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Passivity Based Control for Permanent-Magnet Synchronous Motors

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1. Introduction

The Passivity based control (PBC) is a well established technique which has proved very powerful to design robust control for physical system, especially electrical machinery. The PBC have clear physical interpretation in terms of interconnection system with its environment, and are robust overlooked non dissipative effects modelled. These features are extremely valuable in practical implementations of controllers. In this chapter, we show how the PBC can be used to control the speed of permanents magnets synchronous motor (PMSM). In first part, we consider the Euler-Lagrange model in the αβ-referential to design the Passivity Based Voltage Controller. The dq-model of the PMSM is considered to design the Passivity Based current Controller in the second part.

The idea of Passivity Based Control (PBC) design is to reshape the natural energy of the system and inject the required damping in such a way that the control objective is achieved. Expected advantages of this approach are the enhanced robustness properties, which stem from the fact that conciliation of system nonlinearities is avoided.

The technique has its roots in classical mechanics (Arnold, 1989) and was introduced in the control theory in the seminal paper (Takegaki & Arimoto, 1981). This method has been instrumented as the solution of several robot manipulator (Ailon & Ortega, 1993; Ortega & Spong; Takegaki & Arimoto, 1981) induction motor (Gökder & Simaan, 1997; Kim et al., 1997; Ortega et al., 1996, 1997; Ortega & Loria), and power electronics (Sira-Ramirez et al., 1995), which were intractable with other stabilization techniques.

PBC was also combined with other techniques (Achour & Mendil, 2007; Ortega & García-Canseco 2004a, 2004b; Qiu & Zhao, 2006; Petrović et al., 2001; Travieso-Torres et al., 2006, 2008). The design of two single-input single-output controllers for induction motors based on adaptive passivity is presented in (Travieso-Torres et al., 2008). Given their nature, the two controllers work together with field orientation block. In ((Travieso-Torres et al., 2006), a cascade passivity-based control scheme for speed tracking purposes is proposed. The scheme is valid for a certain class of nonlinear system even with unstable zero dynamic, and it is also useful for regulation and stabilization purposes. A methodology based on energy shaping and passivation principles has been applied to a PMSM in (Petrović et al., 2001). The interconnection and damping structures of the system were assigned using the Port-Controlled Hamiltonian (PCH) structure. The resulting scheme consists of a steady state feedback to which a nonlinear observer is added to estimate the unknown load torque. The authors in (Qiu & Zhao, 2006) developed a PMSM speed control law based on PCH that achieves stabilization via system passivity. In particular, the PCH interconnection and damping matrices were shaped so that the physical (Hamiltonian) system structure is preserved at the closed-loop level. The difference between the physical energy of the system and the energy supplied by the controller forms the closed-loop energy function. A review of the fundamental theory of the Interconnection and Damping Assignment Passivity Based Control technique (IDA-PBC) can be found in (Ortega & García-Canseco 2004a, 2004b). In the concerned papers it was showed the role played by the three matrices (i.e. interconnection, damping, Kernel of system input) of the PCH model in the IDA-PBC design.

The permanent-magnet synchronous motor (PMSM) has numerous advantages over other machines that are conventionally used for ac servo drives. It has a higher torque to inertia ratio and power density when compared to the Induction Motion or the wound-rotor Synchronous Motor, which makes it preferable for certain high-performance applications like robotics and aerospace actuators. However, it presents a difficult control problem. This is due to the following reasons: first, the dynamical model of PMSM is nonlinear. Second, the motor parameters (e.g., stator resistance) can vary considerably from the nominal values. Also, the state variable (velocity and current) measurements are often contaminated with a considerable amount of noise. Generally, velocity and current sensors are omitted due to the considerable saving in cost, and volume.

In Section 2, we propose a design strategy that utilizes the passivity concept in order to develop a combined controller-observer system for Permanent-Magnet Synchronous Motors (PMSM) speed control using only rotor position measurement and voltages applied to the stator windings. To this end, first a desired energy function for the closed loop system is introduced, and then a combined controller-observer system is constructed such that the closed loop system matches this energy function. A damping term is included to ensure asymptotic stability of the closed loop system. The interesting feature of this approach is the fact that it establishes a duality concept between the controller and observer design strategy. Such a duality feature is unique for nonlinear systems. Simulation tests on the combined controller-observer design are provided to show the feasibility and the performances of this method.

The work of Section 3 is related with previous work concerning the voltage control of PMSM (Achour & Mendil, 2007). The PBC has been combined with a variable structure compensator (VSC) in order to deal with important parameter uncertainties plant, without raising the damping values of the controller. The dynamics of the PMSM were represented as feedback interconnection of a passive electrical and mechanical subsystem. The PBC is applied only to the electrical subsystem while the mechanical subsystem has been treated as a passive perturbation. A new passivity based current controller (PBCC) designed using the dq-model of PMSM is proposed in this Section 3.

2. Passivity based controller-observer design for permanent magnet synchronous motors

In this part, we develop a control algorithm based on the passivity concept that forces the PMSM to track desired velocity and torque vectors without the need for velocity and stator current measurements, but using only rotor position and stator voltage measurements.

The passivity-based controller design proceeds as follows. First, we carry out a decomposition of the system dynamics as a feedback interconnection of passive subsystems,

where the outputs of the forward subsystem are the regulated outputs. Second, we design an inner feedback loop that, via the injection of a nonlinear damping term, ensures the controlled subsystem defines a strictly passive map from control signals to regulated outputs. Third, the passivity-based technique is applied to this subsystem leaving the feedback subsystem as a "passive perturbation". This last step involves the definition of the desired closed loop energy function whose associated "target" dynamics evolves on a subspace of the state space ensuring zero error tracking.
The main contribution is in the design of an observer that utilizes the high quality position information and voltage for reconstructing the velocity and current signals. The proposed observer is inspired from the passivity based controller design concept. The problem is tackled by constructing an observer that forces the estimated error to match a desired energy function, thereby preserving the passivity property. In addition, for asymptotic stabilization, damping has to be included in the loop. The main feature of this approach is in the fact that it establishes a concept duality between the controller and observer design strategy. Using passivity concept solves stability of the combined controller-observer design. We will introduce a desired energy function that consists of two parts, one for the closed loop controller dynamic and the other for the closed loop observer dynamic.
The organization of this Section is as follows: In Subsection. 1.2 we present the two phases αβ model of PMSM described by Euler-Lagrange (EL) equations, and his properties. The design procedure and the stability problem of the combined controller-observer are given in Subsection. 1.3. Simulation results are presented in Subsection. 1.4. Finally, concluding remarks are given in Subsection. 1.5.

2.1 Permanent-magnet synchronous motor model

2.1.1 Model

The PMSM uses surface mounted rare earth magnets. We consider the following assumptions: -No significant saliency effects; -negligible damping effects in the rotor; -negligible saturation effects; -ideal symmetrical phases and sinusoidal distributed phase windings; -negligible capacity effects in stator windings, considering rigid shaft and not magnetic material in stator. Under the assumptions above, the standard two phases αβ model of PMSM obtained in (Ortega et al., 1997) via direct application of EL equation is given by:

$$D_e\ddot{q}_e + W_2(q_m)\dot{q}_m + R_e\dot{q}_e = U \tag{1}$$

$$D_m\ddot{q}_m + R_m\dot{q}_m = \tau(\dot{q}_e, q_m) - \tau_L \tag{2}$$

$$\tau(\dot{q}_e, q_m) = W_2^T(q_m)\dot{q}_e \tag{3}$$

where

$$D_e = diag\left\{L_d, L_q\right\}; R_e = diag\left\{R_a, R_a\right\}$$
$$W_2(q_m) = \frac{d\,\mu(q_m)}{d\,q_m}.$$

$\dot{q}_e = \left[\dot{q}_{e\alpha}, \dot{q}_{e\beta}\right]^T \in \Re^2$ is stator current vector; $(q_m, \dot{q}_m) \in \Re^2$ are the rotor angular position and velocity respectively; $\mu(q_m)$ is the flux linkages due to permanent magnets; L_d, L_q are the direct and quadrate stator inductance respectively; D_m is the rotor inertia; $R_m \geq 0$ is the mechanical friction; $U = \left[u_\alpha, u_\beta\right]^T$ is stator voltage vector; and τ, τ_L are the generated and load torque respectively. The subscripts $(.)_e$, $(.)_m$, $(.)^T$ denotes the electrical, mechanical and vector transposition respectively.

