Abstract

This paper addresses the design, simulation and experimental validation of a Permanent-Magnet Synchronous Motor speed control. The control is based on the passivity strategy; special attention is paid on the stability control issue. SIMNON simulations illustrating the torque and speed response of the motor control are carried out to validate the control strategy. The algorithm control was programmed on a Texas Instruments TMS320F28335 Digital Signal Controller. Experimental results validating the control strategy are illustrated and analyzed.

Keywords: Type your keywords here, separated by semicolons;

1. Introduction

Permanent Magnet Motors are gaining interest on industrial and transport applications due mainly to the fact that these motors own high power densities and efficiencies. Permanent Magnet brushless DC motors are
preferred in those applications where the high ripple torque of these motors is not an inconvenient and the simplicity to control it is an advantage. The Permanent Magnet Synchronous Motor (PMSM) is preferred in applications where high performance control is needed. The cost of the PM motors and Digital Signal Controllers (DSC) are decreasing continuously doing these motors attractive for high performance and efficiency applications [1].

Field Oriented Control (FOC) and Direct Torque Control (DTC) are two control strategies used to control these motors. In the FOC the motor equations are transformed to a coordinate system which is rotating synchronously with the permanent magnet flux. By means of the transformation the flux and torque equations are separated which allows controlling the flux and torque separately by using PI controllers for currents resulting in a control similar to a brushed DC motor control [2]. On the other hand DTC strategy is based on selecting voltage vectors that are applied to the motor stator; the vector selected depends on the error between the reference and measured torques and the reference and estimated fluxes [3]. An advantage of the DTC is that it needs the stator resistance as the only motor parameter.

Passivity concept has gained importance in many control areas [4]. Passivity was used originally in classical mechanics problems and then it was extended to control problems [5]. Passivity-based control with energy molding for induction motors was carried out in [6]. The passivity-based control for PMSM was carried out by Qjo [7]. Achour [8] proposed to use the magnetic fluxes as the state variables instead of the currents for a PMSM. In the work presented in this paper the passivity-based control signal is calculated according to Linares [9]. SIMNON simulations were carried out to validate the passivity model then the algorithm was programmed on a Texas Instruments Digital Signal Controller TMS320F28335 to validate experimentally the motor speed control.

Passive systems are dynamic systems where the energy exchange is a key concept. Two functions are needed to define the passivity: the rate at which the energy flows towards the system and the amount of energy stored in the system. These functions are related by means of the dissipative inequality which states that the trajectories of the dissipative system of the supply is not less that the increment in the storage. That means that it is not possible to store more energy than that supplied by the power supply.

The passivity-based control is implemented on two stages. The first stage is the energy molding in which a new potential energy function with a unique minimum equilibrium point is generated, on the second stage a damping characteristic is added to modify the dissipation function to guarantee asymptotic stability. The resultant control law is a proportional-derivative controller. The control strategy used to control the motor speed uses the feedback of the passive error output. Essentially it is a linear controller for a nonlinear system, which is based on the storage of the total energy of the system to dissipate it then by means of the feedback of the passive output, which makes the system to lose energy resulting in a way to regulate the physical variables of the motor to the reference values [9].

2. Motor Model

The stator winding of a PMSM is similar to that of an induction motor. According to the way the permanent magnets are fitted in the rotor the motors are classified in two types: Interior Permanent Magnet Synchronous Motor (IPMSM) and Surface Permanent Magnet Synchronous Motor (SPMSM). The first type of motors is sturdy and is ideal for high speed applications. The second type of motor has the magnets
mounted on the surface of the rotor resulting in a motor with radial flux and airgap reluctance nondependent of the rotor position.

Similar to an induction motor the PMSM stator have three coils spatially separated by 120° each other, the coils are supplied with a three-phase voltage. The resulting phase currents depend on the self inductance and resistance of each coil and the coupling fluxes among the coils and the permanent magnet flux. The analytic model of a PMSM is complex if it is worked out in the static three-phase reference frame; nonetheless the model is simplified if Clark and Park transformations are applied. The \(a\beta\) (Clark) transformation transforms a three-phase static system into a two-phase static system. The \(dq\) (Park) transformation transforms a three-phase static system into a two-phase rotating system, the two-phase axis rotate at the rotor speed.

