DOUBLE STAR PM MACHINE: ANALYSIS AND SIMULATIONS

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Acronyms and Symbols

*PWM*: pulse width modulation

*VSI*: voltage source inverter

*DTC*: direct torque control

*DC*: direct current

*AC*: alternating current

*MMF*: magneto motive force

*BDCM*: brushless DC machine

*IGBT*: insulated gate bipolar transistor

*PM*: permanent magnet

*DFPM*: doubly fed permanent magnet motor

*EMF*: electromotive force

*SVPWM*: space vector pulse width modulation

*FOC*: field oriented control

*MTPA*: maximum torque per ampere method

*PMSM*: permanent magnet synchronous machine

*IPM*: interior permanent magnets

*SPM*: surface permanent magnets

*CSI*: current source inverter

*DFIG*: doubly fed induction generator

*WT*: wind turbine

*TSR*: tip speed ratio

*TF*: transfer function

*PI*: proportional integral regulator
\( V \), : voltage and current, respectively

\( n \): number of phases of an electrical machine

\( \alpha \): displacement between any two consecutive stator phases

\( k, i \): number of windings and number of sub phases

\( N \): number of three phase winding sets

\( N_s \): number of series connected turns per phase

\( \Phi_p \): per pole flux

\( \Phi_g \): air gap flux

\( K_w \): winding factor

\( f \): frequency of electrical quantities \((Hz)\)

\( L \): useful core length

\( D \): diameter of the air gap

\( p \): number of poles

\( n_p \): number of pole pairs

\( n_{rpm} \): rotor rotational speed \( rpm \) of the electrical machine

\( N_t \): number of series connected turns per coil

\( q \): number of slots per pole per phase

\( b \): number of parallel path per phase

\( \tau_s \): slot pitch

\( \sigma_s \): current density

\( \lambda_s \): electric loading

\( C \): output coefficient

\( P, P_m \): electrical (active) and mechanical power, respectively

\( A_t, A_s \): turn and slot cross section area

\( K_f \): fill factor of a stator slot

\( a_i, b_i, c_i \): phases’ name in the \( i^{th} \) winding set

\( I_{sc} \): short circuit current

\( E_0, E_{L2L} \): no load EMF phase and line to line value, respectively

\( \omega_e \): electrical angular speed of the rotor \((el. \ rad/s)\)

\( R_s \): stator resistance

\( L_s \): stator inductance

\( L_{sd} \): leakage inductance
\( L_{sm} \): excitation inductance
\( T_{\text{max}} \): maximum torque
\( T_e \): electromagnetic torque
\( O_1, O_2 \): neutral point of the first and second winding set respectively
\( T_m \): mechanical resistance torque
\( \Phi_e, \Phi_a \): excitation and armature flux, respectively
\( \gamma \): angle between excitation and armature flux
\( B \): magnetic induction
\( H \): magnetic field intensity
\( \mu_0 \): absolute permeability
\( A_m, h_m, V_m, U_m \): area, length, volume and magnetic voltage of the permanent magnet
\( A_g, h_g, V_g, U_g \): area, length, volume and magnetic voltage of the air gap
\( \Psi_{PM} \): permanent magnets flux
\( L, M \): self and mutual inductance
\( \omega_m \): mechanical angular speed of the rotor (\text{rad/s})
\( \theta_e \): electrical angle
\( \theta_m \): mechanical angle
\( \alpha - \beta \): stationary reference frame
\( d - q \): reference frame synchronous with rotor position
\( S \): wind passage area in wind turbine
\( v \): wind speed
\( \Omega_r, n_r \): angular speed of wind turbine’s blades (\text{rad/s} and \text{rpm}, respectively)
\( D_b, r \): diameter and radius of the blades of the wind turbine
\( \rho \): air density
\( c_p \): power coefficient
\( \lambda \): tip speed ratio
\( \beta \): pitch angle
\( m_a, m_f \): amplitude modulation ratio and frequency modulation ratio
\( f_h, h \): frequency of harmonic component and harmonic order respectively
\( J \): inertia
\( B_v \): viscous friction coefficient
\( K_p, K_i \): PI regulator parameters
$g_1, g_2$: gate control signals
$V_{DC}$: DC bus voltage
$I_{DC}$: DC side current
$\eta$: efficiency

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Abstract

The work deals with multiphase machines, which are electrical machines with a number of phases greater or equal than 3. These machines are nowadays quite common and a brief presentation of general multiphase machines theories are illustrated. Due to development of power electronic components and materials improvement, the use of multiphase machine drives became easier and their particular features returned to be studied, particular attention is drown by split phase machines. Main focus of the thesis is the so called double star permanent magnets synchronous machine, which is a particular split phase machine, constructed with two three phase winding sets. Recently, there are more systems which employ multiphase machine drives, some of them are analyzed and discussed. A part of the work is dedicated to the mathematical model of the double star PMSM. Problems arise in the insertion of a number of three phase sets greater than one, due to additional magnetic coupling between winding sets with a consequent complication in the development of a control system logic. It will be shown that using a particular Park’s transformation a fictitious model will be obtained which introduces a very important simplification in the machine modeling. Then the machine will be tested to verify the correctness of machine model and control system, different system configurations are illustrated with attention to the main system variables. Additionally, a possible newer system configuration is presented and explained.
Summary

Chapter 1 is a brief introduction of multiphase machines’ generalities. In particular, an introduction about their sizing equation stator and rotor windings, advantages and disadvantages arises by adding a number of machine phases greater or equal than three.

Chapter 2 illustrates some applicative cases in the use of the double star $SM$. Its use is spread in many application areas such as: more electric aircraft, ship applications, automobile and rail traction drives.

Chapter 3 Shows the mathematical model of the double star $PMSM$. Phase variables model is transformed using a particular Park’s transformation and obtained results, such as: decoupling, torque expression simplification, possible use of general three phase machine theory are presented and discussed.

Chapter 4 Brief introduction to wind turbine systems: maximum extractable wind power, possible system configuration, with particular attention to $PMSM$ system configuration. Some power converter’s control strategies, such as: $PWM$ and space vector modulation are presented.

Chapter 5 Simulations based on a real machine data are carried out in different system configurations. Most important, wind turbine system simulation which shows the main part of the electrical drive and steady state quantities.
Chapter 1

Introduction

In this chapter will be presented an overview of existing multiphase machine technologies with particular attention on their advantages and construction. Also, a brief history of reasons which brought interest in this kind of machines is presented along with innovations that made possible their use and applications in modern electrical systems. Their main features, such as: fault tolerance, energy segmentation and performances are presented and discussed.

1.1 Historical reasons of multiphase machines

The roots of multiphase variable speed drives can be traced back to the late 1960s, the time when inverter-fed ac drives were in the initial development stage.

Due to the six step mode of three-phase inverter operation, one particular problem at the time was the low frequency torque ripple. Since the lowest frequency torque ripple harmonic in an \( n \)-phase machine is caused by the time harmonics of the supply of the order \( 2n \pm 1 \), increase number of phases seem to be the best solution to avoid the problem. Hence, significant efforts have been put into the development of five phase and six-phase variable-speed drives supplied from both voltage source and current source inverters.

This is an advantage of multiphase machines that is nowadays somewhat less important since pulse width modulation (PWM) of voltage source inverters (VSI) enables control of the inverter output voltage harmonic content. The other main historical reasons for early developments of multiphase drives, are better fault tolerance and the possibility of splitting the motor power (current) across a higher number of phases. By increasing the number of phases it is also possible to increase the
torque per $rms$ ampere for the same volume machine. Improvement of noise characteristics and reducing the stator copper loss are other advantages of multi-phase systems. The pace of research started accelerating in the second half of the 1990s, predominantly due to the developments in the area of electric ship propulsion, which remains nowadays one of the main application areas for multiphase variable-speed drives.

Other applications of this kind of machines are: locomotive traction, industrial high-power applications, electric and hybrid-electric vehicles (propulsion, integrated starter/alternator concept, and others), the concept of the “more-electric” aircraft and wind turbine generation. Most of the recent works related to these applications are that typically high-performance motor control is utilized, which are vector control or direct torque control ($DTC$).

1.2 About power converters

Variable speed $AC$ drives are nowadays invariably supplied from power electronic converters. Since the converter can be viewed as an interface that decouples three-phase mains from the machine, the number of machine’s phases is not limited to three any more. As an example can be seen Figure 1.1 in which is well shown an advantage introduced by development of power electronic converters. In any case, three phase machines are the most utilized. Such a situation is expected to persist in the future.

Figure 1.1: possible configuration with multiphase machines
The power rating of the converter should meet the required level for the machine and driven load. However, the converter ratings cannot be increased over a certain range due to the limitation on the power rating of semiconductor devices. One solution to this problem is using multi-level inverter where switches of reduced rating are employed to develop high power level converters. Multi-phase machines can be used as an alternative to multi-level converters. In multi-phase machines, by dividing the required power between multiple phases, more than the conventional three, higher power levels can be obtained and power electronic converters with limited power range can be used to drive the multi-phase machine. Whether it is better to use multi-phase machines or multi-level converters is debatable and in fact it is extremely application dependent. Insulation level is one of the limiting factors that can prohibit the use of high voltage systems. Therefore, multi-phase machines that employ converters operating at lower voltage level are preferred.

1.3 Types of multiphase machines

The types of multiphase machines for variable-speed applications are in principle the same as their three-phase counterparts. There are induction and synchronous multiphase machines, where a synchronous machine may be with permanent magnet excitation, with field winding, or of reluctance type. Three phase machines are normally designed with a distributed stator winding that gives near-sinusoidal MMF distribution and supplied with sinusoidal currents. (the exception is the permanent magnet synchronous machine with trapezoidal flux distribution and rectangular stator current supply, known as brushless DC machine, or simply BDCM) Nevertheless, spatial MMF distribution is never perfectly sinusoidal and some spatial harmonics are inevitably present. Multiphase machines show more versatility in this respect. A stator winding can be designed to yield either near-sinusoidal or quasi-rectangular MMF distribution, by using distributed or concentrated windings, for all ac machine types. Near sinusoidal MMF distribution requires use of more than one slot per pole per phase. As the number of phases increases it becomes progressively difficult to realize a near-sinusoidal MMF distribution. For example, a five-phase four-pole machine requires a minimum of 40 slots for this purpose, while in a seven-phase four-pole machines at least 56 slots are needed (for a three-phase four-pole machine the minimum number of slots is only 24).

In both stator winding designs, there is a strong magnetic coupling between the stator phases. If the machine is a permanent magnet synchronous machine, then concentrated winding design yields a behavior similar to a BDCM. A permanent magnet multiphase synchronous machine can also be of
so-called modular design where an attempt is made to minimize the coupling between stator phases. (a three-phase permanent magnet machine may be designed in the same manner, but the most important benefit of modular design, fault tolerance, is then not exploited to the full extent). It should be noted that the spatial flux distribution in permanent magnet synchronous machines is determined by the shaping of the magnets.

1.4 Stator windings

Stator windings of an $n$-phase machine can be designed in such a way that the spatial displacement between any two consecutive stator phases equals $\alpha = \frac{2\pi}{n}$, in which case a symmetrical multiphase machine results. This will always be the case if the number of phases is an odd prime number. This winding topology requires a non-conventional $n$-phase inverter (and control strategy) to be used for motor supply.

![Figure 1.2: types of stator slots(open slots)](image)

However, if the number of phases is an even number or an odd number that is not a prime number, stator winding may be realized in a different manner, as $k$ windings having $i$ sub phases each (where $n = i \cdot k$). In such a case, the spatial displacement between the first phases of the two consecutive sub phase windings is $\alpha = \frac{\pi}{n}$, leading to an asymmetrical distribution of magnetic winding axes in the cross section of the machine (asymmetrical multiphase machines).
A multi-phase winding topology which is very commonly used in high-power electric machinery is the so-called “split-phase” configuration. As illustrated in Figure 1.3, this results from splitting the winding into $N$ three-phase sets, displaced by $\frac{60/\pi}{N}$ electrical degrees apart. When the machine is used as a motor, a split-phase winding arrangement can be desirable as it allows for $N$ three-phase conventional inverter modules to be used for its supply. Some of the advantages of multiphase machines, when compared to their three-phase counterparts, are valid for all stator winding designs while others are dependent on the type of the stator winding.

1.4.1 Machines with sinusoidal winding distribution

This kind of machines are characterized with the following features:

- Fundamental stator currents produce a field with a lower space-harmonic content.
- The frequency of the lowest torque ripple component, being proportional to $2n$, increases with the number of phases.
- Since only two currents are required for the flux/torque control of an AC machine, regardless of the number of phases, the remaining degrees of freedom can be utilized for other purposes.

One such purpose, available only if the machine is with sinusoidal $MMF$ distribution, is the independent control of multi-motor multiphase drive systems with a single power electronic converter supply. As a consequence of the improvement in the harmonic content of the $MMF$, the noise emanated from a machine reduces and the efficiency can be higher than a three-phase machine.
1.4.2 Concentrated winding machines

In concentrated winding machine a possibility of enhancing the torque production by stator current harmonic injection exists. Given the phase number \( n \), all odd harmonics in between one and \( n \) can be used to couple with the corresponding spatial \( MMF \) harmonics to yield additional average torque components.

This possibility exists if the phase number is odd, while the only known case where the same is possible for an even phase number is the asymmetrical six-phase machine with a single neutral point. Torque enhancement by stator current harmonic injection is one possible use of the additional degrees of freedom, offered by the fact that only two currents are required for flux and torque control due to the fundamental stator current component. The table below shown other possible ways to use additional degrees of freedom.

![Figure 1.4: four phases windings machine](image)

<table>
<thead>
<tr>
<th>Stator winding type</th>
<th>Fault-tolerant operation</th>
<th>Torque enhancement by stator current higher harmonic injection</th>
<th>Multi-motor drives with single inverter supply</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sinusoidally distributed</td>
<td>Yes</td>
<td>N/A</td>
<td>Yes</td>
</tr>
<tr>
<td>Concentrated (distributed)</td>
<td>Yes</td>
<td>Yes</td>
<td>N/A</td>
</tr>
<tr>
<td>Modular design</td>
<td>Yes</td>
<td>Possible (if magnets give quasi-rectangular spatial flux distribution)</td>
<td>N/A</td>
</tr>
</tbody>
</table>

Table 1: Uses of additional degrees of freedom

1.5 Sizing equations for multi-phase machines

Although characterized by a variety of possible phase arrangements, multi-phase windings can be treated in the same way from machine sizing viewpoint, under the only hypothesis that each pole span encompasses exactly as many phase belts as the phases are \( (n) \). This hypothesis is verified in the vast majority of multi-phase designs; only those designs are not covered where successive phases are shifted by \( rac{360}{n} \) electrical degrees in space, as happens in symmetrical windings with an even number of phases.

