Study and implementation of a PMSM & Study of a sensorless control method

REPORT

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Deadline: June 2017

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Abstract

The procedure followed in the project begins with a brief introduction of the features that the studied motor, a permanent magnet synchronous motor (PMSM), has. The fact that the motor is synchronous permanent magnet has to do with its greater efficiency in comparison with the induction motor, which is the most used nowadays.

Then, the project is conducted in two steps. The first one is the study of the PMSM mathematical modelling and the subsequent control method applied. The second one is the study of a sensorless control algorithm. Traditionally for speed dependent applications, some kind of sensor is used to read the motor speed and position and feed the value back to the controller. However, extra sensors require extra physical space in the application and it also introduces another source of failure in the system. Thus, with the additional purposes of reducing cost and maintenance needs, the sensor can be replaced by an estimator that mathematically estimates the speed or position of the rotor.

All these implementations have been simulated with MATLAB/Simulink based on the mathematical models.

To design a controlled drive, the stability characteristics of PMSM under open-loop control (without having any feedback for speed) are analysed. The analysis shows that the PMSM becomes unstable after exceeding a certain applied speed. After tuning the controllers, it has been analysed that the maximum speed that the closed-loop can control with a reasonable settling time is 750 rpm. The more speed that the motor achieves, the more settling time appears. Thus, there is an upper limit for the speed. For all the simulations, an optimal speed of 550 rpm has been used. The control structure and the design of the controllers are described.

A rotor position estimation technique for sensorless operation is studied. The estimator uses predictor-corrector method where the difference between the estimated current and the measured current (current error) is used to correct a predicted rotor position. More investigations are still required for accurate rotor position estimation.
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1. Glossary

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1.2. Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Unity</th>
<th>Meaning</th>
</tr>
</thead>
<tbody>
<tr>
<td>( B_m )</td>
<td>[Nm s rad(^{-1})]</td>
<td>Viscous friction coefficient</td>
</tr>
<tr>
<td>( I_a, I_b, I_c )</td>
<td>[A]</td>
<td>Phase a, b, c instantaneous stator current</td>
</tr>
<tr>
<td>( I_d, I_q )</td>
<td>[A]</td>
<td>(d)- and (q)- axis components of stator current</td>
</tr>
<tr>
<td>( J )</td>
<td>[kg m(^2)]</td>
<td>Inertia of the rotating system</td>
</tr>
<tr>
<td>( K_d )</td>
<td>[-]</td>
<td>Derivative constant</td>
</tr>
<tr>
<td>( K_i )</td>
<td>[-]</td>
<td>Integral constant</td>
</tr>
<tr>
<td>( K_p )</td>
<td>[-]</td>
<td>Proportional constant</td>
</tr>
<tr>
<td>( L_d, L_q )</td>
<td>[H]</td>
<td>(d)- and (q)- axis stator self-inductances</td>
</tr>
<tr>
<td>( L_s )</td>
<td>[H]</td>
<td>Average inductance</td>
</tr>
<tr>
<td>( L_x )</td>
<td>[H]</td>
<td>Inductance fluctuation</td>
</tr>
<tr>
<td>( P )</td>
<td>[-]</td>
<td>Number of poles of the motor</td>
</tr>
<tr>
<td>( R_s )</td>
<td>[Ω]</td>
<td>Stator resistance</td>
</tr>
<tr>
<td>( s )</td>
<td>[-]</td>
<td>Laplace operator</td>
</tr>
</tbody>
</table>
Study of a PMSM

\( T \) [Nm] Torque

\( T_L \) [Nm] Load torque

\( t_r \) [s] Settling time

\( T_s \) [s] Sampling time

\( V_a, V_b, V_c \) [V] Phase a, b, c instantaneous stator voltage.

\( V_{d}, V_{q} \) [V] \( d \) and \( q \)-axis components of stator phase voltage

\( \alpha \) \([s^{-1}]\) Modulator constant

\( \theta \) \([^\circ]\) Electrical angle between \( a \)-axis and \( q \)-axis

\( \theta_r \) \([\text{rad}]\) Electrical rotor position

\( \lambda_{ds}, \lambda_{qs} \) \([\text{V s rad}^{-1}]\) Stator flux linkage in rotor fixed \( d,q \) frame

\( \lambda_m \) \([\text{V s rad}^{-1}]\) Peak flux linkage due to permanent magnet

\( \omega \) \([\text{rad/s}]\) Angular velocity of rotation

\( \omega_m \) \([\text{rad/s}]\) Machine’s mechanical speed

\( \omega_r \) \([\text{rad/s}]\) Machine's electrical rotor speed
1.3. Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
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<tbody>
<tr>
<td>AC</td>
<td>Alternating Current</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>EMF</td>
<td>Electromotive Force</td>
</tr>
<tr>
<td>EV</td>
<td>Electric Vehicle</td>
</tr>
<tr>
<td>FOC</td>
<td>Field Orientated Control</td>
</tr>
<tr>
<td>IPMSM</td>
<td>Interior Permanent-Magnet Synchronous Motor</td>
</tr>
<tr>
<td>PMSM</td>
<td>Permanent-Magnet Synchronous Motor</td>
</tr>
<tr>
<td>SC</td>
<td>Scalar Control</td>
</tr>
<tr>
<td>SPMSM</td>
<td>Surface Permanent-Magnet Synchronous Motor</td>
</tr>
<tr>
<td>VC</td>
<td>Vector Control</td>
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</table>
2. Preface

Many types of electric motors have been used in the industry for different purposes: cranes, spinning machines, public transportation and so on. AC motors are widely used and AC drives are subject of study for many researchers. Recently, AC drives in vehicle applications are gaining attention due to pollution and fuel price problems. [13]

A large part of these applications are triggered by induction motors for their ease of use and their presence throughout history. But with the technology currently available these can be replaced by permanent magnet synchronous motors in order to improve their efficiency and performance. [13]

Apart from replacing induction motors for permanent magnet synchronous motor, more improvements can be done. The rotational speed of a PMSM is commonly measured by sensors. High precision sensors are expensive, occupy space and the total complexity of the motor drive is increased. Sensorless control provides the option of reducing the size and complexity of the motor drive, and consequently, the price is also reduced. [5]
3. Introduction

AC electrical motors are widely used in industry. An AC motor is a machine that converts alternating current (AC) into torque, contrary to the DC motor, which produces torque from direct current. Transportation (like trains or cars), washing machines and industrial cranes are some applications where AC motors are used.[3]

