.

Fig. 6.32. Simulation result for GSC current with and without offset control, top is with offset control and bottom is without offset control.

Fig. 6.33. Compensation current reference ($\bar{i}_{g0}$) for removing the DC voltage offset.

In order to have a fair evaluation of the predictive control for the FSTP GSC performance, the GSC currents of the proposed control for the FSTP was compared to the currents of a six-switch three-phase (SSTP) at steady state for the same condition. A predictive control was applied for the SSTP GSC study [3]. The generator speed was constant at 800 rpm and the stator active and reactive powers were set to 3kW and 0 var, respectively. The currents and current THD for the FSTP and SSTP are shown in Fig. 6.34. The currents for the FSTP and SSTP are balanced and sinusoidal. The current THD for the FSTP and SSTP is 3.18% and 2.03%, respectively, which shows the SSTP has a better current THD. However, it should be noted that the main differences are in the switching frequency harmonics which could be removed by a custom designed filter. In the low-order harmonics, both topologies have almost the same harmonic values.

Fig. 6.34. The GSC current and current THD: (a) for FSTP; (b) For SSTP.

The proposed model predictive control is compared with two other predictive controls, which were presented for FSTP topologies. The first model predictive control is the conventional model predictive control which is based on one voltage vector [1]. The second one is two vectors based model predictive control [2]. The results from this comparison are shown in Fig. 6.35. In this study, the switching period time for all three methods was $250 \mu\text{s}$ and the generator was injecting 3 kW and 0 var to the grid, while the rotor speed was constant at 800 rpm. As it can be seen in Fig. 6.35, the currents are distorted in the conventional model predictive control. The current waveforms get better by employing two vectors in every switching period compared to conventional model predictive control. However, these two model predictive controls have notable low order harmonics. As it can be clearly seen, the proposed model predictive control has the best current waveform if compared with the two other model predictive controls. Moreover, the current THD for conventional model predictive control, two vectors based model predictive control and the proposed model predictive control are 15.46%, 7.96% and 3.18%, respectively.

Fig. 6.35. The GSC current and current THD for FSTP with different predictive strategies: (a) predictive control based on one voltage vector [1]; (b) predictive control based on two voltage vectors [2]; (c) proposed predictive control.

In order to evaluate the effect of the parameter variation on the performance of the proposed PDPC, it has been assumed that the grid side converter inductance is 15 mH whereas the actual inductance is 10 mH. The GSC current for this case when the generator is injecting 3 kW and 0 var into the grid at the speed of 800 rpm is shown in Fig. 6.36(b). The GSC current without inductance error is shown in Fig. 6.36(a). As it can be seen, both figures have the same ripple and performance. The THD of the current in Fig. 6.36(a) and Fig. 6.36(b) is 6.61% and 6.71%, respectively. This shows that in the case of 50% error in the grid side inductance, the current THD will increase slightly.

Fig. 6.36. GSC current: (a) without any variation in parameters (b) with 50% variation in the grid side inductance.

6.3.2. Experimental results: analysis and discussion

The proposed control for the FSTP GSC was further evaluated on a 5 kW DFIG laboratory prototype as shown in Fig. 6.37. The proposed controller was implemented on a dSPACE 1104. Each of the converters was controlled by a separate dSPACE 1104. The sampling period and the switching frequency for the GSC dSPACE were set as 125 μ s and 4 kHz, respectively. Beside the hardware protection in IGBT drivers to prevent short circuits, 3.3 μ s dead time was selected in the dSPACE code as a software protection. The computation burden for the whole program in Dspace is 60 μ s. Many tests were carried out in order to show the performance during the steady and transient states. The experimental results were acquired by the dSPACE control desk and data were registered at 500 μ s sampling rate. In the whole experimental test, the stator was absorbing 1500 var reactive power from the grid in order to supply a part of the magnetization current from the stator and to avoid over current in the rotor windings. The stator active power was set to 1 kW during the different tests except for the last one. The offset control was employed during the whole experimental tests.

Fig. 6.37. Experimental DFIG setup.

First, the dynamic performance was evaluated by changing the GSC reactive power reference. The rotor speed was fixed at 0.8 p.u. and the DC link voltage was set to 450 V. The reactive power was reduced in a step from 200 var (GSC injects to the grid) to -300 var (GSC absorbs) at $t=0.1$ s. The results are presented in Fig. 6.38. The proposed control followed the reactive power reference accurately while the DC link voltage remained constant, and both capacitors had a same average voltage value.

The current THD when the reactive power was -300 var, was estimated by the means of a power analyzer PPA1530, and it is shown in Fig. 6.39.

Fig. 6.38. Experimental results during GSC reactive power reference change.

Fig. 6.39. Harmonic spectrum of GSC current when $Q_g = -300$ var, $P_s = 1$ kW.

Second, the DC link voltage reference was changed from 450V to 410V at $t = 0.4$ s, as shown in Fig. 6.40. The results indicate that the proposed control follows the DC link voltage accurately. However, an overshoot can be observed in the DC link voltage due to the existence of the capacitor C_3 in the prototype. The offset suppression control works correctly during the transient states and the steady state.

Fig. 6.40. Experimental results during a step down of DC link voltage.

Fig. 6.41 shows the performance of the GSC during sub and hyper synchronous speed. In this test, the rotor speed was increased from 810 rpm to 1080 rpm while the stator active and reactive power references were constant. By increasing the speed, the GSC current is decreased as it can be seen. Additionally, the voltage deviation control maintains the capacitor voltage offset to zero during the test.