The rated phase voltage is given by:

\[
V = \frac{1}{\sqrt{2}} \Phi_p N_s K_w (2\pi f)
\]  

(1.1)
where $\Phi_p$ is the flux per pole, $N_s$ the number of series-connected turns per phase, $K_w$ the winding factor and $f$ the rated frequency. Quantities $\Phi_p$, $N_s$ and $f$ can in turn be expressed as follows:

\[
\Phi_p = B_g L \pi D / (2n_p) \quad (1.2)
\]

\[
N_s = q (2n_p) N_t / b \quad (1.3)
\]

\[
f = n_{RPM} n_p / 60 \quad (1.4)
\]

In terms of: the average flux density $B_g$ in the air-gap, the useful core length $L$, the machine average diameter $D$ at the air-gap, the number of pole pairs $n_p$, the number of slots per pole per phase $q$ (possibly fractional), the number of series-connected turns per coil $N_t$, the number of parallel paths per phase $b$ and the speed $n_{RPM}$ in revolutions per minute.

A further design figure, called the output coefficient $C$, can be also introduced to describe the degree of utilization of the machine volume (roughly proportional to $D^2L$) in terms of useful machine torque (proportional to $P / n_{RPM}$, where $P$ is the rated active power):

\[
C = P / (n_{RPM} D^2L) \quad (1.5)
\]

Substituting equations from 1.1 to 1.4 into 1.5 yields:

\[
V = \frac{\pi^2 \sqrt{2} (2n_p) P B_m K_w q N_t}{120 b D C} \quad (1.6)
\]

Knowing that the slot pitch expression is:

\[
\tau_s = \frac{\pi D}{q n (2n_p)} \quad (1.7)
\]

And introduce it into equation 1.6:

\[
V = \frac{\pi^2 \sqrt{2} P B_m N_t K_w}{120 b C n \tau_s} \quad (1.8)
\]

The slot pitch can be alternatively expressed in terms of current density $\sigma_s$ and the electric loading $\lambda_s$:
\[
\tau_s = \frac{2N_t A_t \sigma_s}{\lambda_s}
\]  

(1.9)

The two following equations are the applied definitions for \(\sigma_s\) and \(\lambda_s\):

\[
\sigma_s = \frac{l/b}{A_t} = \frac{2N_t (l/b)}{A_s K_f}
\]

(1.10)

\[
\lambda_s = \frac{2N_t (l/b)}{\tau_s}
\]

(1.11)

Where \(A_s\) and \(K_f\) are the cross section area and the filling factor of a stator slot. It can be finally find:

\[
\frac{P}{V} = k \left( \frac{C\sigma_s K_f}{B_m \lambda_s K_w} \right) \left( \frac{A_s}{N_t} b n \right)
\]

(1.12)

where \(k\) is a non-dimensional constant whose value only depends on the units used to express the other quantities. Coefficient in the first brackets does not depend on the winding structure, but only on the magnetic, thermal and electrical loading of the machine; therefore, for machines of homogeneous design in terms of thermal class, insulation technology, cooling system effectiveness, etc., first brackets can be regarded as a constant to a good approximation level. Hence, equation 1.12 expresses the explicit relationship between the following design quantities:

- machine power \((P)\) and voltage \((V)\) ratings;
- winding structure in terms of slot cross-section area \((A_s)\), number of turns per coil \((N_t)\), number of phases \((n)\) and number of parallel ways per phase \((b)\).

Equation 1.12 shows that if the power rating \(P\) increases while the voltage \(V\) below a certain level, this naturally leads to decrease the number of turns per coil \(N_t\), which may result in the need for Roebel bars \((N_t=1)\) above a given power level. Equation 1.12 also demonstrates that there are three design “levers” available to counteract the decrease of \(N_t\), namely:

- increasing the slot cross-section area \(A_s\);
• increasing of the number $b$ of parallel ways per phase;
• the increase of the number of phases $n$.

The first strategy is of limited help, since it generally implies a growth of the overall machine size.
The second strategy can be actually pursued until the number of parallel ways $b$ equals $2p$ since the
number of parallel ways cannot exceed the number of machine poles in any case.
Hence, it is easily understood that, after the limit $b = 2p$ has been reached, the only way left to
avoid the use of Roebel bar construction without incrementing the machine size consists of
increasing the number of its stator phases $n$.

1.6 Advantages of multiphase machines

As it was already said, many are the reasons which brought interest in electrical machines with a
number of phases greater than 3. Now they will be presented in more detail, while in chapter 2 they
will be examinated with some application examples.

1.6.1 Electric power segmentation

The basic reason why, in high-power multiphase machine design practice, it is often necessary to
move from an ordinary three-phase concept to an $n$ -phase one (with $n$ higher than three) is
illustrated in Figure 1.5. It can be seen in Figure 1.5a that in a three phase drive required to deliver
a mechanical power $P$ megawatts to the load, apart from losses, the same power $P$ is to be supplied
by the feeding inverter, each phase of which is thereby demanded to deliver a power $P/3$. When $P$
exceed some certain limits, the available power electronics technology may become inadequate to
achieve such power rating for a single inverter phase. This may be due to single power switch
component current or voltage capability limits or to the limit in the number of series-connected or
shunt-connected switches that can be included in a phase. As a consequence, it may be come
mandatory to split the overall inverter-supplied power $P$ into a higher number of phases. Such a
power segmentation can be achieved, for example, using multiple ($N$) three-phase inverters, each
rated $P/N$ megawatts, instead of a single converter (as shown in Figure 1.5b) or still using a single
inverter but equipped with $n$ phases ($n > 3$), each carrying $P/n$ megawatts.
In high power applications, the solution illustrated by Figure 1.5b (based on a “multiple star” machine) is the most widespread because it enables the designer to use existing and proven three-phase inverter units combined together instead of developing new polyphase topology with the relevant control algorithms.

In fact, in high power applications, where project risk management issues play an important role throughout system development because of the high investment or project costs, the possibility to rely on individually proven and assessed subsystem is often regarded as a preferable option.

This does not exclude that other multiphase topologies can be implemented where, for example, multiple \( N \) symmetrical \( n \)-phase inverters feed a motor equipped with \( N n \)-phase stator windings.
In Figure 1.6a, an example of a 19 MW 15-phase ship propulsion induction motor fed by three 5-phase inverters is shown.

1.6.2 Reliability and fault tolerance

Multiphase design guarantees a higher system reliability and fault tolerance. In fact, in case of a faulty phase, the multiphase system is capable of continuing operation, even without changes in control system strategies, although with degraded operation and at reduced power. It is important for safety but also in those cases where a drive trip and the consequent driven equipment stop causes important economic losses due to production discontinuity.

It is also intuitive and experimentally proven that the higher the number of stator phases the less the degradation and the power de-rating that is to be expected following a fault on a machine phase. Therefore, increasing the number of phases is a provision which generally increments system fault tolerance, in the sense that it reduces the effect of the fault in terms of machine performance.

1.6.3 Performances

It is well known from multiphase machine classical theory that increasing the number \( n \) of stator phases enhances the harmonic content of the air-gap flux density field, making its waveform closer and closer to the sinusoidal profile as \( n \) grows. This can be easily explained because the harmonic rotating fields sustained by different phase sets in a multiple star machine undergo a beneficial mutual cancellation effect.

The benefits which originate for the better air-gap flux waveform due to a high number of stator phases are mainly the following:

- Reduction of rotor losses due to flux pulsations and consequent induced eddy currents in rotor circuits (field, dampers if present) and permanent magnets (if present).
- Improvement of torque quality for reduced amplitude and increased frequency of torque pulsations.

The former benefit is especially important for high-speed multiphase electric machines equipped with permanent magnet rotors (permanent magnet eddy current losses tend to increase as the speed grows). The latter benefit is crucial in those applications where the multiphase machine is subject to strongly distorted phase currents and limits are imposed on the maximum allowed torque
ripple. This is the typical case of synchronous machines supplied by load commutated inverters. The mutual cancellation effects guaranteed by multiphase winding topologies is also important in multiple-star synchronous machines fed by PWM inverters, where considerable phase current harmonic distortions can be accepted without any detrimental effect on the motor torque quality.
Chapter 2

Double star synchronous machine’s applications

In the second chapter main applications of multiphase synchronous machine are presented. It will be shown that this kind of machine represents a very reliable and efficient solution in many applications. Its features, such as: reliability, fault tolerance, higher power quality and high power density are highlighted and discussed. Wind turbine’s applications will be discussed in detail in chapters 4 and 5.

2.1 Electric Drive System of Dual-Winding Fault-Tolerant Permanent-Magnet Motor for Aerospace Applications

In most applications, the failure of a drive has a serious effect on the operation of the system. In some cases, the failure results in lost production whereas in some others it is very dangerous to human safety. Therefore, in life dependent application it is of major importance to use a drive which continues operating safely under occurrence of a fault. The major faults which can occur within a machine or converter are considered as:

- winding open circuit
- winding short circuit (phase to ground or within a phase)
- winding short circuit at the terminals
- power device open circuit, power device short circuit and
- DC link capacitor failure.
In order to limit the short circuit current, the machine should have a sufficiently large phase inductance and in order to avoid loss of performance in healthy phases in faulty condition, mutual inductance between the phases should be small. These two points are required for the reliability of the system.

With the increasing development of aircraft, new more electric and all-electric aircraft in the aviation sector have attracted increasingly more attention. The typical characteristic of the more-electric or all-electric aircraft is that part or all pneumatic and hydraulic systems are replaced by electrical drive systems, which can reduce the running cost, aircraft’s volume and weight, and fuel cost and improve the reliability and maintainability of the aircraft. It is estimated that the weight and fuel cost of an aircraft can be reduced by 10% and 9% adopting more-electric and all-electrical aircrafts, respectively. More-electric aircraft is a transitional scheme from conventional aircraft to all-electric aircraft. At present, the mainly more-electric aircraft includes the European Airbus A380, American Boeing B787 and Lockheed Martin F – 35.

The current research shows that the design of the electrical drive system is one of the key technologies of more-electric aircraft. Compared with the conventional hydraulic fuel pump system, the electrical drive system can not only improve the system efficiency and the flexibility of variable speed control but also reduce the weight and volume of the system in the aircraft fuel pump system. In the aircraft fuel pump system, where continuous operation must be ensured, reliability may be a critical requirement. As we know, the motor and its drive system are the core parts of the electrical drive system, therefore, in addition to meet some specific functions, the motor drive system for aerospace applications must have high reliability and strong fault tolerance. Redundancy technology is a method to improve the reliability of the motor drive system by adding extra resources such as hardware or software. The parallel dual redundancy motor is the mostly used redundancy motor which consists of two sets of independent windings with 30 electrical degree shift in space, two sets of position sensors, and a mutual rotor.

The topology configuration is shown in the Figure 2.1 in which a high fault-tolerant and electromagnetic performance $DFPM$ motor and two sets of three-phase full-bridge drive circuits are adopted in an electrical drive aircraft fuel pump system.
Full rated torque can be provided by isolated $a_1b_1c_1$ windings or isolated $a_2b_2c_2$ windings under fault conditions. Each redundancy consists of an electrically isolated three-phase full-bridge drive and a regenerative energy dump circuit. The drives are composed of many single 600V IGBTs of Microsemi Corporation, and each redundancy is supplied by a separate 270V, as shown in Figure 2.1.

Compared with the existing fault-tolerant PM motor drive systems in which each phase winding was driven by $H$-bridge circuit, the drive system of the proposed DFPM motor can not only reduce the number of power switches and save the system cost but also improve the reliability and power density of the system.

To ensure a limited value of the short-circuit current of the DFPM motor, it must be increase the self-inductance of the windings. Because the DFPM motor has the characteristics of magnetic isolation and the mutual inductance between adjacent phase windings is very small, the steady state short-circuit current $I_{sc}$ is given by:

$$I_{sc} = \frac{E_0}{\sqrt{(\omega_e L_s)^2 + R_s^2}}$$  \hspace{1cm} (2.1)

$$L_s = L_{sm} + L_{s\sigma}$$  \hspace{1cm} (2.2)

where $E_0$ is the no-load back electromotive force (EMF), $R_s$ and $L_s$ are the phase resistance and phase self-inductance, respectively. $\omega_e$ is the electrical angular velocity of the rotor; $L_{sm}$ and $L_{s\sigma}$ are the excitation inductance and the leakage inductance of the motor, respectively. By adopting fractional slot, i.e., deep and narrow slot, the leakage inductance is increased, and the short-circuit current will be reduced and limited.
In addition, the **DFPM** motor has a big overload ratio, if the rated torque of the **DFPM** motor is $T_{e \_nom}$, the maximum torque capability of the **DFPM** motor are $T_{max}$ and $T'_{max}$ under normal and fault conditions, which can be expressed as follows:

$$T_{max} = 3T_{e \_nom}$$  \hspace{1cm} (2.3)

$$T'_{max} = \frac{3T_{e \_nom}}{2}$$  \hspace{1cm} (2.4)

The control block diagram of the proposed **DFPM** motor is shown in Figure 2.3, in which the active–active redundant fault tolerant control strategy based on the failure diagnosis and redundancy communication and the two sets of three-phase full-bridge drive circuits, namely, inverter 1 and inverter 2, are adopted. In addition, the two sets of independent redundancy control strategy consist of the speed controller, the current controller, space vector pulse width modulation (SVPWM), $abc/dq$ conversion, and the inverter.

![Control Block Diagram](image)

**Figure 2.3**: control scheme for fault diagnosis

Motors and power converters have the highest failure rate in the aerospace control system. The two main faults among the highest failure rate per flight hour are open-circuit fault and short-circuit fault in phase winding. The fault diagnosis methods under open-circuit fault and short circuit fault conditions are shown in Table 2.
2.1.1 Open-circuit and short-circuit condition

Figure 2.4a shows the simulated results of the open-circuit fault in the first winding set, when the DFPM motor operated at the rated load of 20 Nm and a speed of 2000 rpm. When the open-circuit fault occurs in the first winding set at 0.02s, the current waveforms of $a_2$, $b_2$ and $c_2$ windings and the current waveforms of $a_1$, $b_1$, and $c_1$ windings are shown in the first and second plot of figure 2.4a, respectively. It is shown, that each set of windings provided 50% power and the peak current of each phase is 26 A before 0.02s. After the open circuit fault occurs in the first winding set, the phase current of $a_2$, $b_2$ and $c_2$ windings reduced to 0 A, while the phase current of $a_1$, $b_1$, and $c_1$ windings are twice as the original one. Its peak value of current in each phase was 52 A, and it provided 100% power to ensure that the output power of the proposed electric drive system keeps constant. The third plot of figure 2.4a shows the corresponding torque waveforms of $a_2b_2c_2$ windings when the open-circuit fault occurs in the first winding set. It can be seen that the output torque generated by $a_2b_2c_2$ windings was 10Nm before 0.02s, which is only half of the rated load. After the open-circuit fault occurs in the first winding set, the normal phase $a_2$, $b_2$ and $c_2$ windings will output the whole rated load power, and the output torque was 20Nm. The last two plots of Figure 2.4a show the output torque and speed waveforms of the DFPM motor at pre- and post-open circuit fault condition. It can be seen that the output torque and speed are all kept constant, which verify the purpose of the control strategy when the proposed electric drive system of the DFPM motor operates at open-circuit fault condition.