The PMSM \(dq\) model is comprised by two differential equations (1) where \(V_d, V_q, i_d\) and \(i_q\) are the voltages and currents on the \(dq\) reference axis, \(L_d\) and \(L_q\) are the inductances, \(R\) the stator resistance, \(\Psi_{PM}\) is the magnitude of the permanent magnet flux and \(\omega_r\) the electrical speed of the rotor [10].

\[
\begin{align*}
V_d &= L_d \frac{di_d}{dt} - \omega \cdot L_q \cdot i_q + R \cdot i_d \\
V_q &= L_q \frac{di_q}{dt} + \left( \omega \cdot L_d \cdot i_d + \Psi_{PM} \right) + R \cdot i_q
\end{align*}
\] (1)

The mechanical response of the motor is modeled by the differential equation (2) where \(\tau_e\) is the electromagnetic torque of the motor, \(T_L\) is the load torque, \(b\) the friction coefficient, \(J\) the rotor inertia constant and \(P\) the pole-pair number. The electromagnetic torque \(\tau_e\) is calculated by using equation (3), for a SPMSM the electromagnetic torque is proportional to the \(i_q\) current.

\[
\tau_e - T_L - b \cdot \omega_e = \frac{J}{P} \frac{d\omega_e}{dt}
\] (2)

\[
\tau_e = \frac{3}{2} P \left[ \Psi_{PM} i_q \right]
\] (3)

The relationship between the mechanical speed \(\omega_e\) in rpm and the electrical speed is \(\omega_{rpm} = (30 \cdot \omega_e) / (P \cdot \pi)\). Substituting equation (3) into (2) the mechanical equation becomes:

\[
\frac{2}{3} \frac{J}{P^2} \frac{d\omega_e}{dt} = \Psi_{PM} i_q - \frac{2}{3P} \left( T_L + b \cdot \omega_e \right)
\] (4)

In order to be able of applying passivity strategy the PMSM model must be converted to a state-space representation where the state vector is \(X=[i_d i_q \omega_e]\). The characteristic state equation of a passive system is given in (5) where \(A\) is a symmetric, positive definite, constant, matrix, \(u=[V_d V_q]^T\) is the input-matrix control, \(J(u)\) is a skew symmetric matrix, \(R\) is the losses matrix and \(\varepsilon\) is the disturbance vector.

\[
A \dot{X} = J(u)X - RX + Bu + \varepsilon
\] (5)
Simple inspection of eqns. (1) and (4) reveal that these are already in state variables representation as shown in eqn. (6). $J(u)$ and $R$ matrices must fulfill the equations $J^T(u) = -J(u)$ and $R^T = R$ to guarantee that the system is controllable by passivity. The PMSM system fulfills these conditions resulting in a system that can be controlled by passivity.

\[
\begin{bmatrix}
L_0 & 0 & 0 \\
0 & L_0 & 0 \\
0 & 0 & \frac{2J}{3P^2}
\end{bmatrix}
\begin{bmatrix}
\frac{di_d}{dt} \\
\frac{di_q}{dt} \\
\frac{d\omega_r}{dt}
\end{bmatrix}
\]

(6)

\[
\begin{bmatrix}
0 & \omega_r L_0 & 0 \\
0 & L_0 & -\Psi_{PM} \\
0 & \Psi_{PM} & 0
\end{bmatrix}
\begin{bmatrix}
i_d \\
i_q \\
\omega_r
\end{bmatrix}
- \begin{bmatrix}
R & 0 & 0 \\
0 & R & 0 \\
0 & 0 & \frac{2b}{3P^2}
\end{bmatrix}
\begin{bmatrix}
i_d \\
i_q \\
\omega_r
\end{bmatrix}
+ \begin{bmatrix}
1 & 0 \\
0 & 1 \\
0 & 0
\end{bmatrix}
\begin{bmatrix}V_d \\ V_q \end{bmatrix}
+ \begin{bmatrix}0 \\ 0 \\ -\frac{2}{3P}\end{bmatrix}T_L
\]

To apply passivity a new system which is a copy of the original system must be defined as shown in eqn. (7) where $X^*$ are the desired state variables and $u^*$ the desired input.