2.1.2 Properties

In this subsection, we present three properties of the PMSM model, which are useful for the methodology of control design.

2.1.2.1 Passivity property of permanent-magnet synchronous motor

Lemma 1

The PMSM represents a passive system, if $\nu = \left[U^T, -\tau_L\right]^T$ and $\dot{q} = \left[\dot{q}_e^T, \dot{q}_m\right]^T$ are considered as inputs and outputs respectively.

Proof

The total energy H of the PMSM is:

$$H(\dot{q}_e, \dot{q}_m, q_m) = \frac{1}{2}\dot{q}_e^T D_e \dot{q}_e + \mu^T(q_m)\dot{q}_e + \frac{1}{2}D_m \dot{q}_m^2 \tag{4}$$

Taking the time derivative of H along the trajectory (1)-(3), we get:

$$\dot{H}(\dot{q}_e, \dot{q}_m, q_m) = -\dot{q}^T R \dot{q} + \dot{q}^T \nu + \frac{d}{dt}\left(\mu^T(q_m)\dot{q}_e\right) \tag{5}$$

Integrating $\dot{H}$ from zero to $\gamma > 0$, and setting $\beta = -\left(H(0) + \left[\mu^T(q_m)\dot{q}_e\right]_0^\gamma\right)$, proves the passivity of the PMSM.

2.1.2.2 Passive Feedback Decomposition

Lemma 2

The PMSM can be represented as the negative feedback interconnection of the electrical and mechanical passive subsystems.

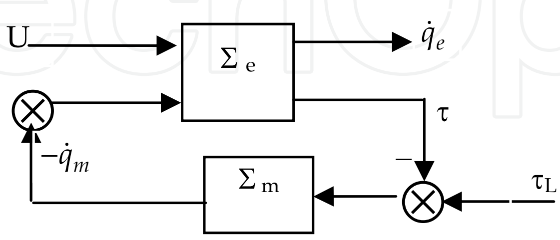


Fig. 1. Passive subsystem decomposition.

$$\Sigma_e: \quad L_{2e}^3 \to L_{2e}^3$$
$$\begin{bmatrix} U \\ -\dot{q}_m \end{bmatrix} \mapsto \begin{bmatrix} \dot{q}_e \\ \tau \end{bmatrix}$$

$$\Sigma_m : \quad L_{2e} \to L_{2e}$$
$$(\tau - \tau_L) \mapsto \dot{q}_m$$

where L^3_{2e} , L_{2e} are the spaces of 3 and 1 dimension respectively of square integral, essentially bounded functions and their extensions.

Proof

Considering the total energy H_e of the electric subsystem Σ_e, that is:

$$H_e(\dot{q}_e, q_m) = \frac{1}{2}\dot{q}_e^T D_e \dot{q}_e + \mu^T(q_m)\dot{q}_e \tag{6}$$

A similar procedure used above to prove the passivity of PMSM can be used to establish the passivity of Σ_e, and for mechanical Σ_m we consider the energy function $H_m(\dot{q}_m) = \frac{1}{2}D_m\, \dot{q}_m^2$ to prove the passivity property.

2.1.2.3 Workless forces

In order to introduce the third property, we note that the model (1)-(3) can be written under the following compact form:

$$D\,\ddot{q} + W(q)\dot{q} + R\dot{q} = MU + \xi \tag{7}$$

Where, $D = diag\{D_e, D_m\};\ R = diag\{R_e, R_m\}$

$$M = [I_2, 0_{1\times 2}]^T;\ \dot{q} = [\dot{q}_e^T, \dot{q}_m]^T;\ \xi = [0_{2\times 1}, -\tau_L]^T$$

$$W(q_m) = \left[W_2^T(q_m)\,\dot{q}_m\ , -\dot{q}_e^T W_2(q_m)\right]^T \tag{8}$$

Based on the passivity property of the PMSM and the relations (1)-(3), we deduce that the "workless forces" are given by:

$$C(q_m) = \begin{pmatrix} 0_{2\times 2} & W_2(q_m) \\ -W_2^T(q_m) & 0_{1\times 1} \end{pmatrix} \tag{9}$$

as $C(q_m)$ verifies:

$$C(q_m) = -C^T(q_m) \tag{10}$$

(i.e., $C(q_m)$ is a skew symmetric matrix.)

Remark

In the present of the saliency effects, the "workless forces" are given by:

$$C(q_m, \dot{q}) = \begin{pmatrix} C_{11} & C_{12} \\ C_{21} & C_{22} \end{pmatrix} \tag{11}$$

Where

$$C_{11} = \frac{1}{2}W_1(q_m)\dot{q}_m$$

$$C_{12} = (\frac{1}{2}W_1\ (q_m)\dot{q}_e + W_2\ (q_m))$$

$$C_{21} = -(\frac{1}{2}\dot{q}_e^T W_1\ (q_m) + W_2^T\ (q_m))$$

$$C_{22} = 0$$

as $C\ (q_m\ ,\dot{q})$ verifies: $\dot{D}\ (q)\ = C\ (q_m\ ,\dot{q}) + C^T\ (q_m\ ,\dot{q})$

(i.e., $(\dot{D}\ (q)\ - 2C\ (q_m\ ,\dot{q}))$ is a skew symmetric matrix).

The previous identification of the workless forces permitted us to write the relation (7) under the following form:

$$D\ \ddot{q} + C(q_m)\dot{q} + R\dot{q} = MU + \xi \tag{12}$$

It is with noting that, these properties have been already derived for Induction machine in (Ortega et al., 1996).

2.2 Problem formulation and design procedure

2.2.1 Problem formulation

The control problem can be formulated as follows: Consider the PMSM model (1)-(3) with state vector $\dot{q} = \left[\dot{q}_e^T\ ,q_m\ ,\dot{q}_m\ \right]^T$; inputs $U\in\Re^2$; regulated outputs $\left(\tau\ ,\ \dot{q}_m\right)$; measurable output q_m; immeasurable outputs $(\dot{q}_e^T\ ,\dot{q}_m)^T$. The problem consists of constructing an observer-based controller such that for all smooth desired output function $\tau^*(t)\ \in\ L_\infty$, with known derivative $\dot{\tau}^*(t)\ \in\ L_\infty$, global torque tracking with internal stability is achieved

2.2.2 Design procedure

The steps to follow are mentioned in section 1. We consider the ideal case to simplify the procedure, where all outputs are supposed available from measurement, then we design an observer to reconstruct the states that we not available.

2.2.2.1 Passivity approach to controller design

The desired dynamics must be compatible with the bounded constraints of the PMSM. From equations (1)-(3), we deduce the following desired dynamics:

$$D_e\ddot{q}_e^* + W_2(q_m)\dot{q}_m + R_e\,\dot{q}_e^* = U^* \tag{13}$$

$$D_m\ddot{q}_m^* - W_2^T(q_m)\dot{q}_e^* + R_m\,\dot{q}_m^* = -\tau_L \tag{14}$$

Where $\dot{q}_e^*$, $\dot{q}_m^*$ is the desired current and desired rotor velocity respectively.