Figure 2.4b shows simulated results of the short-circuit fault in the first winding set, when the DFPM motor operated at the load is 5Nm and the speed is 2000 rpm. When the short circuit fault occurs in the phase- A winding at 0.15s, the current waveforms of $a_2$, $b_2$ and $c_2$ windings and the current waveforms of $a_1$, $b_1$ and $c_1$ windings are shown, again, in the first two plots of Figure 2.4b.
It is shown that the peak value of short-circuit current of $a_1$, $b_1$ and $c_1$ fault windings was nearly 50 $A$, when the system came to steady state. Because 50 $A$ was close to the limited value of the DFPM motor’s short circuit current, it can be seen that the system had the function of inhibiting the short-circuit current.

Meanwhile, the peak current of $a_2$, $b_2$ and $c_2$ windings also increased greatly and stable at 25 $A$. The third plot shows the torque waveforms of $a_2b_2c_2$ windings, when the short-circuit fault occurs in the first winding set. It can be seen that the output torque of $a_2b_2c_2$ windings was 2.5Nm before 0.15$s$, which is only half of the load. After the short-circuit fault occurs in the phase- $A$ winding, the normal phase $a_1$, $b_1$ and $c_1$ windings will compensate for the absent torque of the removed fault phases and offset the pulsating torque of the short-circuit phases. The last two plots of Figure 2.4b show the output torque and speed waveforms of the DFPM motor at pre- and post-short-circuit fault condition. It can be seen that the proposed electric drive system can still operate steadily after a short pulsation, which verifies the proposed fault-tolerant control strategy when the proposed electric drive system of the DFPM motor operates at short-circuit fault condition.

Figure 2.4: (a) open circuit fault (b) short circuit fault
### 2.2 Split-phase machine for electric vehicles applications

In traction systems, split-phase electric motors, can be used with reduced power by operation of one inverter in the case of fault condition. For vehicles using grid power to charge the battery, traction circuit components are not engaged during the charging time, so there is a possibility to use them in the charger circuit to have an on-board integrated charge. This integration leads to reduction in the charger weight, space, and cost in the electrical part of the vehicle. Two independent inverters are utilized to control the split-phase motor with isolated neutral points. By connecting a single-phase AC source to the motor neutral (star) points, the traction system is used as a non-isolated battery charger. The charger is a bidirectional converter with unity power factor operation capability in this configuration.

![Figure 2.5 traction/charge circuit in electric vehicle](image)

For the charging mode, the motor windings are used as inductors, i.e. as energy storage devices. The upper and lower switches in each inverter need to switch simultaneously for a proper operation, that means each inverter is used as a one leg switching device. The system equivalent circuit is presented for the charging mode and it has been shown that the motor does not provide any torque during the charge time. So the motor is in stand still during charge time and there is no need for a mechanical parking brake.

In the traction mode, the two inverters power the motor by using FOC. The system DC bus and two phase currents of each three-phase windings are measured by the controller. The speed control is performed based on the maximum torque per ampere method \( MTPA \).
For battery charging, the source is connected to the motor two neutral points, $O_1$ and $O_2$, by a plug. The motor windings are used as inductors to form a single-phase full-bridge Boost rectifier by using the available power semiconductors in the inverters. Since the component power rating are usually high for traction purposes, the charger is a high-power non isolated battery charger. This is a very efficient and reliable application of split phase electrical machine.
2.3 Ship applications: electric and hybrid electric propulsion

Multiphase machines are also used in ship’s power systems, in recent years, electric and hybrid propulsion is more important due to advantages in electric machine utilization. It’s useful to classify the systems, in particular how is generated the electrical power and how it is used for electric propulsion. As for electrical power production we can have:

- Diesel-electric propulsion: electrical energy is generated by a diesel prime mover.
- Turbo-gas-electric or turbo-electric: when the prime mover is a gas turbine.
- Electrochemical generators: in which the electrical energy is generated by an electrochemical system which uses hydrogen as fuel and oxygen as oxydising, eventually oxygen can be withdraw from air.

The other classification is about the propulsion system itself, so it’d be:

- DC motor propulsion systems
- Induction motors driven by controlled commutation converters
- Synchronous machines driven by cycloconverters
- Synchronous machines driven by synchro-converters

As an example it can be shown the principle circuit of the class T2 tanker, they were construct in the world war 2 by USA and they were drive by a large synchronous machine of variable frequency, from 0 to 60 Hz.
The prime mover was a steam turbine (today steam turbines are abandoned), and as it can be seen from the Figure 2.7 it fed all the electric circuit in the ship, while another small steam turbine fed the excitation of the machine and to all the others 115V circuits.

Another example of electric propulsion is the Queen Elizabeth ship which was fed by steam turbine until the 1987, year in which was decided to use diesel motor instead of steam turbine. Basically, the propulsion electric system is shown in Figure 2.8.

![Figure 2.8: Queen Elizabeth ship principle scheme](image)

In this configuration, the converters were used until 72 rpm of the helix was reached, and then they were short circuited. This ship was very important because is a conjunction between past electric propulsion ships and modern one, in which is always present a converter to regulate the frequency. After a brief presentation of history of synchronous motor drives used in ships applications it will be explain modern uses of double star synchronous machine in cycloconverter and synchro-converter systems.
2.3.1 Cycloconverter systems

A cycloconverter is an electrical device which converts single phase or three phase alternating current into variable magnitude and variable frequency single phase or three phase AC.

![Figure 2.9: block diagram a cycloconverter](image1)

Cycloconverters are used in high power applications driving induction and synchronous motors. They are usually phase-controlled and, traditionally, they use thyristors due to their ease of phase commutation.

In the case of three phase cycloconverter which is driving a synchronous machine, the system will be the one presented in Figure 2.10 in which, a standard synchronous machine was used for propulsion purpose.

![Figure 2.10: three phase cycloconverter](image2)

To improve harmonic content of primary current it shown that can be use a 12 pulse bridge which drives a double star synchronous machine through cycloconverters, so the system can be represented by the Figure 2.11. The use of a double star synchronous machine (in this case with independent excitation) improves reliability and power quality. Cycloconverters allow to regulate
voltages and currents on the "DC" side of the converters. This solution has a high cost but it is justified by the high power of the application.

2.3.2 Synchronous converter systems

Principle scheme for ship's propulsion using synchro-converter it is shown in Figure 2.12

In the scheme can be recognize the following devices:

- A thyristor rectifier which generates a DC system
- A thyristor inverter which fed the synchronous motor with a constant magnitude and variable frequency voltage
- Static excitation of the machine
In the DC side of the system as can be see, an inductor is present so that the DC current can be considered constant in the 10 ms scale. In this way the currents which circulate in the motor have the known trapezoidal waveform, as shown in Figure 2.13.

![Figure 2.13: current waveforms using synchro-converter](image)

As can be seen, the state of conduction of the motor changes after \( T/6 \). (T in the period of the fundamental waveform) This currents situation implies that the magneto motive force generated by the currents will advance in steps, a schematization is shown in Figure 2.14.

![Figure 2.14: step's advancing currents](image)

This characteristic behavior of the magneto motive force generates torque oscillations, since the torque is proportion to the flux itself

\[
T_m = K \Phi_e \Phi_a \sin \gamma
\]  
\[ (2.5) \]
Where:

- $\Phi_e$ and $\Phi_a$ are excitation and armature flux, respectively
- $\gamma$ is the angle between them

In normal operation $\Phi_e$ rotates at nearly constant speed while $\Phi_a$ rotates with steps, this is the reason high torque oscillations are present.

A solution to this problem is the use of a double star synchronous machine with corresponding windings displaced by 30 degrees, in this way the armature flux will assume 12 positions instead of 6 decreasing sensibly the oscillations of the machine.

![Figure 2.15: double star synchronous machine used in synchro-converter fed system](image)

As an example can be seen that in this case was used the double star synchronous machine coupled with a 12 pulse rectifier, to improve harmonic content of the current draw by the grid.
2.4 Locomotive traction system

In the past synchronous motor drives were considered complicated and cumbersome, so asynchronous drives were preferred. The development of power electronic and in particular of electrical drives with intermediate DC circuit, allowed new development in synchronous machine drives for traction systems. In particular, in 1982 was experienced the prototype BB 10004 of SNFC. The favorable results made possible to understand that synchronous converter systems were preferable in high power applications.

2.4.1 BB 26000 locomotive system

The locomotive BB 26000 is an example of traction system in which a double star synchronous machine was used.

The bi-current BB 26000 called SYBIC (syncrones-bicourant), are derive derived from the single voltage BB 10004 and they can generate the nominal performances with both supply systems. The electrical drive is subdivided in two independent sub-drives, their scheme is represented in the two Figures 2.16 and 2.17. Regulation is carried out with a step down converter. (for example in Figure 2.17 can be seen converter CH1, used to regulate one of the two traction motor) Converters are designed for the maximum power, which is, in this case 2800W. The electrical power can be taken from the DC grid through extra-rapid switch or from the ac grid through transformer and single
phase mixed bridge. The aim of the two input stages is to supply the nominal voltage which is 1600 V.

Figure 2.17: input stage of BB 26000 locomotive
Chapter 3

Double star permanent magnets synchronous machine

Chapter three will present the main focus of the work: the so called double star PMSM. The chapter is organized in order to understand primarily the construction of the machine, from which it will be deducted the mathematical model of the machine in the time domain. Then, from the results obtained in the time domain analysis and using an appropriate and particular Park’s transformation, the transformed model is obtained and explained. As it will be seen the variable transformation allows many advantages that will be exploited for the control system in the chapter related to simulations.

3.1 Description of the machine

The double star synchronous machine is a synchronous multiphase machine in which there are two sets of stator windings. The magnetic axes of the two stator windings are displaced each other by \( \alpha = 30 \, \text{deg} \) in order to have \( \alpha = \frac{60}{N} \) (where \( N \) is the number of three phase sets) which is the condition to obtain the so called split-phase configuration. This winding arrangement has different advantages, such as: possibility of use already existing three phase components, greater power density and less vibrating mechanical torque. The air gap situation can be represent as shown in the Figure 3.1. The rotor is the same as its three phase counterparts and it is made with permanent magnets, in particular interior permanent magnets (IPM) and surface permanent magnets (SPM) manufacture techniques are used.
Interior permanent magnet (IPM) machines are widely used taking advantage of the additional torque component due to reluctance dependence on rotor position; such dependence obviously affects the winding inductances, introducing a relevant complication in the case of double star.

The inherent advantage of this machine type is the elimination of the sixth harmonic torque pulsation. Furthermore, using two three-phase sets instead of one increases the redundancy of the system, as it was seen in chapter 2. Park’s transformation is generally applied to three phase machines to model the machine using the rotor reference frame, which, especially for IPM machines, is an important tool to eliminate the rotor angle dependency of inductances. From the control design point of view, such simplifications are usually acceptable since non-idealities can be considered as disturbances, but from the simulation point of view, the model may be oversimplified, and thus, important phenomena may not become visible.

The electromagnetic coupling between the winding sets makes the modeling of double-star machines more complex compared with conventional three-phase machines where such a coupling, obviously, does not exist.

### 3.1.1 Rotor configurations

The rotor is made by permanent magnet materials. In recent years the interest in this materials is improved and this made possible to have better materials at relative low cost, this is the reason why PMSM are nowadays used in much more applications with respect to the past.

The earliest manufactured magnet materials were hardened steel. Magnets made from steel were easily magnetized. However, they could hold very low energy and it was easy to demagnetize. In
recent years other magnetic materials such as Aluminum Nickel and Cobalt alloys (ALNICO), Strontium Ferrite or Barium Ferrite (Ferrite), Samarium Cobalt (First generation rare earth magnet) (SmCo) and Neodymium Iron-Boron (Second generation rare earth magnet) (NdFeB) have been developed and used for making permanent magnets.

<table>
<thead>
<tr>
<th>Property (value @ 20°C)</th>
<th>Unit</th>
<th>Alnico</th>
<th>Anisotropic Ferrite</th>
<th>Sintered Sm-Co</th>
<th>Sintered Nd-Fe-B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Remanence B_r</td>
<td>T</td>
<td>0.6 to 1.35</td>
<td>0.35 to 0.43</td>
<td>0.7 to 1.05</td>
<td>1.0 to 1.3</td>
</tr>
<tr>
<td>Intrinsic Coercivity H_c</td>
<td>kA/m</td>
<td>40 to 130</td>
<td>180 to 400</td>
<td>800 to 1500</td>
<td>800 to 1900</td>
</tr>
<tr>
<td>p.u. recoil permeability H_repl</td>
<td>kJ/m³</td>
<td>20 to 100</td>
<td>24 to 36</td>
<td>140 to 220</td>
<td>180 to 320</td>
</tr>
<tr>
<td>Resitivity μΩ-cm</td>
<td></td>
<td>47</td>
<td>&gt; 104</td>
<td>86</td>
<td>150</td>
</tr>
<tr>
<td>Thermal expansion 10^5/°C</td>
<td>°C</td>
<td>11.3</td>
<td>13</td>
<td>9</td>
<td>3.4</td>
</tr>
<tr>
<td>B_r temperature coefficient %/°C</td>
<td></td>
<td>-0.01 to -0.02</td>
<td>-0.2</td>
<td>-0.045 to -0.05</td>
<td>-0.08 to -0.15</td>
</tr>
<tr>
<td>H_c temperature coefficient %/°C</td>
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<td>0.2 to 0.4</td>
<td>-0.2 to -0.25</td>
<td>-0.5 to -0.9</td>
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<tr>
<td>Max. working temperature °C</td>
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<td>500 to 550</td>
<td>250</td>
<td>250 to 350</td>
<td>80 to 200</td>
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<td>4900</td>
<td>8200</td>
<td>7400</td>
</tr>
</tbody>
</table>

Table 3: main properties of hard magnetic materials

The rare earth magnets are categorized into two classes: Samarium Cobalt (SmCo) magnets and Neodymium Iron Boride (NdFeB) magnets. SmCo magnets have higher flux density levels but they are very expensive. NdFeB magnets are the most common rare earth magnets used in motors these days. A flux density versus magnetizing field for these magnets is illustrated in Figure 3.2
Hard magnetic materials are characterized by a broad main hysteresis loop and by a high coercitivity (for this reason they are called “hard” materials); on the contrary, iron and most of iron alloys have a narrow main hysteresis loop, and a low coercivity (for this reason they are called “soft” materials). This property of the hard materials implies a high value of hysteresis losses (since, when the supply delivers a DC current, the area of the hysteresis loop equals the hysteresis loss). Actually, this is not a trouble, since in normal operating conditions, PMs work with static fields (not with AC fields), thus the PM material do not travel along the hysteresis loop, and the hysteresis losses do not occur.