There are two main classifications of AC motors: induction and synchronous. The difference is that for an induction motor, currents are induced in the rotor windings whenever the speed of the rotor differs from the speed of the rotating magnetic field generated from the currents in the stator windings. While for synchronous motors, the rotor always rotates at the same speed as the rotating magnetic field. The synchronous speed can be achieved by either injecting current into the rotor or using permanent magnets in the rotor. Of the two, induction motors are the most popular in industry, mainly because they are small, robust and cheap. [3]

One particular synchronous motor, the PMSM, is however gaining in popularity due its better power/mass ratio and higher efficiency. Due to the permanent magnets, there are no energy losses in the rotor. However, the PMSM is still more expensive and thus mostly used for high performance applications.[5]

A mature and well-used control strategy for PMSMs is the Field Orientated Control (FOC). The idea with FOC is to separately control the motor flux and torque. This is done by transforming the three stator currents, represented in a three-coordinate reference frame, into a rotating two-coordinate reference frame. To do this, the exact position of the rotor needs to be known. The two current components are then controlled independent on each other. The control output, that is the new motor voltage command, is then transformed back and instructs the voltage inverter to produce the sinusoidal voltages that will be fed to the motor. This way, FOC makes the AC control behave like a DC control.[15]

Instead of using sensors, which have some disadvantages such as the requirement of extra physical space, a sensorless control method will be used. Thus, the sensor will be replaced by a mathematical estimator that estimates the speed or position of the rotor from the given actual rotor position. This method will reduce costs and maintenance needs. [1]

3.1. Objectives of the project

The main aim of this project, therefore, is to study the mentioned motor, implement it with MATLAB/Simulink and tune the controllers of the motor. Another objective is to analyse the sensorless algorithm and understand its great importance. The objective of this sensorless
control method is to control the motor without any sensor position with the objective of reducing the equipment costs and the occupied space, as well as increasing reliability, because sensor positions can lose precision when working in extreme circumstances such as high temperatures, and can lose faculties over time.

3.2. **Scope of the project**

The scope of the present project is summarized in the following points:

- Description of the permanent magnet synchronous motor
- Modelling the permanent magnet synchronous motor
- Control the permanent magnet synchronous motor
- Analysis of the motor response once the control is implemented
- Description of the sensorless control algorithm
4. PMSM Model

4.1. Introduction

The Permanent Magnet Synchronous Motor (PMSM) is an AC synchronous motor whose field excitation is provided by permanent magnets, and has a sinusoidal back electromotive force (EMF) waveform. [3]

Some advantages of this motor are that with permanent magnets the PMSM can generate torque at zero speed; comparing it to AC induction motors, it has a higher torque density. Moreover, it presents high efficiency operation. [1]

In general, PM synchronous machines with approximately sinusoidal back-EMF can be broadly categorized into two types: 1) interior (or buried) permanent magnet motors (IPM) and 2) surface-mounted permanent magnet motors (SPM). In the first category, magnets are buried inside the rotor. Due to this interior-permanent structure, the equivalent air gap is not uniform and it makes saliency effect obvious, although the IPM motor physically looks like a smooth-air-gap machine. As a result, the quadrature-axis synchronous inductance of IPM is larger than its direct-axis inductance \( (L_q > L_d) \), which significantly changes the torque production mechanism. Therefore, both magnetic and reluctance torque can be produced by IPM motor. In the second category, the magnets are mounted on the surface of the rotor so the stator inductances do not depend on the rotor position. Because the incremental permeability of the magnets is 1.02-1.20 relative to external fields, the magnets have high reluctance and accordingly the SPM motor can be considered to have a large and uniform effective air gap. This property makes the saliency effect negligible. Thus the quadrature-axis synchronous inductance of SPM equals its direct-axis inductance \( (L_q = L_d) \). As a result, only magnet torque can be produced by SPM motor, which arises from the interaction of the magnet flux and the quadrature-axis current component of stator currents. Compared to SPM motors, the IPM motor has a mechanically robust and solid structure since the magnets are physically contained and protected. In addition, due to their reluctance torque production, the IPM motors are more suitable for traction applications, which requires constant power output at high speeds over a wide range. However, it is more expensive to manufacture and more complex to control. [3]

The IPMSM type rotor configuration brings some characteristics to the machines, which are not present in a SPMSM. As shown in Figure 1, the magnetic flux induced by the magnets defines the rotor direct \( d' \)-axis radially through the centreline of the magnets. The rotor quadrature \( q' \)-axis is orthogonally (90 electrical degrees) placed with rotor \( d' \)-axis (for four-pole design this is 45 mechanical degrees as shown in Figure 1 (b)). Since the permeability
of PM is almost same as the air, in IPMSM type configuration the effective air-gap of \(d'-\)axis is increased compared to the \(q'-\)axis. Therefore, the \(d'-\)axis reluctance is higher than the \(q'-\)axis reluctance. This results to the fact that the \(q'-\)axis inductance is higher than the \(d'-\)axis inductance in IPMSM \((L_q > L_d)\). [1]

![Image](a) SPMSM (b) IPMSM

**Figure 1: Different rotor configurations for PMSM. (a) SPMSM (b) IPMSM**

For the case study of this project an IPMSM motor type is used and it will be the only type studied. Therefore, for the rest of this report, when referring to a PMSM it is the IPMSM type that is meant if nothing else is stated.

Since PMSMs are synchronous machines, the accurate torque can be produced in these machines only when the AC excitation frequency is precisely synchronized with the rotor frequency. Thus, the fundamental requirement in control design of PMSMs is the assurance of precise synchronization of machine’s excitation with the rotor frequency. The direct approach to achieve this requirement is the continuous measurement of the absolute rotor angular position and the excitation of the machine. This concept is also known as self-synchronization and it assures that the PMSM does not go out of synchronization during operation. [8]

Some applications of this type of motor are:

- Air conditioner & refrigerator (AC) compressors
- Direct-drive washing machines
- Automotive electrical power steering
- Machining tools
- Traction control
4.2. Mathematical Model

In order to study analytically the permanent magnet synchronous motor, it is necessary to build a mathematical model of it, so that the equations relate all the variables of the motor and describe the motor’s behaviour at a given operation point.

In point 4.2 this mathematical model is presented. From this model, the ideal behaviour of the motor can be simulated in order to adjust the appropriate control algorithm that satisfies the application needs.

In a PMSM, the two phase (d,q) equivalent circuit model is commonly used for simplicity and intuition. Using the 2-phase motor model reduces the number of equations and simplifies the control design. [9]

The first step is to implement the mathematical model of the motor without the control; it means an open loop system.