The error dynamic are described by:

$$D_e\dot{e}_e + R_e\,e_e = U - U^* \tag{15}$$

$$D_m\ \dot{e}_m - W_2^T(q_m)\ e_e + R_m\ e_m = 0 \tag{16}$$

Where $e_e = \dot{q}_e - \dot{q}_e^*$, $e_m = \dot{q}_m - \dot{q}_m^*$ are the current error and rotor speed error respectively.

The problem is to find a control law U, which ensures $Lim_{t\to\infty}e(t) = 0$, where $e = \left[e_e^T\ ,e_m\ \right]^T$.

To this end, we shape the energy of the closed loop to match a desired energy function, as:

$$H_e^*(e_e) = \frac{1}{2} e_e^T \; D_e \; \mathrm{e_e} \tag{17}$$

Taking the time derivative of H_e^*, along the trajectory (15), we get:

$$\dot{H}_e^*(e_e) = -e_e^T \; (R_e + (\mathrm{U\text{-}U^*}))\; \mathrm{e_e} \tag{18}$$

In order to ensure the convergence of the e_e to zero, we take:

$$U = U^* \tag{19}$$

Since $R_e = R_e^T \rangle 0$, we has

$$\dot{H}_e^*(e_e) = -e_e^T \; R_e \; \mathrm{e_e} \; \le -\lambda_{\min}\{R_e\}\|e_e(t)\|^2 \;, \forall \; \mathrm{t} \tag{20}$$

we conclude that:

$$\|e_e(t)\| \;\le\; \mathrm{m_e} \; \|e_e(0)\| \; e^{-\rho_e \; t} \tag{21}$$

Where,

$$m_e = \sqrt{\frac{\lambda_{\max}\{D_e\}}{\lambda_{\min}\{D_e\}}} \;\rangle\; 0 \;\; , \;\; \rho_{\mathrm{e}} = \frac{\lambda_{\min}\{R_e\}}{\lambda_{\max}\{D_e\}} \;\;\rangle\; 0 \; .$$

$\lambda_{\min}\{\cdot\}$, $\lambda_{\max}\{\cdot\}$ are the minimum and maximum eigenvalues respectively.

Hence the desired current $\dot{q}_e^*$ is asymptotically attainable. We have the following result:

Proposition 1

Let,

$$U = U^* - K_1 \; e_e \tag{22}$$

where $K_1 = k_e I_2$, $\mathrm{k_e} \rangle 0$, I_2 identity matrix 2x2.

Then the convergence to the desired state trajectory is faster.

Proof

Considering the quadratic function (17), and using the same procedure, we get:

$$\|e_e(t)\| \;\le\; \mathrm{m_e} \; \|e_e(0)\| \; e^{-\rho_{e1} \; t} \tag{23}$$

Where,

$$\rho_{\mathrm{e1}} = \frac{\lambda_{\min}\{R_e + K_1\}}{\lambda_{\max}\{D_e\}} \;\;\rangle\; 0 \tag{24}$$

The control law is:

$$U = D_e \ddot{q}_e^* + W_2(q_m)\dot{q}_m + R_e \dot{q}_e^* - K_1 e_e \tag{25}$$

Remarks

1. Since, we can not control the magnetic fields from the permanent magnets; it is reasonable to expect that we must eliminate the effect on electric subsystem Σ_e of the flux linkages due

to the permanent magnets. Which is seen from (25), the term from the permanent magnets must be concealed out a drawback of the scheme. However, this term is a vector in a measurable quantity (position).
2. In the closed loop system, the positive definite matrix K_1 increases the convergence of the tracking error and overcome the imprecise knowledge of system parameters, if we choose high gain k_e.

2.2.2.2 Desired current and desired torque

The PMSM operating under maximum torque if the direct current i_d in the general reference frame d-q (direct-quadrate) equals to zero.
Under the above condition, the desired current in αβ reference frame is chosen as:

$$\dot{q}_e^* = \frac{2\ \tau^*}{3\ \mathrm{n_p}\lambda_m}\begin{pmatrix} -\sin\ (q_m) \\ \cos\ (q_m) \end{pmatrix} \tag{26}$$

where $\tau^\bullet$ is the desired torque; n_p is the number of pole pairs, and λ_m is the amplitude of the flux linkage established by the permanent magnet.
The desired torque is deduced from the desired mechanical dynamic (14), we have:

$$\tau^* = D_m\ddot{q}_m^* + R_m\,\dot{q}_m^* + \tau_L \tag{27}$$

It has been proved in (Kim et al., 1997), that this scheme has two drawbacks, it is an open loop scheme (in the speed tracking error), and its convergence rate is limited by the mechanical constant time (D_m / R_m). In (14) $\tau^\bullet$ is defined as:

$$\tau^* = D_m\ddot{q}_m^* - z + \tau_L \tag{28}$$

$$\dot{z} = -a\ z + b\ \mathrm{e_m}, \quad \text{and } a, b \rangle\ 0\,. \tag{29}$$

With this choice, the convergence rate of the speed error $\omega_m - \omega_m^*$ does not depend only on the natural mechanical damping. This rate can be adjusted by means of the positives gains *b* and *a* have the same role of proportional-derivative (PD) control law.

Remark

If, v and $\dot{q}_e$ are considered as input and output, then it is easy to prove the strict passivity of the closed loop system.

$$v = D_e\ddot{q}_e^* + R_e\,\dot{q}_e^* \tag{30}$$

2.2.2.3 A passivity Approach to observer design

The problem is to construct an auxiliary dynamic system that asymptotically reconstructs the current and velocity signals from input-output measurements, i.e., stator voltage U and rotor position q_m, respectively. To this end we will use a passivity approach. An interesting feature of this approach is that it establishes a conceptual duality, between the strategies of PMSM controller and observer design. Such a duality feature is rather unique for nonlinear systems.
Based on the physical structure of the PMSM model (1)-(3) and the controller structure (25), we introduce the current and velocity observer systems as follows:

$$D_e\ddot{\tilde{q}}_e + W_2(q_m)\dot{\hat{q}}_m + R_e\dot{\hat{q}}_e = U - L_e\dot{\tilde{q}}_e \tag{31}$$

$$D_m\ddot{\tilde{q}}_m - W_2^T(q_m)\dot{\hat{q}}_e + R_m\dot{\hat{q}}_m = -\tau_L - L_v\dot{\tilde{q}}_m \tag{32}$$

where $\dot{\hat{q}} = \left[\dot{\hat{q}}_e^T \;\; ,\dot{\hat{q}}_m \;\; \right]^T$ is the observer state; $\dot{\hat{q}}_e$, $\dot{\hat{q}}_m$ represents the estimated current and estimated velocity respectively; $\dot{\tilde{q}}_e = \dot{\hat{q}}_e - \dot{q}_e$, $\dot{\tilde{q}}_m = \dot{\hat{q}}_m - \dot{q}_m$ are the estimated current error and estimated velocity error; where:

$$L_e = L_e^T \;\rangle\, 0 \,,\;\; L_v \,\rangle\, 0 \tag{33}$$

The model (31), (32) can be written under the following form:

$$D\,\ddot{\hat{q}} + C(q_m)\dot{\hat{q}} + R\dot{\hat{q}} = MU + \xi - L\dot{\tilde{q}} \tag{34}$$

Where $\dot{\tilde{q}} = \left[\dot{\tilde{q}}_e^T \;\; ,\dot{\tilde{q}}_m \;\; \right]^T$ and $\mathrm{L} = diag\{\mathrm{L_e} \;\;, \mathrm{L_v}\}$

From the equation (12) and (34), we deduce the observer error dynamic:

$$D\,\ddot{\tilde{q}} + C(q_m)\dot{\tilde{q}} + (R+L)\dot{\tilde{q}} = 0_{3\times 1} \tag{35}$$

In order to prove the asymptotic stability of the observer estimated error; we choose the following desired energy error function:

$$H_o^*(\dot{\tilde{\mathrm{q}}}) = \frac{1}{2}\dot{\tilde{q}}^T \; D\,\dot{\tilde{\mathrm{q}}} \tag{36}$$

Taking the time derivative of H_o^*, along the trajectory (35), we get:

$$\dot{H}_o^*(\dot{\tilde{\mathrm{q}}}) = -\dot{\tilde{q}}^T \;(R+L)\;\dot{\tilde{\mathrm{q}}} \tag{37}$$

Since $L = L^T \;\rangle\, 0$, $\dot{\tilde{\mathrm{q}}} = 0$ is asymptotically stable.