![Figure 3.3: working point in hard magnetic materials](image)

Instead, the useful consequence of a broad hysteresis loop is that when such a material is inserted in a magnetic circuit, the working point on the magnetization curve has a high flux density, therefore such material can be used as a flux source.

On the contrary, if a soft material is inserted in a magnetic circuit, the working point on the magnetization curve has a very low flux density, therefore a soft material cannot be used as a flux source. This property is demonstrated if the working point is obtained by a graphical intersection between the magnetic characteristic of the flux source (the $B - H$ curve of the PM in the 2$^{nd}$ quadrant of $B - H$ plane) and the magnetic characteristic of the magnetic, passive load (a straight line, passing through origin of $B - H$ plane, with a negative slope).
In PM materials an important parameter is the energy product, let’s consider the magnetic voltage law in the air gap:

\[ H_m h_m = -H_g g = -U_g \]  \hspace{1cm} (3.1)

And the flux balance:

\[ B_m A_m = B_g A_g = \phi_g \]  \hspace{1cm} (3.2)

show that:

- in order to produce the same flux \( \phi_g \) in the air gap, a PM with a higher flux density \( B_m \) requires a lower section area \( A_m \).
- in order to produce in the air gap the same magnetic voltage \( U_g \), a PM with higher magnetic strength \( H_m \) requires a lower magnetization length \( h_m \).

If equations 3.1 and 3.2 are multiplied term by term, and by taking into account the link between \( B \) and \( H \) in the air gap

\[ B_g = \mu_0 H_g \]  \hspace{1cm} (3.3)

It results

\[ B_m H_m A_m h_m = -\frac{B_g^2}{\mu_0} A_g g = -U_g \phi_g \]  \hspace{1cm} (3.4)

By defining \( V_m = A_m x h_m \) and \( V_g = A_g x g \) the PM and air gap volume respectively, equation 3.4 becomes

\[ B_m H_m V_m = -\frac{B_g^2}{\mu_0} V_g \]  \hspace{1cm} (3.5)
And solving for $V_m$:

$$V_m = - \frac{B_g^2}{\mu_0} V_g \frac{1}{B_m H_m}$$

(3.6)

The minus sign in equation 3.6 occurs because the $PM$ working magnetic strength $H_m$ is negative, due to the operation in the 2$^{nd}$ quadrant. Equation 3.6 shows that, if the air gap volume $V_g$ and flux density $B_g$ are defined, the $PM$ volume $V_m$ is in inverse proportion with respect to the product $B_m H_m$: this quantity is an energy per unit volume $W/m^3$, and for this reason it is called energy product. Thus, it seems to be advisable that the $PM$ working point be in that portion of the characteristic where the product $B_m H_m$ assumes its maximum value, since in this point (with fixed value of $B_g$, $H_g$, $g$, $A_g$) the $PM$ volume has a minimum. In a similar way, since the energy stored in the air gap is expressed by:

$$W_g = - \frac{1}{2} \frac{B_g^2}{\mu_0} A_g g = - \frac{1}{2} B_m H_m V_m$$

(3.7)

if the $PM$ volume is defined, the air gap energy is the higher, the higher is the energy product $B_m H_m$. But we must investigate where is the point with maximum value of the product $B_m H_m$. Figure 3.5 shows the behavior of the product $B_m H_m$ as a function of $B$. Usually, on the $B - H$ plane, together with the demagnetization curves, also the hyperbola loci of constant $B_m H_m$ are reported; this allows to read the $B_m H_m$ value in each point, and in particular its maximum value $(B_m H_m)_{max}$. The same figure shows that the point with $(B_m H_m)_{max}$ is on the knee of the $PM$ curve (this usually occurs for materials which have a sharp knee). In applications where, during the operation, the $PM$ is subjected to varying demagnetizing fields, and/or the geometrical sizes of the circuit change, and/or the temperature varies, the $PM$ working point can move, and may position itself below the knee, causing the $PM$ irreversible demagnetization.

Therefore, as regards the $PM$s whose knee is in the 2$^{nd}$ quadrant, it is advisable to make them work not in the point $(B_m H_m)_{max}$, but in a point shifted on the right, in order to avoid the cited demagnetization risk.
PM motors are broadly classified by the direction of the field flux. The first field flux classification is radial field motor meaning that the flux is along the radius of the motor. The second is axial field motor meaning that the flux is perpendicular to the radius of the motor. Radial field flux is most commonly used in motors and axial field flux have become a topic of interest for study and used in a few applications.

PM motors are classified on the basis of the flux density distribution and the shape of current excitation. They are PMSM and PM brushless motors (BDCM). The PMSM has a sinusoidal-shaped back EMF and is designed to develop sinusoidal back EMF waveforms. They have the following features:

- Sinusoidal distribution of magnet flux in the air gap
- Sinusoidal current waveforms
- Sinusoidal distribution of stator conductors.

BDCM has a trapezoidal-shaped back EMF and is designed to develop trapezoidal back EMF waveforms. They have the following features:

- Rectangular distribution of magnet flux in the air gap
- Rectangular current waveform
- Concentrated stator windings.

In PM machines, the magnets can be placed in two different ways on the rotor. Depending on the placement they are called either as surface permanent magnet motor or interior permanent magnet motor. Surface mounted PM motors have a surface mounted permanent magnet rotor. Each of the PM is mounted on the surface of the rotor, making it easy to build, and specially skewed poles are easily magnetized on this surface mounted type to minimize cogging torque. This configuration is
used for low speed applications because of the limitation that the magnets will fly apart during high-speed operations. These motors are considered to have small saliency, thus having practically equal inductances in both axes.

The permeability of the permanent magnet is almost that of the air, thus the magnetic material becoming an extension of the air gap. For a surface permanent magnet motor $L_d = L_q$. The rotor has an iron core that may be solid or may be made of punched laminations for simplicity in manufacturing. Thin permanent magnets are mounted on the surface of this core using adhesives. Alternating magnets of the opposite magnetization direction produce radially directed flux density across the air gap. This flux density then reacts with currents in windings placed in slots on the inner surface of the stator to produce torque. Figure 2.3 shows the placement of the magnet.

![Figure 2.3: Placement of the magnet.](image)

**Figure 3.6(a): surface permanent magnets (b) interior permanent magnets**

Interior $PM$ motors have interior mounted permanent magnet rotor as shown in figure 2.4. Each permanent magnet is mounted inside the rotor. It is not as common as the surface mounted type but it is a good candidate for high-speed operation. There is inductance variation for this type of rotor because the permanent magnet part is equivalent to air in the magnetic circuit calculation. These motors are considered to have saliency with $q$-axis inductance greater than the $d$-axis inductance ($L_d > L_q$).
3.1.2 No load back EMF

The no-load back-EMF waveform in PM synchronous machines depends on the rotor electrical rotational speed $\omega_e$ and the $\Psi_{PM}$ flux produced by the PMs. Commonly, $\Psi_{PM}$ is assumed constant in transformed models. In the steady state, $\omega_e$ can be assumed to be constant with a negligible error. Assuming both the values constant results in sinusoidal voltage waveforms. However, assuming $\Psi_{PM}$ constant requires the air-gap flux density distribution to be sinusoidal. It is worth mentioning that the harmonics in the no-load flux linkage can be taken into account in transformed models also.

Generally, the voltage induced in the winding is given by:

$$E_0 = -\frac{d\Psi_{PM}}{dt}$$

Where $\Psi_{PM}$ is the flux linkage of the winding. The parameters having the greatest effect on the induced voltage are the winding factor, the number of turns of the coil, the cross-sectional area of the coil, and the flux density in the air gap.

In PM machines, the air gap flux density is obviously affected by the magnet configuration and the magnetization direction, as well as by the stator geometry, e.g. stator slotting, which can be taken into account by an analytical method, based on the determination of the air-gap magneto motive force and the flux path permeance function. Besides being rather complex, such a model requires (similarly as an FE model) the knowledge of the geometry details of the machine and the material characteristics. In addition to the stator slotting, the winding layout has an important impact on the harmonics generated in the back EMF.

The greatest winding factor for the fundamental wave in the case of a conventional three phase machine is achievable with a full-pitch winding layout. As a drawback, the winding factors of the harmonics (especially the $5^{th}$ and $7^{th}$) increase as well. The double-star full-pitch windings generally exhibit even higher harmonic components. As a benefit, higher voltage is obtained, the same air gap flux being maintained, even if with a greater harmonic content.
3.1.3 Hypotheses

In developing the mathematical model, the following assumptions are made:

- The set of three-phase stator winding is symmetrical.
- The capacitance of all the windings can be neglected.
- Each of the distributed windings may be represented by a concentrated winding.
- The change in the inductance of the stator windings due to rotor position is sinusoidal and does not contain higher harmonics.
- Hysteresis loss and eddy current losses are ignored.
- The magnetic circuits are linear (not saturated) and the inductance values do not depend on the current.

3.2 Phase variable model

3.2.1 Model Structure

The phase-variable model can be written in the following vector notation:

\[ V = R_s I + p\Psi \quad (3.9) \]

where is \( R_s \) the stator winding resistance matrix, \( p \) is the time derivative operator, and \( \Psi \) is the flux linkage vector composed of the no-load flux \( \Psi_{PM} \) generated by the PMs and the armature reaction as follows:

\[ \Psi = L(\theta)I + \Psi_{PM} \quad (3.10) \]

The vectors \( V, I, \Psi \) and \( \Psi_{PM} \) are in the form:

\[ f = [ f_{a1} \ f_{b1} \ f_{c1} \ f_{a2} \ f_{b2} \ f_{c2} ]^T \quad (3.11) \]

Currents are selected as state variables and the equations to be solve are derived from the system of equation 3.9, obtaining:

\[ pI = L(\theta)^{-1}(V - p\Psi_{PM} - R_s I + pL(\theta) \cdot I) \quad (3.12) \]
3.2.2 Inductances

The dependence of the self- and mutual inductances on rotor position $\theta$ is generally conveniently expressed by Fourier expansions. For example, a general equation defining a self-inductance and taking into account higher order harmonics, can be written as:

$$L_i(\theta) = L_{i0} + \sum_{n=1}^{\infty} L_{i2n} \cos(2n\theta_i + \varphi_{in}) \quad (3.13)$$

where the subscript $i$ denotes the $i^{th}$ winding ($i = a_1, ..., c_2$), $L_{i0}$ is the average value, $L_{i2n}$ is the coefficient of inductance harmonics ($2^{nd}$, $4^{th}$, $6^{th}$, ...), and $\varphi_{in}$ is the displacement angle of the corresponding harmonic component. A similar equation can be written to define the mutual inductance as well:

$$M_{ij}(\theta) = M_{ij0} + \sum_{n=1}^{\infty} M_{ij2n} \cos(n(\theta_i + \theta_j) + \varphi_{in}) \quad (3.14)$$

where the subscript $i, j$ denotes the coupling between the $i^{th}$ and $j^{th}$ windings ($i, j = a_1, ..., c_2, i \neq j$).

3.2.3 Inductance waveforms

In general, when an electrical machine is modeled using Park’s transformation, the $d-q$ axes inductances are considered constant, meaning that the higher-order inductance harmonics are neglected. The fundamental waveforms are defined by equations 3.13 and 3.14 with $n = 1$.

Using the inbuilt post-processing procedures of the Maxwell FEM software, the inductance waveforms that also include higher-order harmonics can be obtained. Figure 3.7 shows the work made in [3], in which inductance waveforms related to coil $a_1$ are illustrated. The relative amounts of harmonic components with respect to the fundamental one are shown in Table 4. According to Table 4, significant amounts of harmonics, especially even harmonics, are present. The presence of even harmonics is clearly visible also in Figure 3.7 because the negative and positive half-cycles are not symmetrical. Particularly, the values of the $4^{th}$-order inductance coefficients ($n = 2$) are significantly high.
3.2.4 Inductance matrix

Considering only the first inductance harmonic equations 3.13 and 3.14 become:

\[ L_i(\theta) = L_{i0} + L_{i0} \cos(2n\theta_i) \] (3.15)

\[ M_{ij}(\theta) = M_{ij0} + M_{ij0} \cos(\theta_i + \theta_j) \] (3.16)

It can be shown that the inductance matrix is made in this way:
\[ L(\theta) = \begin{bmatrix} L_1(\theta) & M_{12}(\theta) \\ M_{21}(\theta) & L_2(\theta) \end{bmatrix} \]  
\hspace{1cm} (3.15)

It is a 6\times6 full matrix in which we can recognize four main matrices, in particular we have:

- \( L_1(\theta) \): matrix associates with the first three phase winding, it expresses its self and mutual inductance.
- \( L_2(\theta) \): matrix associated with the second three phase winding.
- \( M_{12}(\theta) \), \( M_{21}(\theta) \): these two matrices represent the effect of one three phase winding on the other.

Each of them are 3\times3 full matrices where every single term is \( \theta \) dependent, later will be shown that using Park’s transformation we can simplify these matrices and eliminate the rotor position dependence.

For the sake of clarity can be shown the complete matrix:

\[
\begin{bmatrix}
L_{a1}(\theta) & M_{a1b1}(\theta) & M_{a1c1}(\theta) & M_{a1a2}(\theta) & M_{a1b2}(\theta) & M_{a1c2}(\theta) \\
M_{b1a1}(\theta) & L_{b1}(\theta) & M_{b1c1}(\theta) & M_{b1a2}(\theta) & M_{b1b2}(\theta) & M_{b1c2}(\theta) \\
M_{c1a1}(\theta) & M_{c1b1}(\theta) & L_{c1}(\theta) & M_{c1a2}(\theta) & M_{c1b2}(\theta) & M_{c1c2}(\theta) \\
M_{a2a1}(\theta) & M_{a2b1}(\theta) & M_{a2c1}(\theta) & L_{a2}(\theta) & M_{a2b2}(\theta) & M_{a2c2}(\theta) \\
M_{b2a1}(\theta) & M_{b2b1}(\theta) & M_{b2c1}(\theta) & M_{b2a2}(\theta) & L_{b2}(\theta) & M_{b2c2}(\theta) \\
M_{c2a1}(\theta) & M_{c2b1}(\theta) & M_{c2c1}(\theta) & M_{c2a2}(\theta) & M_{c2b2}(\theta) & L_{c2}(\theta) \\
\end{bmatrix}
\hspace{1cm} (3.16)

In which it can be recognize the four main matrices presented earlier.