Nevertheless, before that, an equivalent 2-phase circuit model of a 3-phase IPM machines is derived in order to clarify the concept of the transformation and the relation between 3-phase quantities and their equivalent 2-phase quantities.

![Figure 2: PM Synchronous Motor](image-url)
The following parameters are defined:

\( P \)  
Number of poles of the motor

\( I_a, I_b, I_c \)  
Phase a, b, c instantaneous stator current

\( V_a, V_b, V_c \)  
Phase a, b, c instantaneous stator voltage.

\( I_d, I_q \)  
d- and q- axis components of stator current

\( V_d, V_q \)  
d and q- axis components of stator phase voltage

\( R_s \)  
Stator resistance

\( p \)  
\( \frac{d}{dt} \)

\( L_d, L_q \)  
d- and q- axis stator self-inductances

\( L_s \)  
Average inductance \( L_s = \frac{1}{2} \cdot (L_q + L_d) \)

\( L_x \)  
Inductance fluctuation \( L_x = \frac{1}{2} \cdot (L_q - L_d) \)

\( \lambda_m \)  
Peak flux linkage due to permanent magnet

\( \theta \)  
Electrical angle between a-axis and q-axis in degrees

\( \omega = p \cdot \theta \)  
Angular velocity of rotation (in electrical rad/sec)

Figure 2 illustrates a conceptual cross-sectional view of a 3-phase, 2-pole IPM synchronous motor along with two reference frames. To illustrate the inductance difference \( (L_q > L_d) \), rotor is drawn with saliency although actual rotor structure is more likely a cylinder. The stator reference axis for the a-phase is chosen to the direction of maximum magneto-motive force when a positive a-phase current is supplied at its maximum level. Reference axis for b- and c- stator frame are chosen 120° and 240° (electrical angle) ahead of the a-axis, respectively. Following the convention of choosing the rotor reference frame, the direction of permanent magnet flux is chosen as the d-axis, while the q-axis is 90 degrees ahead of the d-axis. The angle of the rotor q-axis with respect to the stator a-axis is defined as \( \theta \). Note that as the machine turns, the d-q reference frame is rotating at a speed of \( \omega = \frac{d\theta}{dt} \), while the stator a-, b-, c- axes are fixed in space. [9]
The electrical dynamic equation in terms of phase variables can be written as:

\[ V_a = R_s \cdot I_a + p \cdot \lambda_a \]  
(4.1)

\[ V_b = R_s \cdot I_b + p \cdot \lambda_b \]  
(4.2)

\[ V_c = R_s \cdot I_c + p \cdot \lambda_c \]  
(4.3)

while the flux linkage equations are

\[ \lambda_a = L_{aa} \cdot I_a + L_{ab} \cdot I_b + L_{ac} \cdot I_c + \lambda_{ma} \]  
(4.4)

\[ \lambda_b = L_{ab} \cdot I_a + L_{bb} \cdot I_b + L_{bc} \cdot I_c + \lambda_{mb} \]  
(4.5)

\[ \lambda_c = L_{ac} \cdot I_a + L_{bc} \cdot I_b + L_{cc} \cdot I_c + \lambda_{mc} \]  
(4.6)

Considering symmetry of mutual inductances such as \( L_{ab} = L_{ba} \). Note that in the above equations, inductances are functions of the angle \( \theta \). Since stator self-inductances are maximum when the rotor q-axis is aligned with the phase, while mutual inductances are maximum when the rotor q-axis is in the midway between two phases. Also, note that the effects of saliency appeared in stator self and mutual inductances are indicated by the term \( 2\theta \). [9]

\[ L_{aa} = L_{so} + L_{sl} + L_x \cdot \cos(2\theta) \]  
(4.4)

\[ L_{bb} = L_{so} + L_{sl} + L_x \cdot \cos(2\theta + 120) \]  
(4.5)

\[ L_{cc} = L_{so} + L_{sl} + L_x \cdot \cos(2\theta - 120) \]  
(4.6)

\[ L_{ab} = -\frac{1}{2} L_{so} + L_x \cdot \cos(2\theta - 120) \]  
(4.7)

\[ L_{bc} = -\frac{1}{2} L_{so} + L_x \cdot \cos(2\theta) \]  
(4.8)

\[ L_{ac} = -\frac{1}{2} L_{so} + L_x \cdot \cos(2\theta + 120) \]  
(4.9)

For mutual inductances in the above equations, the coefficient \(-\frac{1}{2}\) comes due to the fact that stator phases are displaced by 120°, and \( \cos(120) = -\frac{1}{2} \). Meanwhile, flux-linkages at the
stator windings due to the permanent magnet are

\[ \lambda_{ma} = \lambda_m \cdot \cos(\theta) \]  
\[ \lambda_{mb} = \lambda_m \cdot \cos(\theta - 120) \]  
\[ \lambda_{mc} = \lambda_m \cdot \cos(\theta + 120) \]

For this model, input power \( P_i \) can be represented as

\[ P_i = V_a \cdot I_a + V_b \cdot I_b + V_c \cdot I_c \]  

Unfortunately, the output power \( P_o \) and the output torque \( T = \left( \frac{P}{\omega} \right) \cdot \frac{P_o}{\omega} \) cannot be simply derived in this 3-phase model. The torque can be expressed as

\[ T = \left( \frac{P}{\omega} \right) \cdot (L_q - L_d) \cdot \left[ \{(I_a^2 - 0.5I_b^2 - 0.5I_c^2 - I_a \cdot I_b - I_a \cdot I_c + 2 \cdot I_b \cdot I_c) \cdot \sin(2\theta) \right. \\
+ \left. \sqrt{3} \cdot (I_b^2 + I_c^2 - 2 \cdot I_a \cdot I_b + 2 \cdot I_a \cdot I_c) \cdot \cos(2\theta) \right] + \lambda_m \\
\cdot \{(I_a - 0.5 \cdot I_b - 0.5 \cdot I_c) \cdot \cos(\theta) + \sqrt{3} \cdot (I_b - I_c) \cdot \sin(\theta)\} \]

Now, let \( S \) represent any of the variables (current, voltage and flux linkage) to be transformed from the a-b-c frame to d-q frame. The transformation in matrix form is given by

\[
\begin{bmatrix}
S_q \\
S_d \\
S_0
\end{bmatrix} = \frac{2}{3} \begin{bmatrix}
\cos(\theta) & \cos(\theta - 120) & \cos(\theta + 120) \\
\sin(\theta) & \sin(\theta - 120) & \sin(\theta + 120) \\
0.5 & 0.5 & 0.5
\end{bmatrix}
\begin{bmatrix}
S_a \\
S_b \\
S_c
\end{bmatrix}
\]