Following the same procedure used in section II.2.1, we conclude that:

$$\left\|\dot{\tilde{q}}(t)\right\| \;\le\; \mathrm{m_o}\;\left\|\dot{\tilde{q}}(0)\right\|\; e^{-\rho_o\, t} \;,\forall\; \mathrm{t.} \tag{38}$$

where $m_o = \sqrt{\dfrac{\lambda_{\max}\{D\}}{\lambda_{\min}\{D\}}} \;\rangle\; 0 \;\;,\;\; \rho_o = \dfrac{\lambda_{\min}\{R+L\}}{\lambda_{\max}\{D\}} \;\;\rangle\; 0$

We conclude that, the observer (34) reconstructs asymptotically the current and velocity signals.

Remark

We can notice that the gain matrix L has the same effect than that of matrix K_1 in (25), i. e; L is the damping that is injected in the observer system to ensure the asymptotic stability of the observation error.

2.2.2.4 Combined Controller-Observer Design

The desired dynamics, when only rotor position is measurable are:

$$D_e\ddot{q}_e^* + W_2(q_m)\dot{\hat{q}}_m + R_e\dot{q}_e^* = U^* \tag{39}$$

$$D_m\ddot{q}_m^* - W_2^T(q_m)\dot{q}_e^* + R_m\dot{q}_m^* = -\tau_L - k_m e_m \tag{40}$$

Where, km > 0.
We have the following result: The controller law becomes:

$$U = D_e\ddot{q}_e^* + W_2(q_m)\dot{\hat{q}}_m + R_e\dot{q}_e^* - K_2 e_e \tag{41}$$

In order to establish the stability of the closed loop system with presence of the observer, we consider equation of state error (35). We get from (25), (16), (40) and (41):

$$D\,\dot{e} + G(q_m)e + N(q_m)\dot{\tilde{q}} = 0 \tag{42}$$

Where,

$$G(q_m) = \begin{pmatrix} (R_e + K_2) & 0_{2\times1} \\ -W_2^T(q_m) & (R_m + k_{m2}) \end{pmatrix}$$

$$N(q_m) = \begin{pmatrix} L_{e2} & -W_2(q_m) \\ 0_{1\times2} & (R_m + l_{m2}) \end{pmatrix}$$

Proposition 2

Consider the PMSM model (1)-(3) in closed loop with the observer-controller (32)-(33) and (41)-(43). Then, the closed loop system is asymptotically stable provided that:

$$\begin{aligned} k_{e2} &> \frac{\mathrm{l_{e2}}}{4} - R_a^2 \\ k_m &> \frac{\mathrm{l_{v2}}}{4} - R_m^2 \\ \mathrm{l_{v2}} &> 4\mathrm{l_{e2}} + 4R \end{aligned} \tag{43}$$

Proof

To prove the convergence of the vector error $z_o = \left[\mathrm{e}^{\mathrm{T}}, \dot{\tilde{\mathrm{q}}}^{\mathrm{T}}\right]^T$, let consider the desired energy function error as:

$$H_{co}^{cl}(e, \dot{\tilde{q}}) = \frac{1}{2}e^T D\,e + \frac{1}{2}\dot{\tilde{q}}^T D\,\dot{\tilde{q}} \tag{44}$$

The time derivative of $H_{\mathrm{co}}^{\mathrm{cl}}$ along the trajectory (35), (42), gives:

$$\dot{H}_{\mathrm{co}}^{\mathrm{cl}} = -e^T G\,(q_m)\,\mathrm{e} - e^T N\,(q_m)\,\dot{\tilde{\mathrm{q}}} - \dot{\tilde{q}}^T(R+L)\,\dot{\tilde{\mathrm{q}}} \tag{45}$$

Which can be written as,

$$\dot{H}_{co}^{cl} = -z_o^T Q\, z_o \tag{46}$$

Where,

$$Q = \begin{pmatrix} G\,(q_m) & \frac{1}{2}N\,(q_m) \\ \frac{1}{2}N^T\,(q_m) & (R+L) \end{pmatrix}$$

Then, if matrix Q is positive, we can conclude that the closed loop system is asymptotically stable.
Matrix Q is positive if and only if the following inequality is satisfied:

$$G\ (q_m)\ (\mathrm{R}+\mathrm{L})\text{-}\frac{1}{4}N(q_m)N^T(q_m)\ >\ 0 \tag{47}$$

which can be written after calculations;

$$G\ (q_m)\ (\mathrm{R}+\mathrm{L})\text{-}\frac{1}{4}N(q_m)N^T(q_m)=\begin{pmatrix} F_{11} & F_{12} \\ F_{21} & F_{22} \end{pmatrix}$$

Where,

$$\mathrm{F}_{11}=\ (\mathrm{R_e}+K_2)(\mathrm{R_e}+L_{e2})-\frac{1}{4}(L_{e2}L_{e2}+W_2W_2^T)\mathrm{R_e}+K_2(\mathrm{R_e}+L_{e2})-\frac{1}{4}(L_{e2}L_{e2}+W_2W_2^T)$$

$$F_{12}=\frac{l_{v2}}{4}W_2$$

$$F_{21}=\frac{l_{v2}}{4}W_2^T-W_2^T(\mathrm{R_e}+L_{e2})$$

$$F_{22}=\ (\mathrm{R_m}+k_{m2})(\mathrm{R_m}+l_{v2})-\frac{l_{v2}^2}{4}$$

for simplicity, we have chosen:

$$L_{e2}=l_{e2}\ I_2\ ,\ K_2=k_2\ I_2, \text{where } \mathrm{k}_2\ >0;\ \mathrm{l}_{e2}\ >\ 0\ .$$

We note that if conditions see that the matrix Q is positive definite if conditions (43) are satisfied.
A block diagram representing the passivity-based method is show in Fig. 2.

2.3 Simulation results

The performance of the controller-observer system was investigated by simulation. We used a PMSM model, whose parameters are given in the Appendix 1.
The filter and damping parameters taken in the simulation are; a=100; b=87.5; k_{e2}=100; l_{e2}=1000 and l_{v2}=1500. We have limited the desired stator current and chosen the initial observer conditions equal to zero.
Fig. 3 shows the time response, of the motor, where a load torque τ_L of 1.35 Nm is applied to the PMSM at the starting phase and we take a speed reference of 150 rad/s. The rotor speed converges with of setting time of 0.4s. The estimated observer current and speed errors converge to zero.
Fig. 4 illustrate the time response of the closed loop system without load torque, and speed reference of (150 rad/s if t<=0.65 and –150 rad/s if t>0.65). We can see that the rotor velocity tracks its reference, and the estimation error converges.
In Fig. 5, we show the robustness of the combined controller-observer system. We take these uncertainties in the parameters of PMSM ($3R_a$, $2R_m$, $2L_d$, $2L_q$, $1.5D_m$, $0.75\lambda_m$). We note that, the rotor speed converges, but the setting time is increased lightly.