\[
A^*X^* = J(u^*)X^* - RX^* + Bu^* + \varepsilon
\]

(7)

Subtracting eqn. (5) from (6) results in eqn. (8), where $e = X - X^*$ and $e_u = u - u^*$.

\[
A^*e = [J(u) - J(u^*)]X - Re + Be_u + J(u^*)e
\]

(8)

With the aim of linearizing eqn. (8) $J(u)$ is linearized by Taylor series as follows:

\[
J(u) = J(u^*) + \left. \frac{\partial J(u)}{\partial u} \right|_{u^*} e_u
\]

(9)

As $J(u)$ is a constant matrix its partial derivative is zero. By substituting this result in eqn. (8) the resultant equation is:

\[
A^*e = -Re + Be_u + J(u^*)e
\]

(10)

To demonstrate the stability of eqn. (10) the second Lyapunov principle is used. The Lyapunov candidate function proposed is the positive defined function $V(e) = (1/2)e^T Ae$. Taking the derivative with respect to the time we get eqn. (11).
\[ \dot{V}(e) = e^T A e = -e^T Re + e^T Be_u + e^T J(u^*)e \]  
\hspace{2cm} (11) 

If \( e_u = -\gamma B^T e \) and accounting for the fact that \( e^T J(u^*) \) is nonsymmetrical then \( e^T J(u^*)e = 0 \), substituting this result in eqn. (11):

\[ \dot{V}(e) = -e^T [R + B\gamma B^T]e \]  
\hspace{2cm} (12) 

To make \( \dot{V} \) negative defined the inequality \( [R + B\gamma B^T] \geq 0 \) has to be fulfilled which guarantees the controller stability. To find the control law it is assumed that \( e_u = u - u^* = -\gamma B^T e \), whence equation (13) is derived. Finally the control law \( u = u^* - \gamma B^T e \) can be derived as shown in eqn. (14).

\[ e_u = -\begin{pmatrix} \gamma_1 & 0 & 1 & 0 \\ 0 & \gamma_2 & 0 & 1 \\ i_d - i_d^* & i_q - i_q^* & \omega_r - \omega_r^* \\ \end{pmatrix} = -\begin{pmatrix} \gamma_1 (i_d - i_d^*) \\ \gamma_2 (i_q - i_q^*) \\ \end{pmatrix} \]  
\hspace{2cm} (13) 

\[ u = \begin{bmatrix} V_d \\ V_q \end{bmatrix} = \begin{bmatrix} V_d - \gamma_1 (i_d - i_d^*) \\ V_q - \gamma_2 (i_q - i_q^*) \end{bmatrix} \]  
\hspace{2cm} (14) 

The desired values for \( V_d^*, V_q^*, i_d^* \) and \( i_q^* \) can be gotten from the system equilibrium point \( \dot{X} = 0 \) given by eqn. (13) resulting in the following equation.

\[ \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} \omega^* L_d i_d^* - Ri_d^* + V_d^* \\ -\omega^* (L_d i_d^* + \Psi_{PM}) - Ri_d^* + V_q^* \\ \Psi_{PM} i_q^* - \frac{2}{3P} (T_L + b \omega^*) \end{bmatrix} \]  
\hspace{2cm} (15) 

Taking into account the fact that to get the maximum motor torque \( i_d = 0 \) and solving for \( V_d^*, V_q^* \) and \( i_d^* \) we get the desired values. Substituting eqn. (16) into eqn. (14) the control law is gotten.

\[ \begin{align*} 
V_d^* &= -\omega^* L_d i_d^* \\
V_q^* &= \omega^* (L_d i_d^* + \Psi_{PM}) \\
i_q^* &= \frac{2}{3P \Psi_{PM}} \left( T_L + \frac{b \omega^*}{P} \right) 
\end{align*} \]  
\hspace{2cm} (16)
3. Simnon simulation

Simnon simulations were carried out to test the passivity controller. The inputs to the controller are reference speed and load torque, those inputs are used to calculate currents and voltages by using eqn. (16) and the control law is calculated by using eqn. (14). The controller tuning is done varying the gains \( \gamma_1 \) and \( \gamma_2 \) (must be positive) until the best response is achieved. The outputs of the controller are the \( V_d \) y \( V_q \), voltages then the power inverter convert them to a three-phase voltage \( (V_a, V_b, V_c) \). The a and b motor phase currents are measured meanwhile c phase current is calculated assuming that the motor is a balanced system. The controller was programmed in a discrete time file and the plant in a continuous time file. The sampling time used was of 10 \( \mu \)s. The motor parameters are shown in table I.