Fig. 6.41. Experimental result when the generator speed is increased.

Finally, the stator active power was reduced in a step from 2 kW to 1 kW at $t = 0.3$ s. whereas the generator speed was 820 rpm at the beginning as shown in Fig. 6.42. Since the DC motor voltage was constant, the speed of the generator was increased when the injected stator active power was reduced as it can be observed in this figure. The GSC current was decreased after the reduction in stator power. No offset can be observed in this case in the capacitor voltages. It should be noted that the DC link voltage was increased slightly at the beginning of this step change.

Fig. 6.42. Experimental result when the stator active power changed.

The performance of the proposed predictive control is evaluated with three different grid side inductances. In this test, the initial situation in Fig. 6.38 was repeated. The GSC is injecting 200 var in this case while the DC link voltage is fixed at 450V. The GSC current results for 20 mH, 10 mH and 5mH are shown in Fig. 6.43(a), Fig. 6.43(b) and Fig. 6.43(c), respectively. As it can be seen, by decreasing the inductor, the ripple increase. However, the proposed control works accurately.

The effect of the grid side inductance variation on the proposed predictive control is investigated when the assumed inductance is 50% smaller than the actual value. The GSC current with this inductance error and without it, is shown in Fig. 6.44(b) and Fig. 6.44(a), respectively. In this test, the GSC was injecting 200 var into the grid while the DFIG stator was injecting 1kW and absorbing 1.5 kvar, and the rotor speed was 800 rpm. The actual

inductance is 10 mH for this test while the assumed value was 5 mH. These two figures show that the proposed control is not sensitive to a 50% error in the grid side inductance.

Fig. 6.43. GSC current with different inductance (L_{fg}): (a) with 20 mH inductance (b) with 10 mH inductance (c) with 5 mH inductance.

Fig. 6.44. GSC current: (a) without any variation in parameters (b) with 50% variation in the grid side inductance.

6.4. Summary and conclusions

In this chapter, the performance of the proposed predictive controls for FSTP MSC of SPMSG and for FSTP GSC of DFIG simulated in the Matlab Simulink environment and applied in a laboratory setup have been presented.

Firstly, in section 6.2 of this chapter, the proposed Model Predictive Current control (MPCC) for the FSTP machine side converter has been tested. Despite the lack of two IGBTs in the MSC, the proposed MPCC system has proven to be able to set and keep the references, basically the torque and the magnetic field, controlled by i_{sq} and i_{sd} , as well as to remove the voltage offset of the capacitors, during both steady-state and speed transients. Also, when it comes to references' transients, the experimental results showed that this system features a very fast dynamic response and total independence in the control of i_{sd} and i_{sq} . Besides, the proposed voltage offset suppression control works perfectly and rapidly even when the rotor speed is low, and the speed variations are high.

Additionally, the proposed method for the FSTP machine side converter was compared with two other previous model predictive controls for the FSTPC. The results showed that the proposed predictive control has less low-order harmonics in the stator current and lower current THD compared to predictive controls based on one voltage vector and two voltage vectors. In this study, the current THD for predictive control was 3.47% while for the model predictive controls, based on one voltage vectors, and two voltage vectors were 10.22% and 5.96%, respectively. Moreover, the highly-demanding speed test corresponding to the control of an OWC-based power plant has demonstrated the capability of this control system to follow the maximum efficiency speed reference of the Wells turbine, whereas the capacitor voltage offset was eliminated during the test. As a result, the power plant could keep generating power despite the fault in the power converter.

Secondly, in section 6.3 of this chapter, the proposed model predictive method for a four-switch three phase GSC was first evaluated in Matlab Simulink. The result indicates that the voltage deviation in the capacitors for a four-switch three phase system can be removed by the use of the presented voltage offset control in the capacitors. Moreover, by eliminating the DC offset in capacitors, the currents will be balanced. The performance of the FSTP GSC was compared to a SSTP system working in the same conditions. The current THD for FSTP and SSTP was 3.18% and 2.03%, respectively. The result showed that, although the current THD is better in the SSTP, the low-order harmonics in both systems are almost the same.

A series of experimental tests were later carried out in section 6.3 for FSTP GSC of DFIG in order to assess the transient performance of the proposed control. The results reveal that the proposed method can follow the references while the DC voltage offset in capacitors remains at zero value. Furthermore, during hyper-synchronous speed and sub-synchronous speed, the proposed method works effectively. Finally, it can be concluded that the proposed predictive control for the FSTP GSC is an appropriate solution during one leg failures of six-switch GSC.

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Conclusions and Future works

7.1. Conclusions

The work described in this document has been focused on improving the control of the power converters in a DFIG and a SPMSG. A novel model predictive control has been presented in this dissertation for both power converters, GSC and MSC, of these two types of generators. According to the MPC strategies that have been presented in this document for the two-level power converters of DFIG and SPMSG, the following conclusions can be drawn:

A. Proposed predictive direct power control (PDPC) for the rotor side converter (RSC) of DFIG:

The proposed PDPC for the RSC of DFIG was presented for balanced and unbalanced grid voltages. In this method, four rotor voltage vectors are used within a fixed sampling period. The appropriate and correct voltage vector sequence in each period is recognized according to the estimated duration times of the selected vectors. The negative value of the estimated duration time of the selected vectors has been also considered in this strategy. The proposed strategy is examined during balanced and unbalanced grid voltage operation. The simulation and laboratory experimental results showed that the proposed strategy can track the power references perfectly keeping the switching frequency constant. Moreover, the proposed strategy could obtain smooth active and reactive powers or eliminate electromagnetic torque pulsations or provide sinusoidal and balanced stator currents under unbalanced grid voltage, without using any PI or resonant controllers. In the proposed PDPC, neither the estimation of the rotor flux nor using a phase-locked loop (PLL) is needed. The voltage vector sequences are selected according to a simple algorithm. The proposed control strategy works perfectly under disturbed grid voltage and near the synchronous speed and can inject oscillating active and reactive power components into the grid. However, the proposed algorithm for selecting the appropriate vectors is somewhat more complicated than previous PDPC strategies for the RSC. The comparison between the proposed PDPC with vector control, direct power control, a previous predictive control during balanced grid voltage has shown that the proposed method has a faster dynamic response and lower stator current THD which make it superior compared to other control methods. Moreover, the comparison between the proposed PDPC and two previous PDPC strategies during disturbance, indicates that the proposed strategy can obtain the desired control target with less power ripple and

better current THD, which again constitutes a remarkable contribution to the state of the art in this field.

B. Proposed model predictive direct current control (MPDCC) for the machine side converter (MSC) of SPMSG:

The proposed MPDCC for the MSC of SPMSG is easy to implement in a microcontroller and its computing time means not a bigger burden than the amount of time typically spent in classical vector control. In this method, the required stator voltage sector is predicted, and four voltage vectors are applied in every switching interval. Results have shown that the MPDCC can effectively predict the appropriate sector and voltage vector sequences while keeping the switching frequency fixed as well. The magnetic field and torque are controlled by means of the predictive stator current control in the d-q frame. In this strategy, the ripple in the stator currents is very low. The performance of the proposed method has been validated by simulation and experimental results and a fast dynamic performance was observed. The simulation results revealed that the proposed MPDCC has a significantly better performance compared to the conventional one voltage vector and two-voltage vector MPC methods. Moreover, electric currents present a low current THD and the rest of the variables have low ripple. The experimental tests carried out using an OWC power plant emulator have shown how the MPDCC is able to make the rotary speed follow the reference although it changes in a wide range and in a very irregular manner. Therefore, it can be said that the proposed MPDCC presents a suitable behavior to control a SPMSG in wave energy conversion. This new control MPDCC proposed is a contribution itself in the control of MSC of SPMSG and its application in the control of OWC wave generators presents a very interesting solution in this field.

C. Proposed model predictive direct power control (MPDPC) for the GSC of DFIG and PMSG

In chapter 3 of this dissertation, the proposed MPC strategy was extended for the GSC of DFIG and PMSG. Since the GSC of DFIG is a bidirectional converter, the proposed MPDPC control considers this issue. Without using a PLL controller and grid sector position, the voltage vector sequences are selected and employed. In this method, the GSC active and reactive power references are directly tracked according to a cost function. Simulation and experimental results were carried out to check the performance of this new method for the GSC of DFIG and SPMSG. The results indicated that it can handle the sub-synchronous and hyper-synchronous operation of the DFIG so bidirectional performance was achieved. The results showed that the power ripples are very low with this method and during the transients it follow the power references very quickly. Moreover, the proposed method was compared with two previous MPC methods obtaining a current THD for the proposed method which was lower than that in the other MPC strategies. It is important to realize that this novel control for the GSC can also be applied to any DC system connected to the ac grid, such as energy storage systems connected to the grid, PV generation systems and fuel cell generators connected to the grid. Thus, the applications of this MPDPC proposed in this Thesis are not only restricted to the wind and wave generators treated in this work.

Regarding the use of a four-switch three phase topology in the converter, which is the resulting configuration of a typical six switch three phase bridge after certain type of failures, such as losing one leg after a switch breakdown, some contributions have been also presented in this work. The following conclusions can be drawn according to predictive control strategies that have been presented in this document for the four-switch three phase converters of SPMSG and DFIG:

D. Proposed model predictive current control (MPCC) for four-switch three phase converter for SPMSG

In chapter 4 of this document, a simple model predictive current control with capacitor voltage offset elimination control for the four switch three phase converter feeding a SPMSG was presented. The proposed method features a fixed switching frequency and a fast dynamic response despite the fact that it is working with only four switches instead of six. In every switching period, three voltage vectors are employed according to the current references and the rotor speed. The proposed capacitor voltage suppression strategy can remove the voltage offset in the capacitors rapidly without needing to obtain the mean values of the capacitor voltages using a low-pass filter, which would be difficult at low speeds. Despite the lack of two IGBTs in the MSC, the proposed MPCC system has proven to be able to set and keep the references, basically the torque and magnetic field, controlled by i_{sq} and i_{sd} respectively, as well as to remove the voltage offset of the capacitors, during both steady-state and speed transients. Also, the experimental results showed that this system features a very fast dynamic response during transients. Besides, it was proved that the proposed voltage offset suppression control works perfectly and rapidly even when the rotor speed is low, and the speed variations are high, which is a very important issue for wave energy applications. Furthermore, the proposed method was compared to other two previous MPC strategies for the four-switch three phase converters, and the results showed that the proposed predictive control has less low-order harmonics in the stator current and lower current THD compared to predictive controls based on one voltage vector and two voltage vectors. In this comparison, the current THD for the predictive control was 3.47% while for the predictive controls based on one voltage vector, and two-voltage vector were 10.22% and 5.96%, respectively. The highly demanding speed test corresponding to the control of an OWC-based power plant has demonstrated the capability of the proposed MPCC system to follow the maximum efficiency speed reference of the Wells turbine whereas the capacitor voltage offset was eliminated during the test. As a result, the power plant could keep generating power despite the fault in the power converter. This feature could be also helpful in the case of buoy-based WECs since it allows to keep the control of buoy in the case of a breakdown in the power converter.