### 3.2.5 Electromagnetic torque

The torque equation can be derived starting from the energy stored in the magnetic fields and defined as follow:

\[
T_e = n_p \frac{1}{2} I^T \frac{\partial L}{\partial t} I + I^T \frac{\partial}{\partial t} \Psi_{PM} \omega_e
\hspace{1cm} (3.17)
\]

Where \( n_p \) is the number of pole pairs.
3.3 Transformed variables model

3.3.1 Introduction

General tools for multiphase machine modeling have been developed in the first half of the 20th century. The well-known space vector and $d - q$ models of three-phase machines are only particular cases of the universal $n$-phase machine models. Since the phase-variable model of a physical multiphase machine gets transformed using a mathematical transformation, the number of variables before and after transformation must remain the same.

This means that an $n$-phase machine will have $n$ new stator current (stator voltage, stator flux) components after the transformation. If a machine is with sinusoidal-field distribution, standard modeling assumptions apply and only the first harmonic of inductance terms exists in the phase-variable model. Application of the decoupling (Park’s) transformation produces a set of $n$ equations. The transformation will eliminate the dependence on the rotor position, simplifying the modelization of the machine and the design of the drive.

3.3.2 Park transformation matrix

Park’s transformation for a three-phase system with no neutral connection is defined by the $2 \times 3$ matrix as:

$$T_p(\delta) = \sqrt{\frac{2}{3}} \begin{pmatrix} 
\cos(\delta) & \cos\left(\delta - \frac{2\pi}{3}\right) & \cos\left(\delta + \frac{2\pi}{3}\right) \\
-sin(\delta) & -sin\left(\delta - \frac{2\pi}{3}\right) & -sin\left(\delta + \frac{2\pi}{3}\right) 
\end{pmatrix} \quad (3.18)$$

The transformation matrix that transforms the phase variables, for instance, the $L(\theta)$ 6x6 matrix, into the $d - q$ reference frame is composed by two Park’s transformation matrices and can be expressed as:

$$T_1(\theta) = \begin{pmatrix} T_p(\theta + \alpha) & 0_{2,3} \\
0_{2,3} & T_p(\theta - \alpha) \end{pmatrix} = \begin{pmatrix} T_p(\theta + \alpha) & 0_{2,3} \\
0_{2,3} & T_p(\theta - \alpha) \end{pmatrix} \quad (3.19)$$

where $0_{i,j}$ is a null matrix with $i$ rows and $j$ columns. Applying matrix 3.19 to 3.16, the resulting matrix $L_{T_1}$, related to two couples of orthogonal windings $d^* - q^*$ and $d^{**} - q^{**}$, is given by:
\[ L_{T_1} = T_1(\theta)L(\theta)T_1^T(\theta) \]  \hspace{2cm} (3.20)

\[ L_{T_1} = \begin{pmatrix} L_{dq*} & M_{dq*} \\ M_{dq*} & L_{dq**} \end{pmatrix} \]  \hspace{2cm} (3.21)

Where

\[ L_{dq*} = L_{dq**} = \begin{pmatrix} L_{d*} & 0 \\ 0 & L_{q*} \end{pmatrix} \]

\[ M_{dq*} = \begin{pmatrix} M_{d*} & 0 \\ 0 & M_{q*} \end{pmatrix} \]

And

\[ L_{d*} = L_{s0} + \frac{L_{s2}}{2} - M_{s2} + M_{s2} \]

\[ L_{q*} = L_{s0} - \frac{L_{s2}}{2} - M_{s2} - M_{s2} \]

\[ M_{d*} = \frac{1}{2}(3M_{m0} + M_{m2}) \]

\[ M_{q*} = \frac{1}{2}(3M_{m0} - M_{m2}) \]

The complete \( L_{T_1} \) matrix is shown:

\[ L_{T_1} = \begin{pmatrix} (L_{d*} & 0) & (M_{d*} & 0) \\ (0 & L_{q*}) & (0 & M_{q*}) \end{pmatrix} \]  \hspace{2cm} (3.22)

With \( L_{T_1} \) not being a diagonal matrix, the transformed winding couples \( d^* - q^* \) and \( d^{**} - q^{**} \) are mutually coupled with respect to each other. The matrix of the further transformation required to eliminate such a coupling can be found by diagonalizing \( L_{T_1} \). With a suitable rearrangement of rows and columns, the transformation matrix \( T_2 \) to the final decoupled \( d_1 - q_1 \) and \( d_2 - q_2 \) reference frames (also referred to as \( d - q \) reference frames) is obtained:
\[ T_2 = k \begin{pmatrix} 1 & 0 & 1 & 0 \\ 0 & 1 & 0 & 1 \\ 0 & 1 & 0 & -1 \\ -1 & 0 & 1 & 0 \end{pmatrix} \] (3.23)

where the orthonormalization coefficient \( k = \frac{1}{\sqrt{2}} \) is introduced to enable the power to be invariant. Application of matrix 3.23 to 3.22 yields:

\[ L_{T_2} = \begin{pmatrix} L_{d_1} & 0 & 0 & 0 \\ 0 & L_{q_1} & 0 & 0 \\ 0 & 0 & L_{d_2} & 0 \\ 0 & 0 & 0 & L_{q_2} \end{pmatrix} \] (3.24)

where the inductances are defined by

\[
\begin{align*}
L_{d_1} &= L_{d*} + M_{d*} \\
L_{q_1} &= L_{d*} + M_{q*} \\
L_{d_2} &= L_{q*} - M_{q*} \\
L_{q_2} &= L_{d*} - M_{d*}
\end{align*}
\]

Finally, by combining matrix 3.19 with 3.23, the whole transformation results in

\[
T_3(\theta) = \begin{pmatrix} T_p(\theta + \alpha) & T_p(\theta - \alpha) \\ T_p(\theta + \alpha + \frac{\pi}{2}) & T_p(\theta - \alpha - \frac{\pi}{2}) \end{pmatrix}
\] (3.25)

Figure 3.8: graphical representation of the transformation
It is worth mentioning that $L_{T_2}$ is directly derivable by the application of $T_3(\theta)$. The proposed transformation eliminates, like Park’s one, the triplex harmonics from the transformed quantities, although they are generally not relevant in symmetric three-phase systems.

### 3.3.3 Transformed model

Starting from the phase variable model:

$$v_s = R_s i_s + p\Psi$$   \hspace{1cm} (3.9)

$$v_s = R_s i_s + L(\theta)p i_s + pL(\theta)i_s + e$$   \hspace{1cm} (3.9a)

Applying $T_3(\theta)$ matrix it can be obtained:

$$v_{dq} = R_s i_{dq} + L_{dq}p i_{dq} + \omega_m J L_{dq} i_{dq} + e_{dq}$$   \hspace{1cm} (3.26)

Where:

$$J = \begin{pmatrix} 0 & -1 & 0 & 0 \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \end{pmatrix}$$   \hspace{1cm} (3.27)

In expanded form the equation will have this form.

$$v_{d_1} = R_s i_{d_1} + L_{d_1}p i_{d_1} - \omega_e L_{q_1} i_{q_1}$$

$$v_{q_1} = R_s i_{q_1} + L_{q_1}p i_{q_1} + \omega_e L_{d_1} i_{d_1} + \omega_e \Psi_{PM}$$   \hspace{1cm} (3.28)

$$v_{d_2} = R_s i_{d_2} + L_{d_1}p i_{d_2} - \omega_e L_{q_2} i_{q_2}$$

$$v_{q_2} = R_s i_{q_2} + L_{q_2}p i_{q_2} + \omega_e L_{d_2} i_{d_2}$$

The model using currents as state variable can be write:

$$p i_{d_1} = \frac{1}{L_{d_1}}(v_{d_1} - R_s i_{d_1} + \omega_e L_{q_1} i_{q_1})$$

$$p i_{q_1} = \frac{1}{L_{q_1}}(v_{q_1} - R_s i_{q_1} - \omega_e L_{d_1} i_{d_1} - \omega_e \Psi_{PM})$$   \hspace{1cm} (3.29)

$$p i_{d_2} = \frac{1}{L_{d_2}}(v_{d_2} - R_s i_{d_2} + \omega_e L_{q_2} i_{q_2})$$
\[ pi_{q_2} = \frac{1}{L_{q_2}} (v_{q_2} - R_s i_{q_2} - \omega_e L_{d_2} i_{d_2}) \]

And it is necessary to add the mechanical equation as well, which is, in its simplified form:

\[ p \omega_e = \frac{n_p}{f} (T_e - T_m) \] (3.30)

### 3.3.4 Electromagnetic torque

The equation for the electromagnetic torque is derived starting from the instantaneous power:

\[ P(t) = v_{d_1} i_{d_1} + v_{q_1} i_{q_1} + v_{d_2} i_{d_2} + v_{q_2} i_{q_2} \] (3.31)

In the steady state and neglecting the resistive voltage drop, the equation 3.26 becomes:

\[ v_{dq} = e_{dq} - \omega_m J_{dq} i_{dq} \] (3.32)

Combining 3.31 and 3.32 and dividing \( P(t) \) by the angular speed \( \omega_m = \frac{\omega_e}{n_p} \) the electromagnetic torque is obtained:

\[ T_e = n_p (i_{q_1} \Psi_{d_1}^* + i_{q_2} \Psi_{d_2}^* - i_{d_1} \Psi_{PM,q_1} - i_{d_2} \Psi_{PM,q_2}) \] (3.33)

With

\[ \Psi_{d_1}^* = i_{d_1} (L_{q_1} - L_{d_1}) + \Psi_{PM,d_1} \]
\[ \Psi_{d_2}^* = i_{d_2} (L_{d_2} - L_{q_2}) + \Psi_{PM,d_2} \]

If PM flux harmonics are negligible but the fundamental one, \( L_{d_2} = L_{q_2} \) and taking into account inductance expression, \( M_{m_2} = L_{s_2} + 2M_{s_2} \), equation 3.33 simplifies to

\[ T_e = n_p \Psi_{d_1}^* i_{q_1} = n_p [\Psi_{PM} i_{q_1} + i_{q_1} i_{d_1} (L_{d_1} - L_{q_1})] \] (3.34)

recalling the simplified expression of the electromagnetic torque of variable-reluctance three-phase machines.

The torque is made by two contributions, the first one is the interaction between the current on the \( q_1 \) axis of the first winding set with the permanent magnet flux on the same axis and it is an
alignment torque, proportional to currents and flux. The second term is due to permanent magnets shape and it is a reluctance torque contribution, IPM machines are hybrid machines, meaning that the total torque is a combination of isotropic PMSM and a reluctance synchronous machine.

3.3.5 Equivalent circuit

To show the machine equivalent circuit it can be take into account the founded $d - q$ decoupled system.

$$
\begin{align*}
    v_{d1} &= R_s i_{d1} + L_{d1} p i_{d1} - \omega_e L_{q1} i_{q1} \\
    v_{q1} &= R_s i_{q1} + L_{q1} p i_{q1} + \omega_e L_{d1} i_{d1} + \omega_e \Psi_{PM} \\
    v_{d2} &= R_s i_{d2} + L_{d2} p i_{d2} - \omega_e L_{q2} i_{q2} \\
    v_{q2} &= R_s i_{q2} + L_{q2} p i_{q2} + \omega_e L_{d2} i_{d2}
\end{align*}
$$

(3.28)

As can be seen the two axes are divided, to obtain two equivalent circuit, in which is possible to recognize the FEM contributes.

![Figure 3.9: d-axis equivalent circuit](image1)

![Figure 3.10: q-axis equivalent circuit](image2)
Chapter 4

control strategies

In the next chapter wind turbine systems and means to control grid side converters will be presented. The aim of the brief presentation of wind turbine systems is to give the needed information to understand the control problem, such as: input variable, aim of the control system, desire value of output variables. Additionally, classical PMSM topologies are presented along with the studied structure which is, as already said, not yet used in this kind of generation systems.

4.1 Wind turbines general characteristics

4.1.1 Generalities

There are horizontal and vertical axes wind turbine, the most installed are the horizontal axes wind turbines and the ones studied in this work. The power density of the wind is not very high, so to have interesting powers the only mean is to have a high swept area by the blades. A wind turbine needs a minimal speed of 3 – 5 m/s (cut-in) to deliver power to the grid instead the rated power is delivered when the wind speed reach 12 – 14 m/s. When the speed is higher of 25 m/s (cut-off) the turbine is stopped for safety reasons.

To the air flow can be associated a power (considering air flow with constant speed $v$ and perpendicular with respect to section $S$):

$$ P_w = \frac{d}{dt} \left( \frac{1}{2} m v^2 \right) = \frac{1}{2} v^2 \frac{dm}{dt} = \frac{1}{2} v^2 (\rho S v) = \frac{1}{2} \rho S v^3 $$  \hspace{1cm} (4.1)
The extraction of power from the air flow implies that the air will slow down and that its section increases. The power can’t be extracted completely from the air flow otherwise the air would stop in the considered section.

The maximum power which can be extracted from the air flow is:

\[
P^* = \frac{1}{2} \rho S v^* (v_1^2 - v_2^2) = \frac{1}{2} \rho S \frac{2}{3} v_1 \left( \frac{8}{9} v_1^2 \right) = \frac{1}{2} \rho S \frac{16}{27} v_1^3
\]

Figure 4.1: (a) wind turbine blades and relative speeds (b) air flow

The value \(16/27 = 0.593\) is the Betz’ limit. The maximum power which can be converted from the undisturbed flow is about the 60%.

In practice the real extracted power is 60 – 70 % of this theoretical value.

Power is:

- proportional to \(S\) i.e. \(D_b^2\)
- the cost is more or less proportional to \(D_b\)

This is the reason why are constructed bigger WTs, furthermore the wind in higher heights have, usually, more strength. Power coefficient \(c_p\), is the ratio between extracted power from the turbine and the power from the undisturbed flow:

\[
c_p = \frac{P}{P_w} = \frac{1}{\frac{1}{2} \rho S v^3} \Rightarrow P = \frac{1}{2} c_p \rho S v^3
\]

The power coefficient is a function of the tip speed ratio (\(\lambda\) or \(TSR\)) and the pitch angle \(\beta\). The expression of \(\lambda\) is:
\[ \lambda = \frac{\Omega r}{v} = \frac{\pi n_r D}{60v} \] (4.4)

Equation 4.4 for a given pitch angle (which means that the aerodynamic is fixed), states that:

- If aerodynamic of the blades is fixed only one optimal TSR exist, in which \( c_p = c_{p_{\text{max}}} \) as shown in Figure 4.3(a)
- When \( v \) varies is necessary to change deliberately the blades’ rotational speed in order to maintain \( c_p = c_{p_{\text{max}}} \)
- TSR low implies lift reductions and higher resistance, until the stall is reach
- If TSR is high lift and resistance must be reduced
- The less is the number of blades faster turbines have to rotate in order to extract the maximum wind power from the wind flow
- Fast turbines have a large \( c_p \) when TSR is high, slow turbines have the contrary characteristic

If we consider for example the curve in Figure 4.3a, looking at equation 4.4 and considering a constant \( \beta \) can be said that:

- \( r = 10 \text{ m (fixed)} \) and \( v = 10 \text{ m/s} \)
- If \( \Omega_r = 40 \text{ rpm} = 4 \text{ rad/s} \), we’d have: \( \lambda = 4 \) and so \( c_p = 0.12 \)
- If \( \Omega_r = 80 \text{ rpm} = 8 \text{ rad/s} \), we’d have: \( \lambda = 8 \) and so \( c_p = 0.42 \)
This simple numerical example well explain why it is better to have a turbine which can varies it’s rotational speed, in fact a constant speed wind turbine cannot exploit the maximum produced power except for the time in which the speed of the wind is so that: $v = v(\lambda_{opt})$.