Here \( S_0 \) component is called the zero sequence component, and under balanced 3-phase system this component is always zero. Since it is a linear transformation, its inverse transformation exists and is

\[
\begin{bmatrix}
S_a \\
S_b \\
S_c
\end{bmatrix} = \frac{2}{3} \begin{bmatrix}
\cos(\theta) & \sin(\theta) & 1 \\
\cos(\theta - 120) & \sin(\theta - 120) & 1 \\
\cos(\theta + 120) & \sin(\theta + 120) & 1
\end{bmatrix}
\begin{bmatrix}
S_q \\
S_d \\
S_0
\end{bmatrix}
\]

Now, by applying the transform of Equation 4.15 to voltages, flux-linkages and currents of Equations 4.1-4.6, we get a set of simple equations as
\[ V_q = R_s \cdot I_q + p \cdot \lambda_q + \omega \cdot \lambda_d \]  \hspace{1cm} (4.16)

\[ V_d = R_s \cdot I_d + p \cdot \lambda_d - \omega \cdot \lambda_q \]  \hspace{1cm} (4.17)

where

\[ \lambda_q = L_q \cdot I_q \]  \hspace{1cm} (4.18)

\[ \lambda_d = L_d \cdot I_d + \lambda_m \]  \hspace{1cm} (4.19)

Here, \( L_q \) and \( L_d \) are called q- and d-axis synchronous inductances, respectively, and are defined as

\[ L_q = \frac{3}{2} \cdot (L_{s0} + L_x) + L_{sl} \]  \hspace{1cm} (4.20)

\[ L_d = \frac{3}{2} \cdot (L_{s0} - L_x) + L_{sl} \]  \hspace{1cm} (4.21)

As noticed in the above equation, synchronous inductances are effective inductances under balanced 3 phase conditions. Each synchronous inductance is made up of self-inductance (which includes leakage inductance) and contributions from other 2 phase currents. Now, a more convenient equation may result by eliminating the flux-linkage terms from equations 4.16-4.17 as

\[ V_q = (R_s + L_q \cdot p) \cdot I_q + \omega \cdot L_d \cdot I_d + \omega \cdot \lambda_m \]  \hspace{1cm} (4.22)

\[ V_d = (R_s + L_d \cdot p) \cdot I_d - \omega \cdot L_q \cdot I_q \]  \hspace{1cm} (4.23)

These two equations are the final electric equations of the model.

Figure 3 shows a dynamic equivalent circuit of an IPMSM based on equations 4.22 and 4.23. Note that in practice, magnetic circuits are subject to saturation as current increases. Especially, when \( I_q \) is increased, the value of \( L_q \) is decreased and \( \lambda_m \) and \( L_d \) is subject to armature reaction. Since \( I_d \) is maintained to zero or negative value (demagnetizing) in most operation conditions, saturation of \( L_d \) rarely occurs. [9]
The torque produced $T$, which is power divided by mechanical speed, can be represented as

$$T = \left(\frac{3}{2}\right) \cdot \left(\frac{P}{2}\right) \left(\lambda_m \cdot I_q + (L_d - L_q) \cdot I_q \cdot I_d\right) \quad (4.24)$$

and defines the electromechanical conversion of the model.

It is apparent from the above equation that the produced torque is composed of two distinct mechanisms. The first term corresponds to “the mutual reaction torque” occurring between $I_q$ and the permanent magnet, while the second term corresponds to “the reluctance torque” due to the differences in d-axis and q-axis reluctance (or inductance). Note that in order to produce additive reluctance torque, $I_d$ must be negative since $L_q > L_d$. [9]

The relationship among the machine produced electromagnetic torque $T_e$, load torque $T_l$ and the machine’s electrical speed $\omega_r$, gives the mechanical equation of the machine and it can be expressed as

$$T_e = J \cdot \frac{2}{n} \cdot p \cdot \omega_r + B_m \cdot \frac{2}{n} \cdot \omega_r + T_l \quad (4.25)$$

where $J$ is the inertia of the rotor and the connected load, and $B_m$ is the viscous friction coefficient.

### 4.3. Implementation and results

#### 4.3.1. Open-loop system

Equations 4.22, 4.23, 4.24 and 4.25 have been implemented on MATLAB/Simulink.

Apart from these equations, it is clear that the open-loop system needs energy to work. Thus, a battery that provides the enough voltage has been added. This voltage is a DC voltage and its value is $400\sqrt{3} \, V$. As the motor works in AC, a power converter DC/AC must be added.
(see second block of Figure 4). The alpha parameter will be later implemented in the control and it is defined as $\alpha = \frac{U_a}{U_e}$, where $U_e$ is the battery voltage and $U_a$ is the voltage of the motor after the whole closed-loop system. The power converter is connected to the electric motor and the combination of the two is referred to as an electric drive.

![Figure 4: Open-loop system](image)

See Appendix C to see the inside of each block.

Nevertheless, with only these five blocks, the speed depends directly from the introduced voltage (as well as the fixed parameters of the motor and the load torque) and as a consequence it cannot be controlled. That is why a close-loop system is added: the control.

Before proceeding to implement the control in the model, it was proved that from fixed voltages $V_d = -178\, V$, $V_q = 284\, V$ and with the load torque $T_l = 12\, Nm$ the motor speed was the expected: $\omega_r = 550\, rpm$; see Figure 5 below.
Figure 5: Motor speed during open-loop system

Figure 6: Direct current and quadrature current

The currents values when stabilized are

\[ i_d = 0 \text{ A} \]

\[ i_q = 5.7 \text{ A} \]

The \( i_d \)-reference has been set to zero, since there is no flux weakening operation.
4.3.2. Closed-loop system

In the realm of motion control, the task of controlling the speed of a moving object or tool is frequently encountered. The actual speed should be made equal to the set speed. The difference between the actual and set speed is known as the speed error. It is the task of the speed controller to keep the speed error as small as possible, preferably equal to zero. To achieve this result, the controller generates the torque reference. [2]

After having analysed the open-loop system, the control part which will close the system will be then implemented.