E. Proposed model predictive direct power control strategy for four-switch three phase operation of grid side converter

In this method, three voltage vectors among the four possible voltage vectors of four-switch three phase GSC are applied to guarantee a constant switching frequency and a low ripple in the powers at any circumstance. Moreover, a voltage offset suppression control was presented for the four-switch three phase GSC without changing the voltage vector sequences algorithm. The proposed predictive control was first evaluated in Matlab Simulink. The result indicates that the voltage deviation in the four-switch three phase system can be removed by

the use of the presented control. Moreover, by eliminating the DC offset in the capacitors, the currents will be balanced. The performance of the four-switch three phase GSC was compared to a six-switch three phase system working in the same conditions. The result showed that, although the current THD is better in the six-switch three phase system, the low-order harmonics in both systems are almost the same. Additionally, a series of experimental tests were carried out in order to assess the transient performance of the proposed control for the four-switch three phase of GSC of DFIG. The results revealed that the proposed method can follow the references normally while the DC voltage offset in capacitors remains at zero value. Besides, it has been proven that during hyper-synchronous speed and sub-synchronous speeds in the DFIG, the proposed method for the GSC works effectively. Finally, it can be concluded that the proposed predictive control for the four-switch three phase GSC of DFIG is an appropriate solution during a one leg failure of six-switch converters.

7.2. Future works

There are still some research aspects for the proposed MPC control that are very interesting and can be addressed in the future works, such as:

- As mentioned in this document, the proposed predictive control has a fast dynamic response. However, in some applications, it is required to have a slow dynamic response and the response time needs to be set with a slower value (desired value). As a result, the MPC control should consider the delay in reaching the reference and the response time should be added in the cost function. Therefore, the equations of duration times of the vectors will be different. This idea could be further investigated in the future.
- In this thesis, the cost function of the MPC control was set to minimize only the power errors or the current errors. However, other objectives such as decreasing the switching power losses can be added to the cost function setting their weighting factors according to the importance. This is a very interesting subject and needs more research to obtain the appropriate duration time for this case. This idea has not been reported before for a fixed switching MPC strategy.
- The GSC can be connected to the grid through an inductive-capacitive-inductive (LCL) filter due to the good features that bring this filter, such as low current THD. However, the performance of the filter near the resonant frequency is a big concern and might cause the instability of the system. Hence, the MPC control that has been presented in this document could be extended for GSC using an LCL filter where the stable operation should be considered, and the ability of the control system should be improved.
- The stability analysis for the proposed predictive control for the power converter of DFIG and SPMSG in microgrids is also an interesting subject, since the stability of the system is an important concern in microgrids. Many instabilities have been reported in the literature in the microgrids, where some of them were the interactions between the renewable power generation systems with other systems in the microgrids. The impedance-based stability analysis is an efficient solution to predict instability. As a result, the impedance-based stability of the proposed method should

be studied in the future. In this case, the proposed MPC method should be linearized and the impedance of the whole system, which consists of the filters and the control, should be estimated.

Appendix

A.1. Appendix. A: The values of k_1 to k_6 of equations (3.21) and (3.22)

The values of k_1 to k_6 , and β_+ , β_- and δ of equations (3.21) and (3.22) are shown below:

$$\begin{aligned} k_1 &= -(\omega_s - \omega_m)Q_s + 1.5\lambda L_r |V_s|^2 - 1.5\lambda L_r \omega_m (|v_s^+| |\varphi_s^+| - |v_s^-| |\varphi_s^-|), & k_2 &= -\lambda L_m |v_s^+| V_{dc}, \\ k_3 &= -\lambda L_m |v_s^-| V_{dc} & k_4 &= -1.5\lambda L_r \omega_m (|v_s^-| |\varphi_s^+| - |v_s^+| |\varphi_s^-|), & k_5 &= (\omega_s - \omega_m)P_s, \\ k_6 &= 1.5\lambda L_r \omega_m (|v_s^-| |\varphi_s^+| + |v_s^+| |\varphi_s^-|), & \beta_+ &= \alpha_+ - (n-1)\pi/3, & \beta_- &= -\alpha_- + (n-1)\pi/3, & \delta &= \alpha_+ - \alpha_- \end{aligned}$$

A.2. Appendix. B: Novel Differential Protection Technique for Doubly Fed Induction Machines

A.2.1. Introduction

Electric machines should use protection relays to safeguard themselves against abnormal operations or faults, such as short circuits or overloads [1]–[3]. These situations could be dangerous both for the facilities and the operators. In case of fault, failure analysis is vital to improve the machine design, the protections, and the maintenance schedule [4].

The rotor-wound asynchronous machine is a well-known machine, traditionally used as motor for high torque starting loads. The use of this machine has been considerably increased recently due to the expansion of the wind energy, and the large-size variable speed pumped-storage power plants.

The increasing importance of wind energy worldwide has originated the development of generators specially designed for this application [5]. In order to increase the wind energy production, it is important that the generator can modify the rotation speed. Thus, a great part of the wind generators manufactured today are doubly fed induction generators (DFIG), since they can vary the speed with a reduced size power converter [6]–[8].