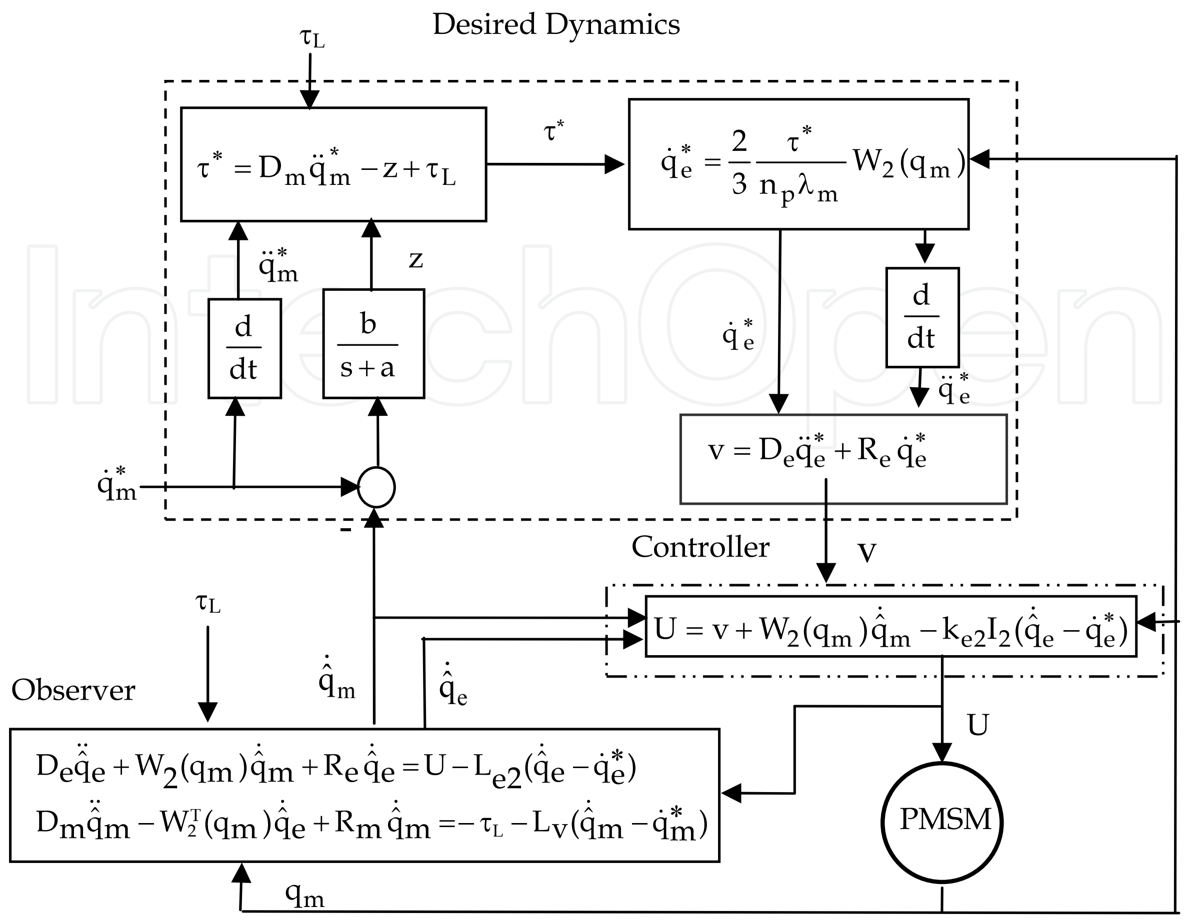


Fig. 2. Block diagrams for the passivity-based method.

a

b

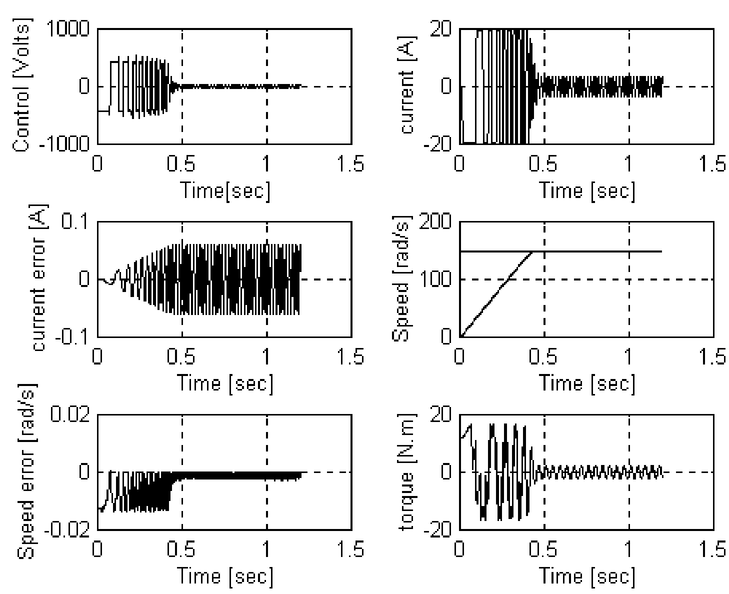


Fig. 3. Control of speed with reference 150 rd/s; a) Estimated current error; b) Estimated velocity error.

a

b

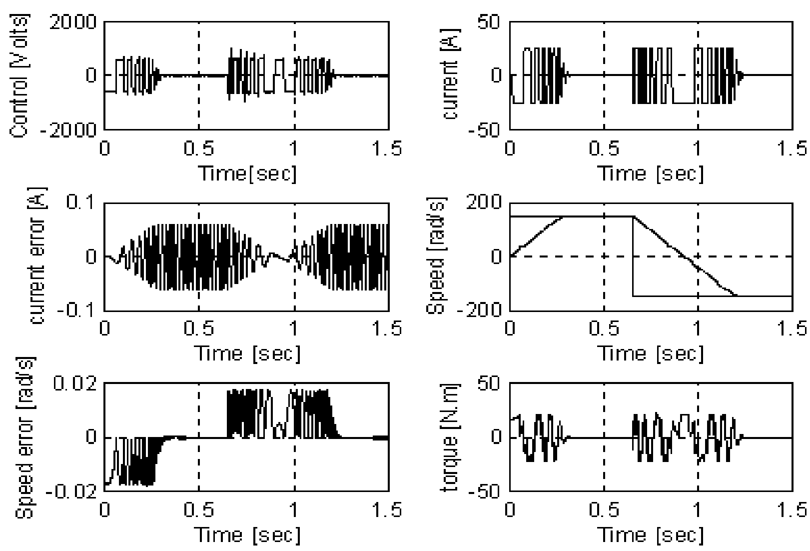


Fig. 4. Control of speed with reference (150 rd/s if t<=0.65 and –150 rd/s if t>0.65),
a) Estimated current error ; b) Estimated velocity error.

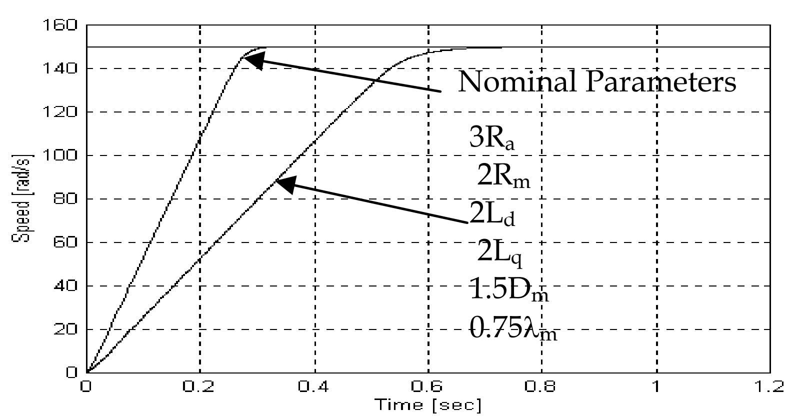


Fig. 5. Robustness test.