The control algorithms are divided into two general groups. The first group is called scalar control (SC). The other group is called vector control (VC), or field-oriented control (FOC). The FOC technique brings over all improvements in drive performance when compared to the scalar control (higher efficiency, full torque control from zero to nominal motor speed, decoupled control of flux and torque, improved dynamics, etc.). [16]

Therefore, high performance motor control is characterized by smooth rotation over the entire speed range of the motor, full torque control at zero speed, and fast accelerations and decelerations. To achieve such control, the Field-Orientated Control (FOC) technique for 3-phase AC motors is used. The basic idea of the FOC algorithm is to decompose the stator current into the magnetic field-generating part and the torque-generating part. Both components can be controlled separately after the decomposition. The structure of the motor controller is then as simple as that for separately excited DC motors. [16]

FOC technique is able to control the field and torque of the motor separately. The aim of the control is to regulate motor speed.

The control structure of the motor is directly deduced from the yet implemented mathematical equations of the motor.

PID controllers will be used. A PID controller continuously calculates an error value $e(t)$ as the difference between a desired setpoint and a measured process variable and applies a correction based on proportional, integral, and derivative terms. The general structure of a PID controller is shown in Figure 7. [12]
Figure 7: Block diagram of a PID controller in a feedback loop

where \( r(t) \) is the desired process value, \( y(t) \) is the measured process value and \( e(t) \) is the error \( e(t) = r(t) - y(t) \).

In this model, three PI controllers will be used, two in the electrical part (d and q currents) and the other one in the mechanical system (speed).

In proportional control, adjustments are based on the current difference between the actual and desired speed, while in integral control, adjustments are based on recent errors. [11]

Note that there is no derivative part. The reason is that in the derivative part of a controller, the noise affects the system, making that these small variations of the input influence in a bigger scale in the output. [11]

The function of the derivative action is to maintain minimum error correcting it proportionally at the same speed that occurs, thus avoiding the error increase. Nevertheless the derivative action occurs when there is a change in the absolute value of the error (if the error is constant, as in this case, only the proportional and the integral part act). [11]

The derivative action, therefore, is appropriate when there is delay between the movement of the control valve and its impact on the controlled variable. When the time derivative action is large, there is instability in the process. It is usually little used because of the sensitivity to noise that manifests and the complications that entails. [11]

In order to prove this fact, some graphics have been made. A first order system has been taken; see Figure 8.
In this case, $H(s)$ is equal to one. $G_{pi}(s)$ and $G_p(s)$ are defined as:

$$G_{pi}(s) = \frac{K_p + K_i}{s}$$  \hspace{1cm} (4.26)$$

$$G_p(s) = \frac{1}{Ls + R}$$  \hspace{1cm} (4.27)$$

The global function $G_{CL}(s)$ is defined as:

$$G_{CL}(s) = \frac{G_{pi}(s) \cdot G_p(s)}{1 + G_{pi}(s) \cdot G_p(s)} = \frac{\frac{K_p + K_i}{s} \cdot \frac{1}{Ls + R}}{1 + K_p + K_i} = \cdots = \frac{K_p + K_i}{Ls^2 + (K_p + R)s + K_i}$$  \hspace{1cm} (4.28)$$

Now, if $G_1(s) = \frac{K_p}{Ls^2 + (K_p + R)s + K_i}$ and $G_2(s) = \frac{K_i}{Ls^2 + (K_p + R)s + K_i}$, it is easier to see if the derivative part affects the response, being $G_1(s)$ the derivative part of the function.
Figure 9: Block diagram of the three functions $G_1(s)$, $G_2(s)$ and $G_{CL}(s)$ used to see the derivative influence.

To do so, three different lines will be graphited, one for $G_1(s)$, another one for $G_2(s)$ and the sum of both functions, $G_{CL}(s)$, as it is shown in Figures 10 and 11. It has been done for two cases, an external loop and an internal loop. The results are the same.

Figure 10: Effect of the derivative term in an external loop.
Study of a PMSM

Figure 11: Effect of the derivative term in an internal loop

It can be seen that $G_1(s)$ (blue line) is not behaving in a good way and this fact affects the global function.

Tuning the controllers

Once having that clear, it is time to tune the PI controllers (setting the proportional and integrative constants, while neglecting the derivative part).

The objective of the controllers, as it can be deduced, is to control the model, but with an important requirement: not changing it.

That will be explained with an example. Imagine that we have this system below (see Figure 12) with two PI controllers with their respective functions $G_a$ and $G_b$ (both are closed-looped systems). When both PI controllers are together (as in this case), a requirement is that the internal loop has to behave as it was equal to 1, in order not to disturb the other PI controller.
The conditions to satisfy this requirement are:

1) The dynamic performance of the external loop must be slower than the internal loop (of around 1/100). Therefore, the respectively settling times could be: 100 ms for the external loop and 1 ms for the internal loop.

2) The internal loop bandwidth must be higher than the external loop bandwidth. To do so, a Bode Diagram with these specifications has been graphed.

As it can be seen, the difference is significant.

On the electrical part, $K_p^{id}$, $K_p^{iq}$, $K_i^{id}$ and $K_i^{iq}$ (direct and quadrature) will have to be fixed, while on the mechanical part, $K_{p_{wr}}$ and $K_{i_{wr}}$ will be fixed.
See Figure 25 and 27 on Appendix C.

The proportional and integral coefficients of the electrical controller will be calculated as

\[ K_{pi}^{id} = \alpha_i \cdot L_d \]  
\[ K_{i}^{id} = \alpha_i \cdot R_s \]  
\[ K_{pi}^{iq} = \alpha_i \cdot L_q \]  
\[ K_{i}^{iq} = \alpha_i \cdot R_s \]

where \( \alpha_i \) is the electrical controller bandwidth.

\[ \alpha_i = \ln(9) / t_r^i \]

While the coefficients of the mechanical part are

\[ K_{p}^{wr} = \alpha_{wr} \cdot J_m \]
\[ K_{i}^{wr} = \alpha_{wr} \cdot B_m \]

where \( \alpha_{wr} \) is the mechanical controller bandwidth.

\[ \alpha_{wr} = \ln(9) / t_r^w \]

where \( t_r \) is the settling time (time required for the process variable to settle from 10% to 90% of the final stabilized value). [12]

\( K_{pi}^{id} \), \( K_{pi}^{iq} \), \( K_{i}^{id} \) and \( K_{i}^{iq} \) impose the dynamics in a first order system on the electrical part.