Nowadays, due to its larger capacity and its high efficiency, pumped-storage hydropower plants are the most suitable backup generators for wind power integration [9].

Generally, the doubly fed induction machines (DFIMs) are protected by protection relays from the stator windings, as there is no specific protection for the rotor in the market. Some protection functions for the rotor are included in the converter controller.

However, in recent years, some investigation started to develop a new protection relay for the rotor of DFIM [10], [11]. The rotor windings of a variable speed DFIM are connected to a back-to-back converter and the rotor current frequency generally oscillates between 0.1 and 6 Hz during normal operation for hydroelectric power generation, and it is usually between 0 to 20 Hz for the wind turbine generation [12]. Thus, new sensors should be used in these new protections due to the low-frequency currents [10], [11].

Due to the fact that the rotor of DFIM is a very expensive component, especially in machines of hundreds of megawatts in pumped-storage stations, some specific protection for temperature monitoring has been recently developed [13].

It is a common practice that the main protection of generators and, in general, any type of electrical machines is the differential protection. This protection is based on the comparison of two or more currents and it detects internal faults. Differential protections are extendedly used in power systems, such as lines, cables, bus bars [14], and in all types of machines, such as transformers [15] or rotating machines [16].

Although the differential protections are invented many years ago, this research field is very active. Many research works have recently been published such as: new communication systems for differential protections [14], new algorithm to avoid unwanted trips [15], [17], [18], new current sensor to avoid saturation [19], [20], or new differential protections for distributed generation [21] among others.

In rotating machines, both induction- or synchronous-type, the differential protection only protects the stator windings through the comparison of the currents at both endings of the windings. According to the best knowledge of the authors, the differential protection was not used to protect the entire DFIM machine, including the rotor, until now.

Differential protection relays will trip in case of three-phase faults, phase-to-phase faults, and double-phase-to-ground faults. Only in case of grounded generators, the differential protection will detect some single-phase-to-ground faults. This section presents a novel differential protection for DFIMs. The main advantage of this new protection is that both the rotor and the stator are protected along different operating speeds and that the implementation of this protection is simple [22], [23] and allows distinguishing between internal and external faults.

Simulations and experimental tests have been performed in case of three-phase and phase-to-phase faults.

This part is structured as follows. Section A.2.2 presents the basic operating principle of the differential protection. Section A.2.3 presents the operation principle of the new protection proposed for the DFIM. The Section A.2.4 analyses the results of the simulations, whereas in Section A.2.5, the results of the experimental tests performed in the laboratory are presented. Finally, Section A.2.6 resumes the conclusions of this part.

A.2.2. Differential protection: state of the art

The differential protection is a very common protection used in a large range of electrical devices, especially in electrical machines. The simplified schematic of a differential

protection for motors and generators is shown in Fig. A.1. Its operation principle is based on the comparison of the input and output currents in each phase winding, as expressed in the following:

$$I_{diff} = |\vec{I}_T - \vec{I}_N| \quad (A.1)$$

Fig. A.1. Simplified schema of a motor/generator differential protection.

Fig. A.2. Simplified schema of a transformer differential protection.

Where the differential current (I_{diff}) is calculated as the difference between the neutral side (I_N) and the terminal side (I_T) currents. The restraining current (I_b) is obtained using different expressions, depending on the relay manufacturer, but the most commonly used expression is presented as

$$I_b = \frac{|\vec{I}_T| + |\vec{I}_N|}{2} \quad (A.2)$$

The differential protection algorithm is based on the comparison of the differential I_{diff} and the restraining I_b currents [1]. Only the stator is protected. For transformers protection, the differential algorithm is more complex, basically because the protection should compensate the vector group, the transformer ratio, and the inrush current of the transformer (see Fig. A.2).

A.2.3. Novel differential protection for DFIM

A.2.3.1. Basic Principles of DFIM

The dynamic model of a DFIM in the synchronous frame is shown in Fig. A.3. According to this figure, the stator active and reactive powers can be presented as follows:

$$P_s = \frac{3}{2}(V_{sd}i_{sd} + V_{sq}i_{sq}) \quad (A.3)$$

$$Q_s = \frac{3}{2}(V_{sq}i_{sd} - V_{sd}i_{sq}) \quad (A.4)$$

Fig. A.3. Dynamic model of DFIM in synchronous frame: (a) d-axis, (b) q-axis

If a voltage oriented control (VOC) is employed, the stator voltage will be placed along the d-axis establishing the synchronous reference frame and considering the stator resistant (R_s) negligible. Therefore, the stator flux could be shown as

$$\varphi_{sd} = 0 = L_m i_{rd} - L_s i_{sd} \quad (A.5)$$

$$\varphi_{sq} = -\frac{V_{sd}}{\omega_s} = L_m i_{rq} - L_s i_{sq} \quad (A.6)$$

Thus, the relation between the stator current and the rotor current in the synchronous frame in a VOC system can be presented as

$$i_{rd} = \frac{L_s}{L_m} i_{sd} \quad (\text{A.7})$$

$$i_{rq} = -\frac{|\vec{V}_s|}{\omega_s L_m} + \frac{L_s}{L_m} i_{sq} \quad (\text{A.8})$$