To produce the highest annual energy is suggested to maintain the $c_p$ as large as possible also when the wind speed changes, Figure 4.3(b) shown a representation of this situation: power varies with $v^3$.

And the $TSR$ is maintained at the condition of maximum power coefficient which corresponds to the maximum extracted power.

### 4.1.2 Variable speed wind turbine’s regulation systems

A wind turbine can be schematized as a mass with a large inertia which rotates along with the electrical generator. To these masses are applied the aerodynamic torque of the turbine and the electromagnetic torque of the generator. For speed lower than nominal one the control system and regulation act to maximize the aerodynamic torque while for speed larger of nominal speed control system has to modulate such torque to maintain the speed below security limits.

In variable speed wind turbines, the electromagnetic torque can be varied independently from aerodynamic torque therefore rotor rotational speed can be varied acting both on aerodynamic torque and electromagnetic torque. Variation of electromagnetic torque is achieved through power converter which regulates the generator currents. Advantages of variable speed wind turbines are:

- Extracted power from wind increases (greater efficiency)
- When wind speed is low rotational speed of the rotor is lower with respect to fixed speed wind turbine, the noise is reduced.
- Power quality increases
• Lower dynamical stresses

The main drawback is about losses in the power converter and higher costs due to electrical equipment.

4.1.3 Pitch regulation system

For moderate wind speed the pitch angle is maintained constant and generator torque is regulated to follow the maximum $c_p$ condition. When wind speed increases, generally the rotor reach its nominal speed before the nominal power is achieved, as can be seen in Figure 4.4(b). In this condition the rotational speed is maintained constant with consequent fluctuation of output power. Combined action of electromagnetic torque and pitch regulation ensure to maintain the power to its nominal value.

The control scheme is shown in Figure 4.4(a) in which is possible to see that the instantaneous rotational speed is compared both with reference rotational speed and $P - c_p$ curve of the generator with $c_{pmax}$ nominal rotational speed as we said is decided based on wind speed. Pitch regulation acts when the rotational instantaneous speed exceed the reference value and reduce it. For a given instantaneous rotational speed, the output power is given by the curve, so the power converter is controlled to reach the reference power or to limit the power to its reference value.

Figure 4.4: (a) wind turbine’s control system (b) generator rotational speed and consequent generated power
4.1.4 Synchronous generator system configuration

Nowadays, PMSG are widely used in wind turbine applications due to their high reliability and good power density. In this paragraph will be shown the classic full converter configuration system, in order to anticipate and explain the studied structure with double star synchronous machine and two power converters. There are other many possible solutions to allow variable speed operation. In particular, the most important are:

- Induction generator with wound rotor and output variable resistance.
- Induction generator with wound rotor in doubly fed configuration
- Induction generator with wound rotor with power converter between stator and grid (full converter configuration)

The classical full converter configuration is made by: synchronous generator, power converter, transformer, gear box (if present), brakes, pitch drive system and control system.

As it was already said, frequency at generator output is proportional to the rotational speed, so to maintain grid frequency as constant as possible a double stage power converter is needed to decouple the generator from the grid:

- First stage converts the electrical quantities in DC ones through diode bridge or controlled bridge
- Second stage converts DC quantities in alternating ones to match with grid requirements

Electromechanical conversion can occur at high, medium or low speed. High speed conversion (up to 2000 rpm) is similar to doubly-fed configuration in which a triple stage gear box is present which is connected with a high speed generator, (usually PM) the main advantage is to use a smaller and
lighter generator. Medium speed conversion (up to 500 \( \text{rpm} \)) uses a single or double stage gearbox coupled with a \( PM \) generator.

Lower speed allows smaller mechanical stresses with the consequence of higher reliability. The generator diameter is larger with respect to the previous case. Low speed conversion (up to 30 \( \text{rpm} \)) is made without gearbox and the generator can be \( PM \) or wound rotor, due the low speed the generator has to have a high pole number which corresponds to an even bigger generator but the advantage of eliminate the gearbox with consequent decrease of mechanical losses and higher reliability makes this solution very interesting.

4.2 Power converter control strategies

The main used strategies in power converter control are sinusoidal pulse width modulation (\( PWM \)) and space vector modulation.

4.2.1 Sinusoidal pulse width modulation

Briefly, in \( PWM \) two signals are compared (carrier and control signal) in order to generate the switching signal. An example of single phase pulse width modulation is shown in Figure 4.7(a) while Figure 4.7(b) presents the harmonic content of the output voltage. The main quantities in pulse width modulation are:

- \( m_a \): amplitude modulation ratio, which is defined as:

\[
m_a = \frac{V_{\text{control}}}{V_{\text{triangular}}} \tag{4.5}
\]
Where $V_{\text{control}}$ is the maximum of the sinusoidal control signal and $V_{\text{triangular}}$ is the maximum of the carrier triangular signal.

- $m_f$: frequency modulation ratio, which is defined as:

$$m_f = \frac{f_s}{f_1} \quad (4.6)$$

![Figure 4.7: (a) pulse width modulation (b) harmonic content](image)

Where $f_s$ is the frequency of the carrier signal and $f_1$ is the frequency of the control signal, in particular $f_1$ is also the frequency of the first harmonic of the output voltage. It can be demonstrated that the harmonic order is proportional to $m_f$, a high $m_f$ means that harmonics will be shifted at high frequency, it will be said how they are distributed.

The harmonic spectrum of $v_{A0}$, (output voltage waveform of the inverter) it brings three important informations:

- The peak amplitude of the fundamental frequency component ($V_{A0}$) is $m_a$ times $\frac{1}{2} V_{DC}$
- The harmonics in $v_{A0}$ appear as sidebands, centered around the switching frequency and its multiplies, that is: $m_f, 2m_f, 3m_f$ and so on. This general pattern holds for $m_a$ in the range $0 - 1$. For $m_f \geq 9$ (which is always the case) harmonic amplitudes are almost independent of $m_f$ and the frequencies and orders at which they occur can be indicated as:

$$f_h = (jm_f \pm k)f_1 \quad (4.7)$$

$$h = j(m_f) \pm k \quad (4.8)$$
- $m_f$ should be an odd integer. Doing that, the result is an odd symmetry [$f(-t) = -f(t)$] as well as a half-wave symmetry [$f(t) = -f(t + \frac{1}{2}T_1)$]. In this way only odd harmonics are present in the output voltage.

In the case of six leg converter the situation will not change, in the sense that the basic concept is the same but it is necessary to have six control signals.

![Figure 4.8: six leg converter structure](image)

It will be better explained later the characteristics of this power components since PWM is the modulation technique adopted in the simulations.

### 4.2.2 Space vector modulation

This technique, in the classical case of three phase system, is based on control the global three phase system and not on control separately each phase. Moreover, it is easier to implement on a microprocessor and shifts more significant harmonics on a higher frequency value. It is well known that the inverter can assume eight different conduction’s state. (which can be characterized with set of three bits)

The situation is well explained in the two following figures in which are presented the six possible positions of the voltage space vector (function of the switches) and the graph of the consecutive transitions from one state to another. (it is useful to minimize the transitions) To obtain the reference space vector shown in Figure 4.9(c) it will be provide in a sampling period two voltages (in figures are vectors 6 and 4) to obtain the resultant reference vector. Each sampling period the reference will move and the provided vectors will change.
Similarly, standard three phase space vector modulation can be used for the two converters. A combinatorial analysis of the inverter switch states shows 64 switching modes. Each voltage vector is represented by a decimal number corresponding to the binary number.

In space vector $PWM$ technique based on vector classification, which is a simple method to control the power converters, each of the inverters independently with space vector $PWM$ strategy. Each inverter will then generate the hexagon voltage vector diagram. This can be achieved by means of two different techniques as shown in Figure 4.10.

First case is, using separate reference signals for each converter but the same switching states. The same space vector modulation can be used, only by changing the input signal to the function. The second case, is using the same reference signal but different position of switching states, the switching states of one of the converter are uniformly displaced by the angle which is equal to the phase difference of the two groups of three phase winding of the generator.
Despite the fact that for this modulation technique only the \((\alpha - \beta)\) subspace is controlled, the current harmonics from the phase currents spectra are considerably reduced compared with the conventional \(SVM\).

In fact, in both methods, the harmonic currents such as 5\(^{th}\), 7\(^{th}\), and 17\(^{th}\), etc. are inherently eliminated without extensive computations. The vector classification is simpler than the vector space decomposition technique from the point of view of implementation complexity, because the switching state vectors are mapped into the inverter basis frame rather than the machine basis frame. And the maximum modulation index increases by 3.59\% compared to vector decomposition technique. Consequently, the vector classification techniques can be easily implemented with low cost fixed-point \(DSP\) controllers.

Figure 4.11: vector classification technique
Chapter 5

Simulations

In the last chapter simulation results are presented. It will be firstly introduced the machine from which the data were taken, then the various blocks implemented in Matlab/SIMULINK. PID tuning procedure is explained along with motor mode operation. In the end, generator mode operation will be largely illustrated, explaining obtained results and how it was possible to conciliate SIMULINK signal logic with SimPowerSystem circuit logic in order to simulate efficiently power electronic and power system components.

5.1 Machine data

Before the presentation of the simulations in Matlab/SIMULINK it is better to describe the machine that was taken as reference in the work. In Table 5 are shown the most important parameters from which the others can be derived.

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal current</td>
<td>22.5 A</td>
</tr>
<tr>
<td>Nominal back-EMF L2L</td>
<td>363 V</td>
</tr>
<tr>
<td>Number of poles</td>
<td>8</td>
</tr>
<tr>
<td>Nominal speed</td>
<td>350 rpm</td>
</tr>
<tr>
<td>Stator resistance</td>
<td>0.53 Ω</td>
</tr>
<tr>
<td>$D_1$-axis inductance</td>
<td>39.1 mH</td>
</tr>
<tr>
<td>$Q_1$-axis inductance</td>
<td>49.1 mH</td>
</tr>
<tr>
<td>$D_2$-axis inductance</td>
<td>11.3 mH</td>
</tr>
<tr>
<td>$Q_2$-axis inductance</td>
<td>6.0 mH</td>
</tr>
</tbody>
</table>

Table 5: machine data
5.2 Machine model

The first step was to write in Matlab/SIMULINK the machine model starting from the derived machine equations. As it was explained in rotating \( d - q \) reference frame it is easy to control the machine because the torque depends just on the current component \( i_{q1} \), due to the particular transformation that was done on machine variables.

\[
\begin{align*}
v_{d1} &= R_s i_{d1} + L_{d1} p i_{d1} - \omega_e L_{q1} i_{q1} \\
v_{q1} &= R_s i_{q1} + L_{q1} p i_{q1} + \omega_e L_{d1} i_{d1} + \omega_e \Psi_{PM} \\
v_{d2} &= R_s i_{d2} + L_{d1} p i_{d2} - \omega_e L_{q2} i_{q2} \\
v_{q2} &= R_s i_{q2} + L_{q2} p i_{q2} + \omega_e L_{d2} i_{d2}
\end{align*}
\]

(3.28)

To use it, it is necessary to make the integral of the current’s derivatives in order to make SIMULINK solve the system of equations. The machine system is shown below in Figure 5.1 in which the different terms shown in the equations are highlighted.

![Figure 5.1 machine equations in Matlab/SIMULINK](image-url)
As can be seen the contributes due to back electromotive forces are treated as disturbances, this is because the only terms able to change the current (absorbed or delivered) are the voltages:

\[
\begin{align*}
    u_{d1} &= R_si_{d1} + pL_{d1}i_{d1} \\
    u_{q1} &= R_s i_{q1} + pL_{q1}i_{q1} \\
    u_{d2} &= R_s i_{d2} + pL_{d2}i_{d2} \\
    u_{q2} &= R_s i_{q2} + pL_{q2}i_{q2}
\end{align*}
\]  

In fact, writing the equations in the Laplace’s domain the transfer function (TF) of the input voltage (without considering disturbances) and the output current can be obtained:

\[
\begin{align*}
    I_{d1}(s) &= \frac{1}{R_s + sL_{d1}} U_{d1}(s) \\
    I_{q1}(s) &= \frac{1}{R_s + sL_{q1}} U_{q1}(s) \\
    I_{d2}(s) &= \frac{1}{R_s + sL_{d2}} U_{d2}(s) \\
    I_{q2}(s) &= \frac{1}{R_s + sL_{q2}} U_{q2}(s)
\end{align*}
\]  

(5.2)
5.3 Torque expression

The electromagnetic torque expression was written considering equation 3.34 which is:

\[
T_e = n_p \Psi_{d1}^* i_{q1} = n_p [\Psi_{PM} i_{q1} + i_{q1} i_{d1} (L_{d1} - L_{q1})]
\]  

(3.34)

The model in Matlab/SIMULINK is shown in Figure 5.3:

![Figure 5.3 torque expression](image)

5.4 PI regulator tuning

Once found the \( T_F s \), the next step was to find current and speed regulators parameters. In this work PI regulators were used and they were tuned using the Matlab application: Tuner.

5.4.1 Current regulator

As we said the currents (and so the torque) are controlled by the \( u \) voltages, the Figure 5.4 shows the scheme of a current regulator, in particular in this scheme it was considered that:

- Disturbances are compensated or not present
- The modulator is a unity gain

For a better representation of the power supply, the unity gain could be replaced with a delay whose value is tied to the switching time of the converter switches. If the modulator is based on a PWM technique, for example, the delay between the instant of the reference variation and its implementation varies from 0 to the entire period of switching. On average, we can assume a delay equal to one half the switching time.
The scheme in Figure 5.4 is representative of just one winding set. Due to the presence of second winding set it was necessary to do the procedure for both sets.