Often times, there is a disturbance in the system that affects the process variable or the measurement of the process variable. It is important to design a control system that performs satisfactorily during worst case conditions. [16]

In general, increasing the proportional gain will increase the speed of the control system response. However, if the proportional gain is too large, the process variable will begin to oscillate. If \( K_p \) is increased further, the oscillations will become larger and the system will become unstable and may even oscillate out of control. [7]

As the proportional gain is increased, the system becomes faster, but care must be taken not to make the system unstable. Once \( K_p \) has been set to obtain a desired fast response, the
integral term is increased to stop the oscillations. The integral term reduces the steady state error, but increases overshoot. Some amount of overshoot is always necessary for a fast system so that it could respond to changes immediately. [2]

It is important to have a relative low settling time. The settling time has to be as low as possible but with less ripples as possible when stabilised. Thus, with the settling times above thought, it was tested and it was proved that it meets the aforementioned aim. So it can be confirmed that the values of the settling times are

\[ t_r^i = 0.001 \text{ s} \]
\[ t_r^{wr} = 0.1 \text{ s} \]

Consequently, the proportional and integral gains are fixed as:

\[ K_p^{id} = 9.1383 \]
\[ K_p^{iq} = 12.5374 \]
\[ K_i^{id} = K_i^{iq} = 725.0841 \]
\[ K_p^{wr} = 0.2213 \]
\[ K_i^{wr} = 0.0449 \]

The equations of the converter (in this case, as it is the control, it is a converter from mechanics to electrics) are the same as in the open-loop system but inverse.

See Appendix C to see the inside of each block.

Now that the system is controlled, the graphic of the new speed response is shown in Figure 14.
Comparing this figure with Figure 5 (open-loop system), it can be seen that with the control, the speed gets stabilized with the same settling time as an open-loop system although ripples appear during the settling time. Nevertheless, the aim of achieving the desired speed of 550 rpm is fulfilled. Thus, simulation results show that the speed controller has a good dynamic response.

Figures 15, 16, and 17 show the behaviour of the motor, with a load torque of 12 Nm and a desired speed of 550 rpm.
Figure 15: Final torque after the control implemented (12 Nm)

Figure 16: Direct current and quadrature current after the control implemented

The fast control loop executes these two independent current-control loops: the direct-axis current ($i_d$) is used to control the rotor magnetizing flux, while the quadrature-axis current ($i_q$) corresponds to the motor torque.
Figure 17: Direct voltage and quadrature voltage with the control

The values when the system is stabilized are the same as in the open-loop system, so it can be affirmed that the control is good implemented.
5. Sensorless control

5.1. Introduction

The use of sensorless control for PMSM is widely used in industrial and offshore applications. By removing the traditional speed sensor, the entire complexity and costs of the motor drive is reduced. A traditional sensor obtains disturbances in the signals at different load conditions for the motor drive. Additionally, the sensor is exposed to the environment. [1]

In order to eliminate the rotor position sensor in the rotor permanent-magnet flux oriented controlled drive system, is of great interest to use a rotor position and velocity estimation technique.

The method that will be used is called flux linkage method. The method estimates the flux-linkage, based on stator currents and voltage signals from the previous control of the motor. A prediction-correction approach is used. [3]

The rotor position and velocity estimation algorithm require the estimation of the stator flux linkage. Using the estimated stator flux, the stator currents are estimated at a predicted rotor position. The difference between the estimated stator currents and the measured currents are used to correct the error in the predicted rotor position. The block diagram of the algorithm is shown in Figure 18. The number in each block of Figure 18 indicates the execution order of the estimation algorithm. [3]

Each step of the algorithm will be discussed.

Thus, the aim of this method is not to measure the speed/position directly, but to employ some indirect techniques to estimate the rotor position instead.
5.2. Flux-linkage method

Figure 18: Block diagram of the rotor position and velocity estimation algorithm
Step 1: Stator flux linkage estimation [1]

The first step of the algorithm is to estimate the flux-linkage integrating the difference between the stator voltage and the ohmic voltage drop. As it is in discrete time, a sampling time $T_s$ is used for the integration. Thus, the stator flux-linkages are obtained as

$$
\lambda^\text{est}_{qs}(k) = T_s \cdot [v^\text{*}_{qs}(k - 1) - r_s \cdot i_{qs}(k)] + \lambda_{qs}(k - 1)
$$

(5.1)

$$
\lambda^\text{est}_{ds}(k) = T_s \cdot [v^\text{*}_{ds}(k - 1) - r_s \cdot i_{ds}(k)] + \lambda_{ds}(k - 1)
$$

(5.2)

The phase voltages are not measured and the controller commanded voltages to the machine in the previous sampling period (note that $v^\text{*}_{qs}(k - 1)$ and $v^\text{*}_{ds}(k - 1)$ are used).

The last term of the equations is the updated value of the estimated flux-linkage (see Step 4). So, those updated values ($\lambda_{qs}(k - 1)$ and $\lambda_{ds}(k - 1)$) in the previous sampling period are used to calculate the stator flux values in the present sampling period.

Step 2: Stator current estimation [1]

Stator currents can be estimated by analysing the change in rotor angle and the updated flux-linkages. As the inductance varies with frequency, an estimated value of the stator currents can be calculated.

$$
i^\text{est}_{qs}(k) = \frac{[L - \Delta L \cos(2\theta^\text{est}_{rp}(k))] \cdot \lambda^\text{est}_{qs}(k) + \Delta L \sin(2\theta^\text{est}_{rp}(k)) \cdot \lambda^\text{est}_{ds}(k) - (L + \Delta L) \cdot \lambda_m \sin(\theta^\text{est}_{rp}(k))}{L^2 - \Delta L^2}
$$

(5.3)

$$
i^\text{est}_{ds}(k) = \frac{[L + \Delta L \cos(2\theta^\text{est}_{rp}(k))] \cdot \lambda^\text{est}_{ds}(k) + \Delta L \sin(2\theta^\text{est}_{rp}(k)) \cdot \lambda^\text{est}_{qs}(k) - (L + \Delta L) \cdot \lambda_m \cos(\theta^\text{est}_{rp}(k))}{L^2 - \Delta L^2}
$$

(5.4)

Where $L = \frac{l_q + l_d}{2}$ and $\Delta L = \frac{l_q - l_d}{2}$.