The differential current in the synchronous frame can be expressed as

$$|i_{rd} - i_{sd}| = \left(\frac{L_s}{L_m} - 1\right) i_{sd} \quad (\text{A.9})$$

$$|i_{rq} - i_{sq}| = \left| -\frac{|\vec{V}_s|}{\omega_s L_m} + \left(\frac{L_s}{L_m} - 1\right) i_{sq} \right| \quad (\text{A.10})$$

Since the leakage inductance is much smaller than the mutual inductance, especially for large induction machines, L_s/L_m is considered to be equal to 1. Hence, the differential current of rotor and stator in the d- and q-axis can be estimated by

$$|i_{rd} - i_{sd}| \approx 0 \quad (\text{A.11})$$

$$|i_{rq} - i_{sq}| \approx \left| \frac{|\vec{V}_s|}{\omega_s L_m} \right| \quad (\text{A.12})$$

As a result, during normal operation, the differential current in the d-axis has a small value, theoretically zero, and in the q-axis is almost equal to the magnetization current of the machine. It is remarkable that the calculation of I_d and I_q in the stator and rotor does not depend on the machine speed and that the proposed method operates correctly at any machine operation conditions.

A.2.3.2. Proposed Differential Protection for DFIM

The proposed protection is based on the comparison of the vector components I_d – I_q of the rotor and stator currents. The block diagram of the proposed differential protection for DFIM is shown in Fig. A.4.

A phase-locked loop is used to detect the stator voltage phase position. This angle (θ_s) is needed to convert the three-phase stator current to dq synchronous frame. Moreover, the difference between the stator voltage angle and the electric rotor angle (θ_m) is utilized to convert the rotor current to the synchronous frame (i_{rd} , i_{rq}).

Then, the rotor current is referred to the stator by multiplying by the turn ratio of stator and rotor windings (N). The rotor and stator currents in the synchronous frame are compared, and differential d and q currents are obtained (I_{diff_d} and I_{diff_q}). Finally, these differential currents are compared to the settings ($I_{diff_d \text{ set}}$ and $I_{diff_q \text{ set}}$) and the machine would be tripped if any of the differential currents were larger than the settings.

Fig. A.4. Simplified schema of a DFIM differential protection.

A.2.4. Simulation Results

In order to evaluate the performance of the proposed differential protection for the DFIM, four fault cases have been studied for a 2-MW DFIM using MATLAB/Simulink. The parameters of the simulated DFIM [24] are included in Table A.1.

Table A.1. Parameters of simulated DFIM

Rated power	2 MW	L_m	3.362 pu
Rated voltage	690 V	DC link nominal voltage	1200 V
R_s	0.0108 pu	Number of pole pairs	2
R_r	0.0121 pu	Lumped inertia constant	0.5
L_{ls}	0.102 pu	Grid frequency	50 Hz
L_{lr}	0.11 pu	N_s/N_r	0.333

For all the cases in this section, the DFIM was working as a generator and the stator was injecting 1.5 MW and 0 Mvar. The generator speed in this case was considered to be fixed at 1.1 p.u. due to the big inertia of the wind turbine (working in the hyper synchronous speed). There are many different control strategies for the rotor side converter (RSC) of DFIM, such as vector controls, predictive controls, direct power controls, sliding mode, etc. Many references have addressed different strategies and improved RSC strategies for the low voltage ride through for DFIM when there is a fault in the grid, so as to overcome the problems. However, the objective of this section is to investigate the performance of the

differential current protection. Hence, the RSC was controlled under fixed V_d and V_q commands during the simulations. In this section, the three-phase fault has been considered in all the simulations. The four fault cases that have been studied are the following.

- 1) External three-phase fault between the RSC and the rotor winding: $t = 1\text{--}1.2$ s.
- 2) Internal three-phase fault in the rotor windings: $t = 2\text{--}2.2$ s.
- 3) Internal three-phase fault in the stator winding: $t = 3\text{--}3.2$ s.
- 4) External three-phase fault in the grid: $t = 4.0\text{--}4.2$ s.

The schematic of the simulated DFIM system is shown in Fig. A.5. In all of the cases, the three-phase fault resistance is $0.5\text{ m}\Omega$. The stator active and reactive powers are depicted in Fig. A.6. As can be seen, when there is a fault in the rotor windings or near the RSC (cases 1 and 2), the machine absorbs reactive power and the injected power to grid increases due to the machine fixed speed. The fault cases 1 and 2 have the same effect on the stator side. Therefore, the performance of the stator active and reactive powers for these two cases is almost the same. Hence, the internal and external faults in the rotor cannot be distinguished from the stator active and reactive powers.

Fig. A.5. Performed faults in the DFIM simulations at 1, 2, 3 and 4 s (1. External rotor fault / 2. Internal rotor fault / 3. Internal stator fault / 4. External stator fault).

The three-phase rotor current and the RSC current are displayed in Figs. A.7 and A.8, respectively. The three-phase stator and the grid currents are shown in Figs. A.9 and A.10, respectively. As can be seen, when there is an external or an internal fault in the stator, the rotor current will increase, having both similar values. This indicates that the fault in the stator will induce the same effect in the rotor current whether or not it is external or internal. Moreover, the stator current for internal and external faults in the rotor is the same in both cases, as can be seen in Figs. A.9 and A.10.

Therefore, the internal and external faults cannot be distinguished by only using the rotor or the stator currents. Additionally, according to Figs. A.8 and A.10, the converter currents and the grid currents increase during both internal and external faults. However, this does not clarify which kind of fault has occurred.