3. Passivity based controller design for a permanent magnet synchronous motor in dq-frame

Within this Section, a new passivity-based controller designed to force the motor to track time-varying speed and torque trajectories is presented. Its design avoids the using of the Euler-Lagrange model and destructuring since it uses a flux-based dq-modelling, independent of the rotor angular position. This dq-model is obtained through the three phase abc-model of the motor, using Park transform. The proposed control law does not compensate the model workless force terms which appear in the machine dq-model, as they have no effect on the system energy balance and they do not influence the system stability properties. Another feature is that the cancellation of the plant primary dynamics and nonlinearities is not done by exact zeroing, but by imposing a desired damped transient. The effectiveness of the proposed control is illustrated by numerical simulation results.

The Section 2 is organized as follows. The PMSM dq-model and the inner current loop design are presented at Subsection 2.2. In Subsection 3, the passivity property of the PMSM in the dq-reference frame is introduced. Subsection 2.4 deals with the computation of the current, flux and the torque references. The passivity property of the closed loop system and the resulting control structure are given in Subsections 2.5 and 2.6, respectively. Simulation results are presented in Subsection 2.7. Subsection 2.8 concludes this Section. The proof of the passivity property of the PMSM in the dq frame is given. The analysis and proof of the exponential stability of the flux tracking error is introduced. Subsections 2.5 contain the proof of the passivity property of the closed loop system.

3.1 Permanent-magnet synchronous motor model in dq frame

The PMSM uses buried rare earth magnets. Its electrical behaviour is described here by the well known dq model (Krause et al., 2002), given by Equation (48):

$$L_{dq}\dot{i}_{dq} + R_{dq}i_{dq} + n_p\omega_m\Im L_{dq}i_{dq} + n_p\omega_m\Im\psi_f = v_{dq} \tag{48}$$

In this equation the following notations have been employed:

$$L_{dq} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix};\ i_{dq} = \begin{bmatrix} i_d \\ i_q \end{bmatrix};\ R_{dq} = \begin{bmatrix} R_S & 0 \\ 0 & R_S \end{bmatrix};\ \psi_f = \begin{bmatrix} \phi_f \\ 0 \end{bmatrix};\ \Im = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix};\ v_{dq} = \begin{bmatrix} v_d \\ v_q \end{bmatrix}$$

In the above-presented relations, L_d and L_q: are the stator inductances in dq frame, R_S: is the stator winding resistance, ϕ_f: is the flux linkages due to permanent magnets, n_p: is the number of pole-pairs, ω_m: is the mechanical speed, v_d and v_q: are the stator voltages in dq frame, i_d and i_q: are the stator currents in dq frame.

The mechanical equation of the PMSM is given by:

$$J\dot{\omega}_m + f_{VF}\omega_m = \tau_e - \tau_L \tag{49}$$

where J is the rotor moment of inertia, f_{VF} is the viscous friction coefficient, and τ_L is the load torque.

The electromagnetic torque τ_e can be expressed in the dq frame as follows:

$$\tau_e = \frac{3}{2}n_p\left(\left(L_d - L_q\right)i_d i_q + \phi_f i_q\right) \tag{50}$$

The rotor position θ_m is given by Equation (51):

$$\dot{\theta}_m = \omega_m \tag{51}$$

where ψ_d and ψ_q are the flux linkages in dq frame.

The interdependence between the flux linkage motor ψ_{dq} and the current vector i_{dq} can be expressed as follow (Krause et al., 2002):

$$\begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} = L_{dq}i_{dq} + \psi_f \tag{52}$$

where ψ_d and ψ_q are the flux linkages in dq frame.

Substituting i_{dq} value obtained by Relation (52) in Equations (48) and (50), yields:

$$\dot{\psi}_{dq} + n_p \omega_m \Im \psi_{dq} = v_{dq} - R_{dq} i_{dq} \tag{53}$$

$$\tau_e = -\frac{3}{2} n_p \psi_{dq} \Im i_{dq} \tag{54}$$

Current controlled dq-model of PMSM

Let us define the state model of the PMSM using the state vector $\begin{bmatrix} \psi_d & \psi_q & \omega_m & \theta_m \end{bmatrix}^T$ and Equations (49), (51), (53) and (54). The reference value of the current vector i_{dq} is denoted by:

$$i_{dq}^* = \begin{bmatrix} i_d^* \\ i_q^* \end{bmatrix}$$

The proportional-integral (PI) current loops, used to force $\begin{bmatrix} i_d & i_q \end{bmatrix}^T$ to track the reference $\begin{bmatrix} i_d^* & i_q^* \end{bmatrix}^T$, are of the form of equations below:

$$v_d = k_{dp}\left(i_d^* - i_d\right) + k_{di}\int_0^t \left(i_d^* - i_d\right) \mathrm{d}t, \quad k_{dp}, k_{di} > 0 \tag{55}$$

$$v_q = k_{qp}\left(i_q^* - i_q\right) + k_{qi}\int_0^t \left(i_q^* - i_q\right) \mathrm{d}t, \quad k_{qp}, k_{qi} > 0 \tag{56}$$

Assuming that by the proper choice of positive gains k_{dp}, k_{di}, k_{qp}, k_{qi}, these loops work satisfactory. Then, the reference vector i_{dq}^* can be considered as control input for the PMSM model. This result on the simplified dynamic dq-model of the PMSM given below:

$$\dot{\psi}_{dq} + n_p \omega_m \Im \psi_{dq} = -R_{dq} i_{dq}^* \tag{57}$$

$$J\dot{\omega}_m + f_{VF}\omega_m = \tau_e - \tau_L \tag{58}$$

$$\dot{\theta}_m = \omega_m \tag{59}$$

$$\tau_e = -\frac{3}{2} n_p \psi_{dq}^T \Im i_{dq}^* \tag{60}$$

This simplified form of the PMSM model is further used to design the control input i_{dq}^* using the passivity approach.

3.2 Passivity property of dq-model

Lemma 3

The PMSM represents a strictly passive system if the reference vector, of the stator currents, i_{dq}^* and the flux linkage vector, ψ_{dq} are considered as the input and the output vectors, respectively.

Proof

First, multiply both sides of Equation (57) by $\frac{\psi_{dq}^T}{R_s}$, yields

$$\psi_{dq}^T i_{dq}^* = -\frac{1}{2R_s}\frac{\mathrm{d}\left(\psi_{dq}^T \psi_{dq}\right)}{\mathrm{d}t} \tag{61}$$

where ψ_{dq}^T is the transposed of vector ψ_{dq}.

Note that the term $\frac{n_p \omega_m}{R_s}\psi_{dq}^T \Im \psi_{dq}$ does not appear on the right-hand side of (61), since $\psi_{dq}^T \Im \psi_{dq} = 0$ due to skew-symmetric property of the matrix $\Im$. Integrating both sides of Equation (61), yields

$$\int_0^t \left(\psi_{dq}^T i_{dq}^*\right)\mathrm{d}t = -\frac{1}{2R_s}\left(\psi_{dq}^T \psi_{dq}\right)(t) + \frac{1}{2R_s}\left(\psi_{dq}^T \psi_{dq}\right)(0) \tag{62}$$

Consider that the i_{dq}^* is the input vector and ψ_{dq} is the output vector. Then, with positive definite function

$$V_f = \frac{1}{2}\psi_{dq}^T \psi_{dq} \tag{63}$$

the energy balance Equation (62) of the PMSM becomes

$$\int_0^t \left(\psi_{dq}^T i_{dq}^*\right)\mathrm{d}t = -\frac{1}{R_s}V_f(t) + \frac{1}{R_s}V_f(0) \tag{64}$$

This means that the PMSM is a strictly passive system (Ortega et al., 1997). Thus, the term $n_p \omega_m R_{dq}^{-1}\psi_{dq}^T \Im \psi_{dq}$ has no influence on the energy balance and on the asymptotic stability of the PMSM also; it is identified as the workless forces term.