The procedure used to find $K_p$, $K_i$ of the regulator (for example for the $q_1$ axis) was:

- Insert the transfer function in Matlab:

\[
q\text{axes} = \frac{1}{0.0491s + 0.53}
\]

- Then the equation was open in PID Tuner, which allows to choose bandwidth and phase margin of the output in the frequency domain. Additionally, time domain response can be tuned as well in order to obtain a more robust and/or faster response. Below, the example of the application is presented, in which time domain and frequency domain responses are considered in order to find the optimal value of the two parameters.
The $PI$ parameters were chosen as a tradeoff between a correct and fast response and the maximum allowed value of the parameters themselves, in particular it is empirically known that the $K_i$ cannot overcome the value 10000. In the work the value of the $K_i$ parameters was maintained around 5000.

- The same was done for the other machine fictitious axis, in this way the regulator were tuned for both type of operation: motor and generator in order to obtain a good steady state and transient response, moreover the system has a good response to external disturbances.

### 5.4.2 Speed regulator design

As it was said the main focus of the work was the use of the machine as generator but, in order to verify the correctness of both the machine model and the regulator tuning process, the simulation was tested as machine operates as motor. To do so, it was necessary to design also the speed regulation loop and insert the mechanical equation, as shown in Figure 5.7.
The transfer function $F_i(s)$ corresponds to the presented current loop. During the tuning process of speed regulator, inner loop transfer function can be considered as a unity gain because the bandwidth of the current loop is near to 1000 $rad/s$ (a decade bigger than the current loop one).

![Figure 5.8 transfer function and sensitivity function](image)

This is possible because the bandwidth of the speed loop transfer function is confined in the zone in which the transfer function $F(j\omega)$ has a gain equal to 1. The function $F(j\omega)$ is the sensitivity transfer function of the loop and represents the transfer function between the output variable and the reference value of the input variable. The possibility of considering the current loop $TF$ equal to 1 is a good advantage because the regulation becomes easier, this method is called frequency decoupling.

### 5.5 Mechanical load equation

For motor machine operation the mechanical equation related to the load was written as well:

$$\omega_e = \frac{n}{B_v + sj} (T_e - T_m)$$  \hspace{1cm} (5.3)

Once obtained the electromagnetic torque, (see Figure 5.9) and subtracting the load torque (which is a disturbance) the rotational speed (in $el.rad/s$) can be obtained through the transfer function associated with the rotational masses: load + electrical motor. The two parameters are the inertia ($J$) of the two masses and the viscous friction coefficient ($B_v$), the chosen values for these two quantities are:

- $J = 0.065\, Kg\, m^2$
- $B_v = 0.21\, Nm\, s$
5.6 Park’s transformations

The last blocks needed are the Park’s transformations blocks. As it was said they transform the phase variable in Park component variables through a transformation. To obtain the transformation, the matrix \( T_3(\theta) \) presented in Chapter 3 was implemented in Matlab/SIMULINK and shown in Figure 5.10. The inverse transformation \( T_3^{-1}(\theta) \) is obtained by transposition of matrix 3.25.

\[
T_3(\theta) = \begin{pmatrix}
T_p(\theta + \alpha) & T_p(\theta - \alpha) \\
T_p(\theta + \alpha + \frac{\pi}{2}) & T_p(\theta - \alpha - \frac{\pi}{2})
\end{pmatrix}
\]  

(3.25)

It is important to notice that since this is the transformation in a rotating frame it takes as an input the rotational speed \( \omega_e \) and then through integration the \( \theta_e(t) \) value of the rotational position is obtained. The \( f(u) \) functions are the actual equations which transform the time variables in Park’s variables while the factor \( \frac{1}{\sqrt{3}} \) is needed to achieve power invariance.
5.7 Motor mode operation

5.7.1 System configuration

The equations and blocks needed are already presented, consequently it is possible to show the obtained results in the case of motor operation. As it was said, it was necessary to perform some simulations in order to verify if the control system worked correctly and if the machine had a behavior similar to the one taken as reference. The modulator wasn’t taken in exam because it was not important for the purpose of the first simulations.

Input variables were:

- The rated electrical speed, \( \omega_e \) which is \( \omega_{e\_nom} = n_p \omega_{m\_nom} \), where the nominal speed was given (350 rpm) and the number of poles \( (p) \) was given as well, and

- The rated resistant torque, which means that to obtain a steady state situation the motor must generates an equal and opposite torque, which is in turns the nominal one obtained through calculations.

In particular, the nominal torque value was obtained knowing that:

\[
P_{el\_nom} = \sqrt{3} E_{L2L} I_{nom} = 14146 W
\]  

(5.4)

From which can obtained the output mechanical power:

\[
\eta = \frac{P_{el\_nom} - 3 R_s I_{nom}^2}{P_{el\_nom}} = 0.943
\]  

(5.5)

Just Joule losses in stator windings were considered in the efficiency computation. Knowing the \( \eta \) can be write:

\[
P_{m\_nom} = \eta P_{el\_nom} = 13339 W
\]  

(5.6)

And, knowing that:

\[
P_{m\_nom} = T_{m\_nom} \omega_{m\_nom}
\]  

(5.7)

The nominal torque is obtained:
\[ T_{m,\text{nom}} = \frac{P_{m,\text{nom}}}{\omega_{m,\text{nom}}} = 364.5 \text{Nm} \]  

(5.8)

Considering this result we can also obtain the nominal \( i_{q,\text{nom}} \) able to generate the nominal torque:

\[ i_{q,\text{nom}} = \frac{T_{m,\text{nom}}}{n_p \psi_{PM}} = 63.75 \text{A} \]  

(5.9)

Where:

\[ \psi_{PM} = \frac{E_{L2L,\text{nom}}/\sqrt{3}}{n_p \omega_{m,\text{nom}}} = 1.4295 \text{Wb} \]  

(5.10)

\( \psi_{PM} \) is the flux of the permanent magnets chosen for the machine and it is chosen as a compromise between high performances and low manufacture’s costs. In this case, it can be observed that the machine rotor was made by IPMs, this is the reason why the torque expression presented in equation 3.34 has two contributes: anisotropy torque and excitation torque contribution. But considering the current \( i_{d,1} = 0 \) actually we treat the machine like an isotropic machine, in which we consider just the excitation torque component. This is done because interior permanent magnets are less expensive and easier to be realized with respect to surface permanent magnets, consequently this solution has been chosen. Then anisotropy torque component is made inefficient because it makes the control system more difficult; this operation, however does not affect too much the performances of the machine.

Figure 5.11: complete simulated model
5.7.1 Simulation results

As it is shown in Figure 5.12 The electromagnetic torque \( T_e \), in green follows very closely the mechanical resistance torque \( T_m \), in black dashed line the mechanical torque reference was given as a step from time 0s to time 4s, the insertion of the load was gradual to avoid high transient variations. Now, it is interesting to understand if with the nominal generated electromagnetic torque, the other quantities are as expected and if the control system works correctly. For example, in Figure 5.13 the electrical angular speed \((el.\text{rad/s})\) is shown. The reference speed at which the machine should rotate is the nominal one.

![Figure 5.12: mechanical vs electromagnetic torque](image)

It can be seen that the control system works correctly, after 4s when the resistant torque remains constant, the rotational speed of the machine follows the speed reference value. This result is possible only if the current \((q_1 \text{ axis})\) controller works efficiently as well. Figure 5.14 shows the
correctness of the most important current component \( (i_{q_1}) \), because as it was said the dynamic of the machine only depends on the value of this current.

![Graph of \( i_{q_1} \) and \( i_{q_1} \) vs time](image1)

**Figure 5.14: \( q_1 \) current component**

The current (in green) follows very closely the reference value (black), reference value which is the output of the speed controller in the first stage. Another important point to highlight is the current value at steady state, which is 65 \( A \), very near to the one we expected to find. This difference will bring to other differences in phase currents and in absorbed electrical power with respect to nominal values, due to dissipations that probably are not given in nominal data. Anyway the difference is not too high and the studied model can be considered reliable.

The correctness of the other three current controllers are verified and shown in the following figures.

![Graph of \( d_1 \) vs time](image2)

**Figure 5.15: \( d_1 \) current component**
The control works correctly and the illustrated model can be considered a good solution to schematize the machine behavior. In the second winding set it can be seen that there are not transient phenomena, this is right since the two currents \((i_{d2}, i_{q2})\) are both taken with a reference value equal to 0 and as expected they are not affected on what happens in the other winding set, due to decoupling. Instead as regards current \(i_{d1}\) it can be highlight that the presence of the motional term (disturbance) in the current equation affects its time variation. Anyway the effect of the disturbance is small and so the controllers are working correctly. The parameters of the \(PIs\) found in this test are the ones used later in the generator mode of operation.

The last important quantities to be checked are the phase currents, obtained anti-transforming the \(d - q\) currents absorbed by the fictitious machine and caused by the reference voltages obtained through the controllers, and the power: mechanical output power and electrical absorbed power. As for the current using the anti-transform block the following results were obtained (Figure 5.18), all the six phase currents are presented in the same plot to highlight the phase shift between currents of the same winding set and currents of different winding sets.
The first winding set currents (in blue) are shifted by 30 \(^\text{deg}\) with respect to currents of the second winding set. This result was expected, in fact, the windings are space displaced by 30 \(^\text{el. deg}\) which means that the magnetic flux generated by the \(\text{PMs}\) invests each coil of different winding sets after 30 \(^\text{el. deg}\). In the following plots, for clarity purposes, just the first winding set currents are shown.

![Figure 5.18: phase currents of the two winding sets](image)

The maximum value of the absorbed currents was expected too, in fact considering that the only current component taken different from 0 was the \(q_1\) component, using the \(T_3(\theta)\) it is obtained that:

\[
\begin{bmatrix}
i_{a_1} \\
i_{b_1} \\
i_{c_1} \\
i_{a_2} \\
i_{b_2} \\
i_{c_2}
\end{bmatrix} = T_3(\theta) \begin{bmatrix}0 \\ i_{q_1} \\ 0 \\ 0 \end{bmatrix}
\]

(5.11)

And knowing that the matrix \(T_3(\theta)\) introduces a coefficient \(\frac{1}{\sqrt{3}}\) it can be write (for the modulus of a generic phase current, in the example was taken current \(i_{a_1}\)):

\[
i_{a_1,\text{max}} = \frac{i_{q_1}}{\sqrt{3}} = 37.5 \text{ A}
\]

(5.12)

Which corresponds to a \(\text{rms}\) value of:
\[ i_{a_{1,rms}} = \frac{i_{a_{1,max}}}{\sqrt{2}} = 26.5 \text{ A} \] (5.13)

Which is a value very close to the nominal one, the slight difference is due to the problem that was already explained.

As concerns the power it was necessary to implement in *Matlab/SIMULINK* two different relations, the first for the input electrical power, which is:

\[ P(t) = v_{d1}i_{d1} + v_{q1}i_{q1} + v_{d2}i_{d2} + v_{q2}i_{q2} \] (3.

In which it is doubtless that the main contribute if not the only one is given by the term \( v_{q1}i_{q1} \), in fact as it was observed the other currents are equal to 0 or different to 0 just in the initial transient part.

![Figure 5.19: electrical power vs mechanical power](image)

Due to stator resistance, efficiency of the simulated machine is not 1, but it should be considered also that mechanical losses are here neglected. Even though they are not so large it is better to consider them in order to explain in a better way the real behavior of the machine. The steady state value considered as nominal was 15886 \( W \).

The second power relation, the one which provides the mechanical output power is:

\[ P_m = T_m \omega_m = T_e \frac{\omega_e}{n_p} \] (5.14)
Which gives, after the transient, a mechanical output power of 13641W.

To conclude it can be said that the model is quite accurate and the control system works correctly for the simulated cases which were considered. Now, it will be presented the main important case which is the generator operation.

5.8 Generator mode operation (modulator as unity gain)

5.8.1 System configuration

In generator operation the machine model was maintained identical to the previous case, changing the control structure and the input variables. In particular, in this operation mode, the electrical speed is decided by the prime mover, which is the wind turbine. It was imagined a system in which turbine blades were directly connected to the machine. The speed at which it rotates is decided from the turbine control system in order to extract the maximum wind power (as it was seen in Chapter 4). This part of the control system was neglected and it was considered that the speed which drives the two rotational masses (wind turbine blades and electrical generator) was the one which guarantees the maximum extracted power. The chosen speed was a ramp and it is shown in Figure 5.20, the rotational speed is increased from 0s to 4s, to avoid sudden stresses.

![Figure 5.20: rotational speed imposed by the prime mover](image_url)

The input variable was the electromagnetic torque, in the simulation the nominal one was taken in order to show the maximum extractable power from the wind turbine (if blades rotates at their nominal speed) but in real systems in which the electrical delivered power has to be controlled, the torque and the power vary considering the equation:
\[ P_m = T_m \omega_m \approx P_{el} \]  
\[ (5.14) \]

In which the speed hasn’t to be considered as a constant, in the evidence of what it was said before, i.e. the necessity to maintain \( c_p \) as high as possible with different wind speed situations.

![Figure: 5.21: generator simulation without modulator](image)

### 5.8.2 Simulation results

Obtained results are shown, the main purpose is to verify if the machine and control system work correctly in nominal operation conditions. To do so, modulator is not taken into account, meaning that the six leg inverter is considered as ideal (as like unity gains).

The chosen torque was \( T_{el,nom} = -364.5 \, Nm \), the torque increased after the speed reached its nominal value, this operation allows better performances of the control system and it was used also for the other simulation tests. The minus sign stands for the fact that with our signs conventions, electrical and mechanical power were taken as they were entering the system. In this way the electrical power was an outgoing power. Choosing the torque, means, in turn, choosing the voltages at machine terminals able to deliver the correct current and so generate the desired resistant electromagnetic torque. This will be shown in a better way when the modulator will be introduced.

As for now, considering again that the reference values of currents \( i_{d1}, i_{d2}, i_{q2} \) are taken equal to zero, the following results are shown:
As it can be seen the actual electromagnetic torque follows very well the reference value of the resistant electromagnetic torque. As it was largely said, since the electromagnetic torque depends on current $i_{q1}$ means that also that current has to follow its reference, which is obtained through the equation:

$$i_{q1,ref} = \frac{T_{el,ref}}{n_p \phi_{PM}}$$  \hspace{1cm} (5.9)

In fact, its dynamical behavior is:
Which is equal to the electromagnetic torque except for a scale factor. The other current components are now shown to illustrate the correctness of the control system that is used in the more important simulation, which is the one with modulator.

Figure 5.24: $i_{d1}$ current component

Figure 5.25: (a) $i_{d2}$ current component (b) $i_{q2}$ current component
For the current components $i_{d_2}, i_{q_2}$ the same comments made for motor mode operation are still valid. As concerns phase currents, it can be stated that imposing a nominal torque means to impose the nominal $rms$ current at the output of the machine. As it will be shown in the next simulation test, in the real machine this is achieved acting on power converter’s switches which are the electrical device able to control the terminal voltages of the machine and, consequently, they control the phase currents. The phase currents of the first winding set are shown in the figure below, which is Figure 5.26.