Simulation results are illustrated in Fig. 1, the reference speed is \( \omega_{ref} = \pm 1000 \text{ rpm} \) and the speed response is \( \omega_{pm} \). The response time is of 60ms only. The steady state speed error is of 10 rpm and 130 rpm which is less than 13%. The controller gains used were \( \gamma_1 = 1 \) and \( \gamma_2 = 30 \). Fig. 1(b) illustrates the electromagnetic torque and the load torque. Even that \( TL = 0 \), \( \tau_e \) is non null because the motor needs to cope with the friction losses which for steady state are of \( \tau_e = 0.4 \text{ Nm} \).

Table I. Motor parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bus voltage ( V_{DC} )</td>
<td>180V</td>
</tr>
<tr>
<td>Pole pair ( P )</td>
<td>4</td>
</tr>
<tr>
<td>Permanent Magnet Flux ( \Psi_{PM} )</td>
<td>0.0256 Wb</td>
</tr>
<tr>
<td>Inertia constant ( J )</td>
<td>0.000031 kg/m²</td>
</tr>
<tr>
<td>Resistance ( R )</td>
<td>2.35Ω</td>
</tr>
<tr>
<td>d inductance ( L_d )</td>
<td>0.0065H</td>
</tr>
<tr>
<td>q inductance ( L_q )</td>
<td>0.0065H</td>
</tr>
<tr>
<td>Viscous friction ( B )</td>
<td>0.0038Ns</td>
</tr>
</tbody>
</table>

![Fig. 1. Simulation results. (a) Speed response (\( \omega_{pm} \)). (b) Electromagnetic torque (\( T_e \))](image-url)
4. Experimental validation

Experimental validation was carried out by using a Power Electronics Texas Instruments system comprised of a 350V, 750W three-phase inverter; sensored-phase current and voltage, quadrature encoder interface, 85-132 V AC rectifier, USB isolated interface, PWM output and over current protection. Fig. 2 illustrates a block diagram of the system [11].

A Texas Instruments TMS320F28335 DSC was used to implement the control strategy. The DSC has a processing speed of 150 MIPs. It is able to implement 32 bit fixed-point operations and IEEE-754 format floating-point operations. By means of a JTAG interface it is capable to connect with a computer in such a way that it can modify the DSC registers without any delay. It has 12 analog inputs and 6 PWM complementary outputs and a dead time generator. It also has 32kWords of RAM and 256kWords of flash memory.

Fig. 3(a) illustrates the experimental motor speed for positive and negative speed reference values. The waveform was captured by the development tool “code composer”, 200 samples/s were captured during 10 s. At the current stage of the project the torque-reference value needed to get the speed value desired is loaded in the program, nor torque measurements neither torque observers have been used yet. The reference speed is ±1000rpm and null load-torque. As illustrated in Fig. 3(a) the experimental steady-state positive and negative speed values are 940rpm and -984rpm being the error less than 6%.

Fig. 3(b) illustrates the experimental waveforms for alpha and beta currents which are sinusoidal of 0.2A and a phase shift of 90° as expected.

\[\text{Fig. 2. (a) Block diagram of the TI system}\]

5. Conclusions

This paper addressed the design, simulation and experimental validation of a passivity-based PMSM control. Stability of the controller was analyzed. Simnon simulations were carried out to validate the controller design. Experimental validation was carried out using a TI systems and the control strategy was programmed on a DSC. Passivity control is a strategy which is easy to program as it is reduced to an algebraic equation. The
The main drawback of the strategy is that it needs a load torque sensor or an observer which has to be programmed in the DSC. The aim for future work is to include varying load-torque which implies to use observers for the load torque besides an electrodynamometer will be built to test the control.

Fig. 3. (a) Experimental speed response; (b) Alfa beta currents

References