3.3 Analysis of tracking errors convergence using passivity-based method

The desired value of the flux linkage vector ψ_{dq} is:

$$\psi_{dq}^* = \begin{bmatrix} \psi_d^* \\ \psi_d^* \end{bmatrix} \tag{65}$$

and the difference between ψ_{dq} and ψ_{dq}^* representing flux tracking error, as:

$$e_f = \begin{bmatrix} e_{fd} \\ e_{fq} \end{bmatrix} = \psi_{dq} - \psi_{dq}^* \tag{66}$$

Rearranging Equation (66)

$$\psi_{dq} = e_f + \psi_{dq}^* \tag{67}$$

Substituting Equation (16) in Equation (68), yields

$$\dot{e}_f + n_p\omega_m \Im e_f = -R_{dq} i^*_{dq} - \left(\dot{\psi}^*_{dq} + n_p\omega_m \Im \psi^*_{dq}\right) \tag{68}$$

The aim is to find the control input i^*_{dq} which ensures the convergence of error vector e_f to zero. The energy function of the closed-loop system is defined as

$$V(e_f) = \frac{1}{2} e_f^T e_f \tag{69}$$

Taking the time derivative of $V\left(e_f\right)$ along the Trajectory (17), gives

$$\dot{V}\left(e_f\right) = -e_f^T \left(R_{dq} i^*_{dq} + \dot{\psi}^*_{dq} n_p\omega_m \Im \psi^*_{dq}\right) \tag{70}$$

Note that the term $n_p\omega_m e_f^T \Im e_f = 0$ due to the skew-symmetric property of the matrix $\Im$.
The convergence to zero of the error vector e_f is ensured by taking

$$i^*_{dq} = -R_{dq}^{-1}\left(\dot{\psi}^*_{dq} + n_p\omega_m \Im \psi^*_{dq}\right) + R_{dq}^{-1} K_f e_f \tag{71}$$

where $K_f = \begin{bmatrix} k_{fd} & 0 \\ 0 & k_{fq} \end{bmatrix}$ with $k_{fd} > 0$ and $k_{fq} > 0$.

The control input signal, i^*_{dq} consists of two parts: the term which encloses the reference dynamics and the damping term injected to make the closed-loop system strictly passive.
The PBCC ensures the exponential stability of the flux tracking error.

3.3.1 Proof of the exponential stability of the flux tracking error

Consider the quadratic Function (69) and its time derivative in Equation (70). Substituting i^*_{dq} of (71) in (70), yields

$$\dot{V}\left(e_f\right) = -e_f^T K_f e_f \le -\lambda_{\min}\left\{K_f\right\}\left\|e_f(t)\right\|^2,\ \forall\ t \ge 0 \tag{72}$$

where $\lambda_{\min}\left\{K_f\right\} > 0$ is the minimum eigenvalue of the matrix K_f and $\|.\|$ is the standard euclidian vector norm.
The square of the standard Euclidian norm of the vector e_f is given as:

$$\left\|e_f\right\|^2 = e_{fd}^2 + e_{fq}^2 = e_f^T e_f \tag{73}$$

Which combined with Relation (69), gives

$$V(e_f) = \frac{1}{2} e_f^T e_f \le \left\|e_f\right\|^2,\ \forall\ t\ \ge 0 \tag{74}$$

Multiplying both sides of (74) by $(-\lambda_{\min}\left\{K_f\right\})$, leads to

$$\left(-\lambda_{\min}\left\{K_f\right\}\right) V(e_f) \ge \left(-\lambda_{\min}\left\{K_f\right\}\right)\left\|e_f\right\|^2,\ \forall\ t\ \ge 0 \tag{75}$$

which combined with Relation (72), gives

$$\dot{V}\left(e_f\right) \le -\lambda_{\min}\left\{K_f\right\} V(e_f), \; \forall \text{ t } \ge 0 \tag{76}$$

Integrating both sides of the Inequality (76), yields

$$V(e_f) \le V(0)e^{-\rho_f t}, \; \forall \text{ t } \ge 0 \tag{77}$$

where $\rho_f = \lambda_{\min}\left\{K_f\right\} > 0$. Considering the Relation (74) at t=0, and multiplying it by $e^{-\rho_f t}$, gives

$$V(0)e^{-\rho_f t} \le \left\|e_f(0)\right\|^2 e^{-\rho_f t} \tag{78}$$

which combined with Relation (77), leads to the following inequality:

$$V(e_f) \le \left\|e_f(0)\right\|^2 e^{-\rho_f t}, \; \forall \text{ t } \ge 0 \tag{79}$$

The Inequalities (74) and (79) give that:

$$\left\|e_f(t)\right\| = \left\|e_f(0)\right\| e^{-\frac{\rho_f}{2}t} \tag{80}$$

The Equation (80) shows that, the flux tracking error e_f is exponentially decreasing with a rate of convergence of $\rho_f/2$.

3.3.2 Flux reference computation

The computation of the control signal i_{dq}^* requires the desired flux vector ψ_{dq}^*. If the direct current i_d in the dq frame is maintained equal to zero, then the PMSM operates under maximum torque. Under this condition and using Equation (52), results in

$$\psi_d^* = \varphi_f \tag{81}$$

$$\psi_q^* = L_q i_q^* \tag{82}$$

The torque set-point value τ_e^* corresponding to ψ_{dq}^* is given by Equation (54). Substituting ψ_d^* from (81) and i_q^* from (82) in (54), it results that:

$$\tau_e^* = \frac{3}{2}\frac{n_p \varphi_f}{L_q}\psi_q^* \tag{83}$$

Therefore the value of the flux reference is deduced as

$$\psi_q^* = \frac{2}{3}\frac{L_q}{n_p \varphi_f}\tau_e^* \tag{84}$$

3.3.3 Torque reference and load torque computation

The desired torque τ_e^* is computed by the expressions (28)-(29).

In practical applications, the load torque is unknown, therefore it must be estimated. For that purpose, an adaptive law (Kim et al., 1997) has been used:

$$\dot{\hat{\tau}}_L = -k_L(\omega_m - \omega_m^*), \quad \mathrm{k_L} > 0 \tag{85}$$

3.4 Passivity property of the closed loop system in the general dq reference frame

Lemma 4

The closed loop system represents a strictly passive system if the desired dynamic output vector given by

$$\vartheta = -R_{dq}^{-1}\left(\dot{\psi}_{dq}^* + n_p\omega_m \Im \psi_{dq}^*\right) \tag{86}$$

and the flux linkage vector ψ_{dq} are considered as input and output, respectively.

Proof

Substituting the control input vector i_{dq}^* from (71) in Equation (57), gives

$$\dot{\psi}_{dq} + n_p\omega_m \Im \psi_{dq} = -R_{dq}\vartheta - K_f e_f \tag{87}$$

where ϑ is given by Relation (86).

Multiplying both sides of Equation (87) by $\dfrac{\psi_{dq}^T}{R_s}$

$$\psi_{dq}^T\vartheta = -\frac{1}{2R_s}\frac{\mathrm{d}\left(\psi_{dq}^T\psi_{dq}\right)}{\mathrm{d}t} - \psi_{dq}^T K_f e_f \tag{88}$$

The term $\dfrac{n_p\omega_m}{R_s}\psi_{dq}^T \Im \psi_{dq}$ disappears from (88), since $\psi_{dq}^T \Im \psi_{dq} = 0$ due to skew-

-symmetric property of the matrix $\Im$. According to Relation (80), the flux tracking error e_f is exponentially decreasing. Thus, the term $\psi_{dq}^T K_f e_f$ becomes insignificant. And Equation (88) is writes as

$$\psi_{dq}^T\vartheta = -\frac{1}{2R_s}\frac{\mathrm{d}\left(\psi_{dq}^T\psi_{dq}\right)}{\mathrm{d}t} \tag{89}$$

Integrating both sides of Equation (45), yields

$$\int_0^t \left(\psi_{dq}^T\vartheta\right)\mathrm{d}t = -\frac{1}{2R_s}\left(\psi_{dq}^T\psi_{dq}\right)(t) + \frac{1}{2R_s}\left(\psi_{dq}^T\psi_{dq}\right)(0) \tag{90}$$

Let us consider the positive definite function V_f from Relation (67). The Energy Balance (90) of the closed loop system becomes

$$\int_0^t \left(\psi_{dq}^T\vartheta\right)\mathrm{d}t = -\frac{1}{R_s}V_f(t) + \frac{1}{R_s}V_f(0) \tag{91}$$

The previous relation shows that, the closed-loop system is a strictly passive (Ortega et al., 1997). Thus, the term $\frac{n_p \omega_m}{R_s} \psi_{dq}^T \Im \psi_{dq}$ has no influence on the energy balance and the asymptotic stability of the closed-loop system; it is identified as the workless forces term.