It is noted that the estimated stator currents are dependent on correct estimation of flux-linkages and inductance.
**Step 3: Position correction** [1]

The most important part of the algorithm is the correction of the predicted rotor position. By comparing estimated and measured (actual) stator currents, the rotor position can be calculated based on the currents errors.

\[
\Delta i_{qs}^s(k) = i_{qs}(k) - i_{qs}^{est}(k) \quad (5.5)
\]

\[
\Delta i_{ds}^s(k) = i_{ds}(k) - i_{ds}^{est}(k) \quad (5.6)
\]

The current errors \( \Delta i_q \) and \( \Delta i_d \) can be obtained by transforming the stationary frame current errors \( \Delta i_{qs}^s \) and \( \Delta i_{ds}^s \) to the predicted rotor reference frame as

\[
\Delta i_q = \Delta i_{qs}^s \cdot \cos(\theta_{rp}^{est}(k)) - \Delta i_{ds}^s \cdot \sin(\theta_{rp}^{est}(k)) \quad (5.7)
\]

\[
\Delta i_d = \Delta i_{qs}^s \cdot \sin(\theta_{rp}^{est}(k)) + \Delta i_{ds}^s \cdot \cos(\theta_{rp}^{est}(k)) \quad (5.8)
\]

And \( i_{qs}^{pre} \) and \( i_{ds}^{pre} \) as

\[
i_{qs}^{pre} = i_{qs}^s \cdot \cos(\theta_{rp}^{est}(k)) - i_{ds}^s \cdot \sin(\theta_{rp}^{est}(k)) \quad (5.9)
\]

\[
i_{ds}^{pre} = i_{qs}^s \cdot \sin(\theta_{rp}^{est}(k)) + i_{ds}^s \cdot \cos(\theta_{rp}^{est}(k)) \quad (5.10)
\]

Two position errors \( \Delta \theta_q \) and \( \Delta \theta_d \) exist and the position correction term \( \Delta \theta \) can be obtained by taking the average of the two position errors:

\[
\Delta \theta_q(k) = \frac{L_q \cdot \Delta i_q}{-\lambda_m + 2\Delta L \cdot i_{ds}^{pre}} \quad (5.11)
\]

\[
\Delta \theta_d(k) = \frac{L_d \cdot \Delta i_d}{2\Delta L \cdot i_{qs}^{pre}} \quad (5.12)
\]

\[
\Delta \theta(k) = \frac{\Delta \theta_q(k) + \Delta \theta_d(k)}{2} \quad (5.13)
\]

The correct rotor position is obtained as

\[
\Delta \theta_{rp}^{est}(k) = \Delta \theta_{rp}^{est}(k) + \Delta \theta(k) \quad (5.14)
\]
Step 4: Updating of flux linkages [1]

In Step 4, the flux is recalculated using the correct rotor position and the measured stator currents.

\[
\lambda_{qs}(k) = [L + \Delta L \cos(2\theta_r^{est})] \cdot i_{qs}(k) - \Delta L \sin(2\theta_r^{est}) \cdot i_{ds}(k) + \lambda_m \sin(\theta_r^{est})
\]

(5.15)

\[
\lambda_{ds}(k) = -\Delta L \sin(2\theta_r^{est}) \cdot i_{qs}(k) + \Delta L \cos(2\theta_r^{est}) \cdot i_{ds}(k) + \lambda_m \cos(\theta_r^{est})
\]

(5.16)

These updated flux values are used in step 1 of the algorithm in the next sampling interval to estimate the flux. In this way, the integrator drift problems can be avoided in the flux estimation in step 1.

Step 5: Prediction of rotor position [1]

The position is predicted assuming the position varies with time as a second-order polynomial

\[
\theta_r = At^2 + Bt + C
\]

(5.17)

Using three estimated previous positions, the position at (k+1) sampling instant is predicted using second-order polynomial curve fitting.

Figure 19: Rotor position prediction using polynomial curve fitting
Assuming $t = 0$ at $(k - 2)$ sampling instant, the rotor position can be obtained as

$$\theta_{\text{est}}(k - 2) = C \quad (5.18)$$

At $(k - 1)$ sampling instant

$$\theta_{\text{est}}(k - 1) = AT^2 + BT + C \quad (5.19)$$

At $(k)$ sampling instant

$$\theta_{\text{est}}(k) = A(2T)^2 + B(2T) + C \quad (5.20)$$

At $(k + 1)$ sampling instant the prediction position is

$$\theta_{\text{est}}(k + 1) = A(3T)^2 + B(3T) + C \quad (5.21)$$

Solving these equations for $A$, $B$ and $C$ and substituting on the predicted position equation at $(k + 1)$ sampling time, the equation is obtained as

$$\theta_{\text{est}}(k + 1) = 3\theta_{\text{est}}(k) - 3\theta_{\text{est}}(k - 1) + \theta_{\text{est}}(k - 2) \quad (5.22)$$

In step 5 of the algorithm, using three previously estimated positions, the position in next sampling instant is predicted using the final equation above.

### 5.3. Problems encountered

This algorithm was implemented in MATLAB/Simulink (see Appendix C to see the whole implementation with the program).

The difficulty appears when calculating $\Delta \theta_d$ in step 3. For small values of $i_{qs}^{\text{pred}}$, $\Delta \theta_d$ can become very large leading to wrong position correction, and therefore $\Delta \theta$. Even though at high loads the problem does not appear, when the machine is operated under low loads the position correction becomes inaccurate leading to fail the position estimation. [$1$]

In order to avoid wrong position correction from very large values of $\Delta \theta_d$, limits to the $\Delta \theta_d$ can be added. The rotor position variation during one sampling period depends on the speed of the rotor and is is given by $\omega_r T$, where $T$ is the sampling period. Therefore, the maximum value of $\Delta \theta$ can be limited to $\pm \omega_r T$. However, at no-load the rotor position does not still track the actual rotor position accurately. [$1$]

For non-salient pole machines (SPMSM) the denominator of $\Delta \theta$ does not include any time varying variable. Therefore, position correction of non-salient pole machines does not face
difficulties as it is seen for salient pole machines. [1]

Because of this problem, the position estimation algorithm was investigated neglecting the saliency in the machine. This entails assuming $\Delta L = \frac{L_q - L_d}{2} \approx 0$ and the position error appears only in q-axis equation. This is different from previously discussed salient pole machine (IPMSM) case, where the position error appears in both q and d equations. [1]

The new equation for the position error is

$$\Delta \theta(k) = \frac{-L_q \cdot \Delta i_q}{\lambda_m}$$

(5.23)

The reason for the position error when the load is increased is due to the assumptions made in the position estimation algorithm. These assumptions are neglecting the saliency when the currents are estimated in step 2 of the algorithm and when updating the flux linkages in step 4 of the algorithm. This makes errors in the current estimation and the updating of flux linkages. The combination of these facts makes increased flux errors, increased current errors and inaccurate position correction when the load is increased in the machine. [1]

Thus, the position error is dependent on the load of the machine and when the load is increased (bigger than the rated load) the position error is also increased.

Another interesting fact is the effect of the speed. When the speed decreases, the estimation is worse.