![Figure 5.26: phase currents in generator mode operation](image)

The results in Figure 5.26 were explained in the previous paragraph as regards motor mode operation. It can be interesting to note that in this operation mode the current is delivered instead of being absorbed.
5.9 Generator mode operation (complete model)

5.9.1 System configuration

The studied configuration is the one presented in Figure 5.28, in which the double star synchronous generator (coaxial with the wind turbine) is connected to the DC link through a six leg converter. It is considered that the grid side bus is an infinite bus, with the meaning that, this bus imposes the voltage and the frequency at its terminal.

In turns, have an infinite bus at the grid terminals means that the voltage at the DC bus is imposed by the grid, the only quantity which can be controlled is the current injected. (and so the active power) It also means that the six leg inverter behaves as if it was supplied with a DC voltage $V_{DC}$. Controlling the inverter switches, the machine terminal voltages are controlled and so the amount of current allowed to circulate and injected in the grid. In this configuration it is possible to work with unity power factor.
Under Rectifier mode operation, the converter works like Boost converter. The upper switch and lower diode work complementarily, the blue line shows the direction of current flow. Under Inverter mode of operation, the converter works like Buck converter. The lower switch and the upper switch and lower diode work complementarily which is shown by red line in Figure 5.29. The relationship between input voltage (assumed sinusoidal) and converter voltage, ignoring the harmonics, can be easily shown by a phasor. From Figure 5.30, the converter voltage is higher than the input voltage in both modes near unity power factor operation. Therefore, the $DC$ link voltage should be sufficiently larger than diode rectification voltage otherwise the converter will behave like common diode bridge.

![Figure 5.30: (a) rectifier mode operation (b) inverter mode operation](image)

Both in unity power factor operation

### 5.9.2 Description of the simulation

The modulator is here added to the simulation in order to illustrate in which way the real system would work. To do so, the signal logic has to be synchronized with the circuit logic introduced by $SimPowerSystems$. The $SimPowerSystems'$ library is needed because it allows to introduce electrical components, such as: resistances, inductors, capacitances, inverters, generators, and so on.

In particular, the inverter and the control system of the inverter itself has been added. The system becomes like the one shown in Figure 5.31 for motor mode operation. It is important is to note that now there is not a direct passage from the control system to the machine terminals but instead, the control system gives the reference values needed to generate the gate signals, which through the inverter can control the terminal voltages and consequently, the currents.
Let us consider now in more detail how the system works.

Starting from the control system illustrated in Figure 5.32, the reference value of the output currents are selected, except for \( q_1 \) axes in which the resistant electromagnetic torque, which is, as it was said before a scaled value of the \( q_1 \) current component.

In the real system, these voltages must be modulated by the inverter. To do so, the inverter receives the gate signals from the two \( PWM \) generators. The process is shown in Figure 5.33, the reference voltages, obtained from the regulation system are anti-transformed through anti-transform Park’s block, then the signals are modified (scaled to obtain a per unit signal) in order to be compared with the triangular waveform inside the block \( PWM \) generator, and so generate the gate signals which control opening and closing of the switches inside the inverter legs.
In this way, gate control signals $g_1$ and $g_2$ are obtained and sent to the six leg inverter which is already inside the circuit logic domain.

### 5.9.3 Six leg converter

The inverter receives the gate control signals $g_1$ and $g_2$. In this way, the inverter legs, generate the desired voltages at the machine terminals. These phase voltages are measured through the voltmeters and sent to the machine terminals using the block *go to* (blue, in Figure 5.34)

As it was said, the measured voltages are sent to the machine model, as shown in Figure 5.35, with the meaning that: with that specific combination of open and closed switches the delivered amount
of current is the one obtained through the mathematical model of the machine (i.e. solving the differential equations of the machine). Before introducing the voltages in the machine model they have to be transformed using the Park’s transformation block.

The obtained currents are the input of the controlled generators Figure 5.34 (in yellow), they are current generators current controlled. The control signals which drive the generators are in the signal domain (SIMULINK) while the generators themselves are in the circuit domain (SimPowerSystem). In this way, the two logics can be easily synchronized and advantages of their combined use can be exploited.

5.9.4 Simulation results

The results are shown, in which as before, the aim is to understand the correct behavior of the machine and control system. It was already proved the correctness of the machine model and the good behavior of the control system in the case in which the modulator was considered as unity gain. Now, it is important to understand if the complete system works correctly and if possibly the control system can be used with a real machine and a real six leg inverter. The input electromagnetic torque is a ramp with maximum value equal to $T_{\text{nom}} = -364.5 \, Nm$. 

Figure 5.35: machine model
In Figure 5.36 the imposed rotational speed is shown. With respect to the previous case the ramp was increased, it reached the nominal speed after 0.1\(s\), it was done to decrease the simulation time, the change does not affect the correctness of the simulation or the performances of the control system.

following the same steps used for the previous case, the first plot shows the time variation of the electromagnetic torque.

It can be seen that follows very closely its reference value. The next is the \(i_{q1}\) current component, shown in Figure 5.38:
And the other current components are:
The control system works correctly, the amplitude of the oscillations around the reference value are very low although they are present. Dynamical stresses in transient operation are not too high and the $PI_s$ regulation is fast. It can be said that if this control system would be applied to a real machine and inverter, it would give good results in terms of performances. However, it is important to show the behavior of the phase currents, which are the actual circulating current components, their time evolution is presented in Figure 5.42, for the first winding set.

These are the currents delivered by the machine, so the currents at the machine terminals. They are generated by the inverter which modulates the voltages, the next plots show these voltages, they are shown for the first winding set.
As it was already said in Chapter 4, PWM modulate the voltages in this way. Important to note is the magnitude and frequency of the first harmonic along with magnitude and frequency position of the other harmonics. Neglecting for now the issues related to higher order harmonics (which are not the main focus of the work) let us consider the first harmonic of the voltage $v_{a_1}$ (The black line in Figure 5.43).

For the sake of clarity was shown just a part of a period. The other phase voltages have the same time behavior as $v_{a_1}$ with a phase displacement due to winding position. To complete the set of the machine terminal’s quantities, phase current, phase voltage of phase $a_1$ and line to line voltage are shown together in the same plot (Just the first harmonic is presented).
Obviously $v_{ab1}$ was found as: $v_{ab1} = v_{a1} - v_{b1}$ and in fact its modulus is $\sqrt{3}$ times the modulus of $v_{a1}$. Another important data to take into account is the phase displacement between $v_{a1}$ and $i_{a1}$. It is desirable to have $cos \varphi = 1$.

From the figure can be seen that:

$$\Delta t = t_2 - t_1 = 0.005 \text{s}$$ (5.15)

Where

- $t_2$: time instant of the current peak
- $t_1$: time instant of the voltage peak

So, knowing that:

$$\omega_e \Delta t = \varphi = 0.733$$ (5.16)

It can be proved that the $cos \varphi$ is equal to 1. Precisely its value is 0.999.

The last important quantities remained to be shown are the DC ones. In particular, the most important is the delivered DC power which is the one injected into the grid. This power was computed as:

$$P_{DC} = V_{DC}I_{DC}$$ (5.17)
Where \( V_{DC} \), as it was said before, is fixed by the infinite bus and its value was taken equal to 750V.

Instead, in Figure 5.46 the direct current \( I_{DC} \) and its mean value are shown.

![Figure 5.46: DC side current](image)

Taking the mean value of the current \( I_{DC} \), the last Figure shows the two compared powers: mechanical input power and electrical output power.

![Figure 5.47: mechanical vs electrical power](image)
5.10 Generator mode operation with uncontrolled diode bridge

5.10.1 System configuration

To reduce the number of power converters, the idea was to substitute one of the two machine side converters with a diode bridge, as shown in Figure 5.48a. The three phase diode bridge has the advantage of being less expensive but it makes impossible to control the electrical quantities of the second winding set. The simulation was carry out to study the behavior of the system in this particular configuration and to understand if it was possible to find a way to mitigate the problems derived by the insertion of an uncontrolled component such as the diode bridge.

![Electrical circuit model and DC bus voltage regulation](image)

Figure 5.48: (a) electrical circuit model (b) DC bus voltage regulation

5.10.2 DC bus voltage regulation

To take into account the insertion of an uncontrolled element it was necessary a more detailed model of the grid converter: the \( DC \) voltage source was substituted by a controlled current source and a capacitance. The \( DC \) control system was tested using last system configuration, which is the one with both controlled power converters.

The control signal of the current source (\( I_{ref} \)) was decided by the control system. Figure 5.48b shows the voltage control loop and the generation of the correct \( I_c \) reference value able to maintain the voltage as constant as possible on its reference value. The capacitance was \( C = 100 \, mF \) and it was charged at \( 750 \, V \). Then, the capacitance current \( I_c \) is compared with the machine output current \( (I_{machine}) \), as shown in Figure 5.49b (the signs are arbitrary but they need to be chosen correctly).

The machine current delivered by the machine was decided through power considerations explained in Chapter 4.

\[
P_m = T_e \omega_m \cong V_{DC} I_{machine} = P_{DC} \tag{5.18}
\]
It wasn’t taken into account mechanical or electrical losses, this fact allows to find an easy expression for the $DC$ output current. From equation 5.18 the machine current can be obtained:

$$I_{machine} = \frac{T_e \omega_m}{V_{DC}}$$  \hspace{1cm} (5.19)

As it was said, $I_{machine}$ is compared with the $I_c$ and the signal sent to the modulator (which is here considered as ideal). The grid side power converter allowed the correct $I_{ref}$ to circulate (taking into account the machine current $I_{machine}$) in order to let $I_c$ through the capacitor. The $DC$ bus voltage was controlled and it follows its reference in all machine operation conditions, it can be said that the voltage control worked correctly, as shown in Figure 5.50.

The system was completely decoupled, grid side variables can be considered independent from the machine side variables.
5.10.3 Simulation

The system configuration shown in Figure 5.48a brought into light difficulties in the prediction of the machine behavior along with different cases in which the diode bridge didn’t work correctly. This situation was anyway expected because, in order to be able to work, power converter needs a DC bus voltage greater or equal than $\sqrt{2}V$, which is the maximum value of the machine voltage waveform. While, the diode bridge needs the DC bus voltage below that value, otherwise diode is not able to conduct. It was decided to consider another system configuration. The two winding sets were divided and connected to two different voltage sources (the DC regulation wasn’t taken into account for this purpose). The second winding set DC bus voltage was set at 150V. With this particular system configuration the following results were obtained:

- The reference torque was equal to $-100\, Nm$, this value was chosen to avoid overload of one of the two machine stars. The actual electromagnetic torque followed the fixed reference value, as shown in Figure 5.51.

![Figure 5.51: electromagnetic torque (with uncontrolled diode bridge)](image)

- The obtained result for the electromagnetic torque is a consequence of a correct work of the control system on the fictitious current components: $i_{d1}$ and $i_{q1}$. In fact, thankfully, the control system succeeded in the current regulation. The results obtained for the torque and these two current components are good starting point for the creation of a newer and more performing control system. In Figures 5.52 and 5.53 $i_{d1}$ and $i_{q1}$ are shown.
As it was expected current components \( i_{d_2} \) and \( i_{q_2} \) were out of control but they never exceed dangerous values and achieved the steady state their value was constant.

- Last important quantities to show are the absorbed mechanical power and delivered \( DC \) electrical powers.

The obtained results were expected and not so terrible but in the end the system didn’t work in acceptable way. This is why it was necessary to find a better way to control the system.
5.10.4 Innovative control set up

As already said, the two power electronic components cannot work with the same DC bus voltage level. Additionally, Figure 5.54 shown that without an appropriate control strategy it is impossible to achieve good performances of the system. In fact, with those initial data the diode bridge delivered most of the output power, while the power converter carried a very low share of the total power. A solution was found using the configuration set up shown in Figure 5.55a, in which DC buses were maintained separated but just the one connected to the diode bridge was controlled. The other three phase set is thought to be in steady state.

![Figure 5.55: (a) system configuration (b) DC bus regulation](image)

The current controlled current source, as before, represents the grid side power converter (modulator wasn’t taken into account). The idea is to use the grid power converter to regulate the diode bridge delivered current acting on the DC bus voltage level. This can be achieved because \( P_{DC} = V_{DC}I_{DC} \). As soon as the delivered power can be considered constant, voltage and current are inversely proportional, thus varying \( V_{DC} \), it is possible to regulate the delivered current.

![Figure 5.56: corrective term control system](image)
The result, however, is a not perfectly constant value of the \( DC \) bus voltage \( V_{DC_2} \). Using this control strategy the most important part is the choice of the correction term, in Chapters 4 and 5 was explained that wind turbine system cannot always deliver the nominal power due to varying wind speed or grid requirements. So, the reference \( DC \) current value for the bridge connected three phase winding was chosen in an easy way: it was half the current demanded to the controlled winding set, which is:

\[
I_{\text{machine}_\text{ref}} = \frac{P_m}{2V_{DC_1}} = \frac{T_e \omega_m}{2V_{DC_1}}
\]

In this way, it was thought that the two winding sets would share the same current amount avoiding overload and ensuring good performances in term of delivered power. Then, \( I_{\text{machine}_\text{ref}} \) is sent to a \( PI \) regulator, the output of this control loop is the corrective term to add or subtract from the \( DC \) voltage bus reference value. Then, through another control loop taking into account the already present capacitance voltage and the instantaneous machine power, the correct signal is chosen to be sent to the grid side converter’s modulator.
Conclusions and further works

The work presented many aspects of the so called double star synchronous machine. In particular, some applications were discussed in which it resulted that this kind of machine has a lot of advantages in high power applications. The main drawbacks are: the higher costs in relation to classical three phase configurations and higher complexity of the control system. To simplify the control system algorithm, the mathematical model presented all along Chapter 3 was implemented and the illustrated solution seems to be a correct way to schematize the machine, in fact obtained results for motor mode operation, in terms of steady state quantities were very similar to the ones obtained through measures on the real machine. The particular Park’s transformation applied in this work allowed a great simplification in control system design. Simulations were carried out for the main focus system configuration, which is wind turbine systems: the machine was thought fixed with the wind turbine blades (gearless system) which rotated at nominal speed and the electrical machine was demanded to deliver the nominal active power. In the first configuration made by one power converter for each winding set, the results were very promising, also considering introduced simplifications. In the second system configuration, which was with just one power converter and a diode bridge, results were unstable but the possibility of using the grid converter to control the diode bridge dynamical behavior makes this cheaper configuration and its control system very interesting.
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