Fig. A.6. The stator active and reactive power (MW, Mvar)

Fig. A.7. Rotor currents (kA) during the four faults studied.

Fig. A.8. Rotor side converter currents (kA) during the four faults studied.

Fig. A.9. Stator currents (kA) during the four faults studied.

Fig. A.10. Grid currents (kA) during the four faults studied.

The rotor currents in d and q axes are shown in Figs. A.11 and A.12, respectively. In all these cases, the rotor current in the synchronous frame has big values during the faults and cannot give any information about the fault. The stator currents in dq synchronous frame are shown in Figs. A.13 and A.14, respectively. In these figures, although the d-axis stator current in the fault case 3 has a big value if compared to the other cases, the q-axis stator current has a big value in cases 1 and 2. This high current in the q-axis shows that the stator is absorbing a lot of reactive power from the grid. Figs. A.11–A.14 indicate that the rotor and stator currents increase during the fault in the dq-axis. However, these currents do not show the type of the fault either.

Fig. A.11. Rotor current in d axis I_{rd} (kA) Fig. A.12. Rotor current in q axis, I_{rq} (kA) Fig. A.13. Stator current in d axis, I_{sd} (kA)

Fig. A.14. Stator current in q axis, I_{sq} (kA)

The differential currents in dq synchronous frame are shown in Figs. A.15 and A.16. According to these two figures, the differential current in the synchronous frame is almost zero during normal operation and with the external fault. However, during the internal fault, the differential currents in dq have a big value. Hence, these two figures clearly show that when there is an internal fault in the induction machine (cases 2 and 3), the differential currents are high, and when there is an external fault in the system, the differential currents are small (cases 1 and 4). These two figures obviously indicate that the differential currents of the rotor and the stator in the synchronous frame could be used to protect the DFIMs during internal faults. Moreover, when there is an external fault in the system, such as in cases 1 and 4, the proposed differential protection of the induction machine will not trip.

Fig. A.16 shows that the differential current in the q-axis has a small value during the normal operations. This small current is for the magnetization of the machine. As a result, the proposed differential protection can be implemented in a DFIM to prevent damages inside the machine.

Fig. A.15. Differential current in d axis, I_{diff-d} (kA) Fig. A.16. Differential current in q axis, I_{diff-q} (kA)

A.2.5. Experimental laboratory test

In order to evaluate the effectiveness of the proposed differential protection, some experimental tests were carried out in a 5.5-kW variable speed DFIM. In the experimental setup, shown in Fig. A.17, the machine side converter of the back-to-back converter is connected to the rotor winding of the machine, whereas the grid side converter is connected to the grid by a step-up 130 V/400 V transformer. The stator winding is directly connected to the 400-V grid. The DFIM machine is driven by a dc motor. The two converters are controlled by two different dSPACE DS1104. The RSC is controlled by a model predictive direct power control [6] and the grid side converter (GSC) is controlled using a vector control strategy [25]. The switching frequency of both converters is set to 4 kHz. The machine speed for all the tests was fixed at 800 r/min, the initial active power is 1 kW, and the reactive power is -1.5 kvar. The experimental results were obtained through the interface dSPACE control desk. The parameters of the machine and the setup are summarized in Table A.2.

Fig. A.17. Experimental Setup configuration.

Table A.2. Doubly Feed Induction Machine Data used in the experiments

DC link voltage	250 V	Stator leakage inductance	8.6 mH
Rated power	5.5 kW	Rotor leakage inductance	8.6 mH
Grid voltage	400 V	Stator rated current	13.3 A
Rated stator voltage	400V	Rotor rated current	21A
Rated rotor voltage	182 V	Turns ratio	2.19
Number of pole pairs	3	Inertia	0.16 kg.m ²
Stator resistance	0.65 Ω	Capacitor value in DC bus	4080 μ F
Rotor resistance	2.2 Ω	GSC inductance	10 mH
Mutual inductance	81 mH		

The three-phase and phase-to-phase internal and external faults were made in the stator and in the rotor of the asynchronous machine to investigate the performance of the proposed differential protection. In order to not damage the machine, reasonable resistor values have been used to limit the fault current during the faults. Eight cases have been studied, which are as follows (see Fig. A.18).

- 1) Rotor external: Three-phase fault between the RSC and the rotor winding.
- 2) Rotor internal: Three-phase fault in the rotor windings.
- 3) Stator internal: Three-phase fault in the stator winding.
- 4) Stator external: Three-phase fault in the grid.
- 5) Rotor external: Phase-to-phase fault between the RSC and the rotor winding.
- 6) Rotor internal: Phase-to-phase fault in the rotor windings.
- 7) Stator internal: Phase-to-phase fault in the stator winding.
- 8) Stator external: Phase-to-phase fault in the grid.

Fig. A.18. Simplified Experimental Setup configuration.

A.2.5.1. Three phase Rotor External Fault

In this case, three 9.5-Ω resistors have been placed between the rotor and the converter to be used as a three-phase external fault. The value of the resistors was selected to limit the fault current in order to prevent the breakdown of the converter switches. This fault was connected to the system during regular operation at $t = 6$ s and removed at $t = 14$ s. The stator and rotor currents in abc frame and synchronous frame for this case are shown in Figs. A.19 and A.20, respectively. The differential d and q currents (I_{diff_d} and I_{diff_q}) are shown in Fig. A.21. As can be seen, the differential currents do not change during this fault as it was expected. Hence, the protection should not trip in this case.