The reference signal is considered as a position measurement from an ideal sensor.

In order to reduce the oscillations, additional filtering or different tuning methods for the controller gains can be applied. This is highly dependent on the criteria given for the operation of the motor drive.

One option to avoid noise is by adding low pass filters with a natural frequency of 50 Hz and a damping ratio of 0.707. This will influence the flux linkage estimation in Step 1, and the error will remain until the sampled current loops. [3]

The main problem with sensorless control lies in the low speed operations. The position algorithm is based on motor parameters and is sensitive to changes in the signal processing.
Conclusions

PMSM are good candidates due to their attractive efficiency characteristics.

Firstly, the whole drive system is simulated by the use of MATLAB/Simulink. With the motor equations, a model for the machine has been developed. The control system requires rotor position feedback in order to perform the self-synchronization function continuously. The basic PI controllers are sufficient for current and speed control in the drive system. The results of the simulation show the good response of the model when tracking a command speed. The design of the controllers has been validated with MATLAB/Simulink.

For an IPMSM, the rotor d,q model is the most convenient, since the position dependent inductances disappear in that model. The electromagnetic torque of the IPMSM is not only produced by the permanent-magnet flux but also by the reluctance difference in rotor d- and q-axes. This is different from SPMSMs where the electromagnetic torque is only produced by the permanent flux.

In order to achieve sensorless operation of the drive, the position and velocity estimation is required and it is important to consider the type of the machine (SPMSM or IPMSM) and the application of the drive (to know the convenient value of the speed and torque) when investigating a rotor position and velocity estimation technique. The saliency in IPMSMs increases the complexity in the algorithm compared to the SPMSMs.

For position sensorless operation of the system a rotor position estimation technique has been investigated. The investigated rotor position estimation technique uses predictor-corrector method, which uses current errors to correct the predicted rotor position. It has shown that, in theory, the expressions exist for both SPMSM and IPMSM to correct the predicted rotor position using current errors. Nevertheless, the correction of the predicted rotor position is difficult for salient pole machines (IPMSM) using the predicted d-q reference frame current errors. For non-salient pole machines (SPMSM), position correction using current errors in the predicted d-q reference frame seems more convenient.

The difficult of position correction in the position estimation algorithm for all operating conditions and the relatively low saliency that exists in the IPMSM used for the analysis has led to make assumptions to the position estimation algorithm.

In order to improve the performance of the drive, more investigations are still required for accurate rotor position estimation of the IPMSM.
Future work

There are still some topics that would be interesting to study in future projects:

- Variate the machine’s parameters and see how much effect give to the performance of the controllers.
- Simulate the sensorless control algorithm and calculate the difference between the real position and the estimated position.
- Analyse the performance of the sensorless system with no load, 50% load and 100% load.
- Analyse the performance of the sensorless system for different speed profiles as well as different speed values (for example: 5%, 50% and 100% of rated speed; note that 0% of rated speed would give inadequate values for the current and voltage).
- Implement the sensorless control algorithm with the rotor position estimation using the saliency in the IPMSM. This may improve the estimation, so an updated algorithm should be thought.
Thanks

In first place, I would like to express my gratitude to my co-tutor Daniel Heredero, the person who taught me everything from the basics that I needed to know in order to carry out this project. His guidance, help and technical support throughout these months have been essential.

I would like to sincerely thank my tutor Daniel Montesinos for the help and feedback he has given me during this project. Also, while I was doing this project he also taught me the subject “Electric Mobility” which was very useful for the project.

I would also like to thank CITCEA-UPC for the opportunity they gave me to be part of their team during these months. I really appreciate all the help that I have received; it has been the reason I have improved in this field of knowledge.
Bibliography


Appendices

A) Data for the IPMSM

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<tr>
<th>Parameter</th>
<th>Value</th>
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<tr>
<td>Type</td>
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<tr>
<td>Number of poles (n)</td>
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</tr>
<tr>
<td>Rated power</td>
<td>2.2 kW</td>
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<tr>
<td>Rated speed</td>
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<td>Rated frequency</td>
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<td>Rated torque</td>
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<td>q-axis inductance ($L_q$)</td>
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<td>Inertia of the rotating system (J)</td>
<td>10.07·10$^{-3}$ kg m$^2$</td>
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<tr>
<td>Viscous friction coefficient ($\beta_m$)</td>
<td>20.44·10$^{-4}$ Nm s rad$^{-1}$</td>
</tr>
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</table>
B) MATLAB file

```matlab
% PMSM motor nameplate - SSR1-42F2AFNL Yaskawa Electric Corporation
n = 6 ; % number of poles
rs = 3.3 ; % stator resistance per phase [ohm]
Ld = 41.59e-3 ; % d-axis inductance [H]
Lq = 57.06e-3 ; % q-axis inductance [H]
lambdam = 0.4832 ; % rotor permanent-magnet flux [V s rad-1]
L=(Lq+Ld)/2 ;
delta_L=(Lq-Ld)/2 ;

% Mechanical characteristics
Jm = 10.07e-3 ; % Inertia [kg m2]
Bm = 20.44e-4 ; % Friction factor [N m rad s-1]

UDC = sqrt(3)*400; % DC bus voltage [V]
TL = 12; % Load torque [Nm]

% Control
tr_idq = 10e-3;
alpha_idq = log(9)/tr_idq;

Kp_id = alpha_idq*Ld;
Ki_id = alpha_idq*rs;
Kp_iq = alpha_idq*Lq;
Ki_iq = alpha_idq*rs;

tr_wr = 100e-3;
alpha_wr = log(9)/tr_wr;

Kp_wr = alpha_wr*Jm;
Ki_wr = alpha_wr*Bm;

% Sampling time
Ts= 1/5000;
```

C) Model implementation with MATLAB/Simulink

![Battery block](image)

*Figure 20: Battery block*
Figure 21: Power converter

Figure 22: Electric block of the open-loop system

Figure 23: Electromechanical conversion block of the open-loop system
Study of a PMSM

Figure 24: Mechanical block of the open-loop system

Figure 25: Mechanical block of the control

Figure 26: Mechanical-Electrical conversion block of the control

Figure 27: Electric block of the control
Figure 28: Alpha modulator block

Figure 29: Closed-loop system with the control implemented

Figure 30: Flux-linkage algorithm
Study of a PMSM

Figure 31: Step 1 of the algorithm

Figure 32: Step 2 of the algorithm

Figure 33: Step 3 of the algorithm
Figure 34: Step 4 of the algorithm

Figure 35: Step 5 of the algorithm