Fig. A.19. Stator and Rotor currents during a Rotor External 3 phase fault using a fault resistance of 9.5Ω .

Fig. A.20. Stator and Rotor d and q currents. Rotor External 3 phase fault with a fault resistance of 9.5Ω .

Fig. A.21. Differential d and q currents I_{diff_d} , I_{diff_q} . Rotor External 3 phase fault with a fault resistance 9.5Ω .

A.2.5.2. Three Phase Rotor Internal Fault

In this section, the experimental result of an internal rotor fault is presented in Figs. A.22–A.24. During this test, the three-phase internal fault was connected at $t = 12$ s and removed at $t = 19.5$ s. As can be seen, the rotor current increases in the d-axis of the synchronous frame, whereas the stator current in that same d-axis does not change. As a result, it is clearly observed from Fig. A.24 that the differential d current ($I_{\text{diff_d}}$) increases from nearly 1 A to almost 3 A during this fault. It is obvious that in this case, the protection should trip.

Fig. A.22. Stator and Rotor currents during a Rotor Internal 3 phase fault using a fault resistance of 9.5Ω .

Fig. A. 23. Stator and Rotor d and q currents. Rotor Internal 3 phase fault with a fault resistance of 9.5Ω

Fig. A.24. Differential d and q currents I_{diff_d} , I_{diff_q} during a Rotor Internal 3 phase fault using a fault resistance of 9.5Ω .

A.2.5.3. Three Phase Stator Internal Fault

The third set of figures, which corresponds to an internal fault in the stator, is shown in Figs. A.25–A.27. So as to limit the fault currents, three $135\text{-}\Omega$ resistors have been used as a three-phase fault resistor in the stator side. This fault was connected at around $t = 8\text{ s}$ and removed at $t = 14\text{ s}$. In this case, the protection should trip. It is clear that the rotor current increases during this fault, since the closed-loop controller continues to inject the same active and reactive powers during the fault. Therefore, the stator current does not change, and the fault current is supplied by the rotor side. This increment in the rotor current is in the d-axis of the synchronous frame. Consequently, the differential d current I_{diff_d} increases notably during the fault.

Fig. A. 25. Stator and Rotor currents during a Stator Internal 3 phase fault using a fault resistance of 135Ω .

Fig. A. 26. Stator and Rotor d and q currents. Stator Internal 3 phase fault with a fault resistance of 135 Ω

Fig. A.27. Differential d and q currents I_{diff_d} , I_{diff_q} . Stator Internal 3 phase fault with a fault resistance of 135 Ω .

A.2.5.4. Three-Phase Stator External Faults

Finally, a three-phase external fault in the stator is presented in this section. The results are shown in Figs. A.28–A.30. The control of the machine and the used fault resistance of 135 Ω allow the current supply by the machine to be kept constant. In this case, the protection system proposed shows a very small differential current in both axes, which should not trip the machine.

Fig. A.28. Stator and Rotor currents. Stator external 3 phase fault with a fault resistance of 135 Ω .

Fig. A.29. Stator and Rotor d and q currents. Stator External 3 phase fault with a fault resistance of 135 Ω

Fig. A.30. Differential d and q currents I_{diff_d} , I_{diff_q} . Stator External 3 phase fault with a fault resistance of 135 Ω .

A.2.5.5. Phase-to-phase Faults

Only d and q differential currents (I_{diff_d} and I_{diff_q}) are presented in phase-to-phase tests. In the case of phase-to-phase faults, only two 9.5 and 135- Ω resistors between phases A and B are used to limit the fault current between rotor and stator, respectively.

In the cases of external faults in the rotor (see Fig. A.31) and stator (see Fig. A.32), the differential currents do not change during these faults and the protection does not trip, as it was expected. On the other hand, in case of internal faults in the rotor (see Fig. A.33) and stator (see Fig. A.34), the differential current in the d-axis increases by almost three times. Therefore, the protection will trip in these cases. The results of phase-to-phase faults are similar to the cases of three-phase faults.

Fig. A.31. Differential d and q currents I_{diff_d} , I_{diff_q} . Rotor External 2 phase fault with a fault resistance 9.5 Ω .

Fig. A.32. Differential d and q currents I_{diff_d} , I_{diff_q} . Rotor Internal 2 phase fault with a fault resistance 9.5 Ω .

Fig. A.33. Differential d and q currents I_{diff_d} , I_{diff_q} . Stator Internal 2 phase fault with a fault resistance of 135Ω .

Fig. A.34. Differential d and q currents I_{diff_d} , I_{diff_q} . Stator External 2 phase fault with a fault resistance of 135Ω .

A.2.6. Conclusions

The protection of the rotor of a DFIM is not an easy task due to the low frequency of the rotor current during normal operation. Hence, the usual protection for the rotor cannot be effective and fast due to the variable low frequency. However, this section presents a simple and practical differential protection for the DFIM based on the comparison of the I_d and I_q of the rotor and the stator, which considers the low frequency of the rotor current. The results of both the simulation and the experimental tests performed show that the proposed protection can effectively protect the rotor and the stator windings during internal faults but without tripping in case of external faults or normal operation. The setting of this protection should be carefully studied. The protection should only trip in case of internal fault, but not in case of external faults or normal operation. The tripping threshold should be calculated to avoid unwanted trips produced by several reasons as saturation of the current sensors, saturation of the magnetizing inductance, error in the turn ratio measurement, or power system voltage unbalance among others.

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