3.5 Passivity based current controller structure for PMSM

The design procedure of the passivity-based current controller for PMSM leads to control structure described by the block diagram in Fig. 6. It consists of three main parts: the load torque estimator given by Equation (85), the desired dynamics expressed by the Relations (28)-(29), (81)-(85), and the controller given by Equations (55), (56) and (71). In this design the imposed flux vector, ψ_{dq}^*, is determined from maximum torque operation conditions allowing the computation of the desired currents i_{dq}^*. Furthermore, the load torque is estimated through speed error, and directly taken into account in the desired dynamics.

The inner loops of the PMSM control are based on well known proportional-integral controllers. Park transform is used for passing electrical variables between the three-phase and dq frame.

The actuator used in the control application is based on a PWM voltage source inverter. Voltage, currents, rotational speed and PMSM angular position are considered measurable variables.

3.6 Simulation results

The parameters of the PMSM used for testing the previously exposed control structure are given in Appendix. 2.

The plant and its corresponding control structure of Fig.6 are implemented using Matlab and Simulink software environment. It employs the PMSM model represented by the Equations (48)-(51) whose parameters are given in appendix 2. The chosen solver is based on Runge-Kutta algorithm (ODE4) and employs an integration time step of 10^{-4} s. The parameter values of the control system are determined using the procedures detailed in Subsections 2.2 and 2.4 as follows. From the imposed pole locations, the gains of the current PI controller are computed as: k_{dp}=95, k_{di}=0.85, k_{qp}=95, and k_{di}=0.8. The gains concerning the desired torque are set at a=75 and b=400 using pole placement method also. The damping parameters values have been obtained by using a trial-and-error procedure starting from guess values based on the stability Condition (71); their final values are k_{fd} = k_{fq} = 650. The gain of the load torque adaptive law is set to k_L=6, value which ensures the best asymptotic convergence of the speed error.

In all tests performed in this study, the following signals have been considered as representative for performance analysis: rotational speed (Fig. 7(a)), line current (Fig. 7(b)), electromagnetic torque (Fig. 7(c)), the stator voltages in dq frame (Fig. 7(d)), zoom of voltage at the output of the inverter (Fig. 7(e)), and zoom of line current (Fig. 7(f)). Fig. 7 shows the motor response to square speed reference signal with magnitude ±150 rad/s. This study concerns the robustness test of the designed control system to disturbances. To this end, a load torque step of τ_L=10 N·m has been applied at time 0.5 second and has been removed at time 4.5 seconds (see Fig. 7). The results of Fig. 7 show that the response of the rotor speed to the disturbance is quite and the electromagnetic torque, τ_e, have been increased to a value corresponding to the load applied. The rotational speed and line current tracks quickly the reference, without overshoot and all other signals are well shaped. The peaks visible on the electromagnetic torque evolution are due to high gradients imposed to the rotational speed. In practice, these peaks can be easily reduced by limiting the speed reference changing rate and by limiting the imposed current i_q^* value. However, such situation has been chosen for a better presentation of the control law capabilities and performances.

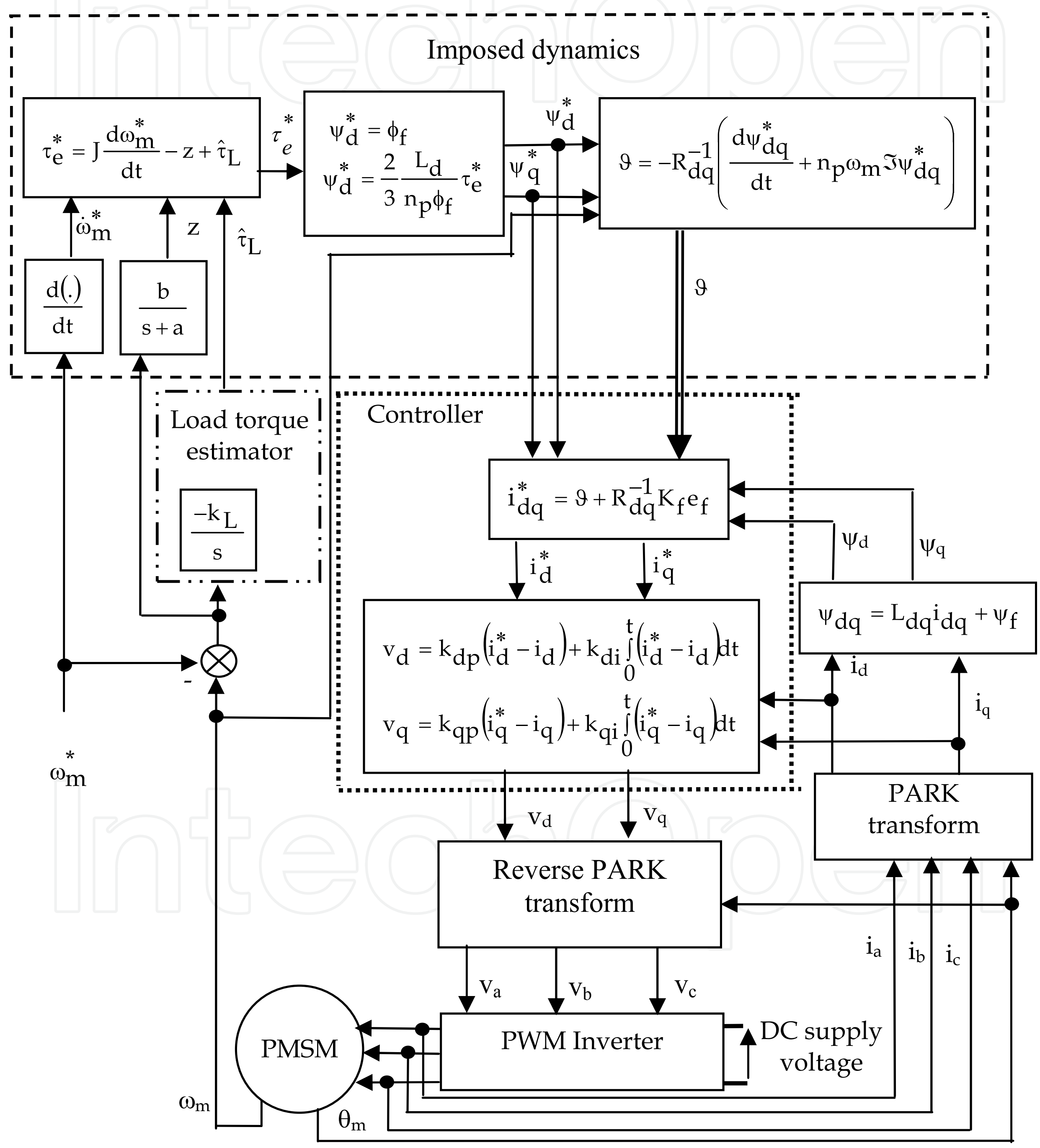


Fig. 6. The block diagram for the passivity-based current controller.

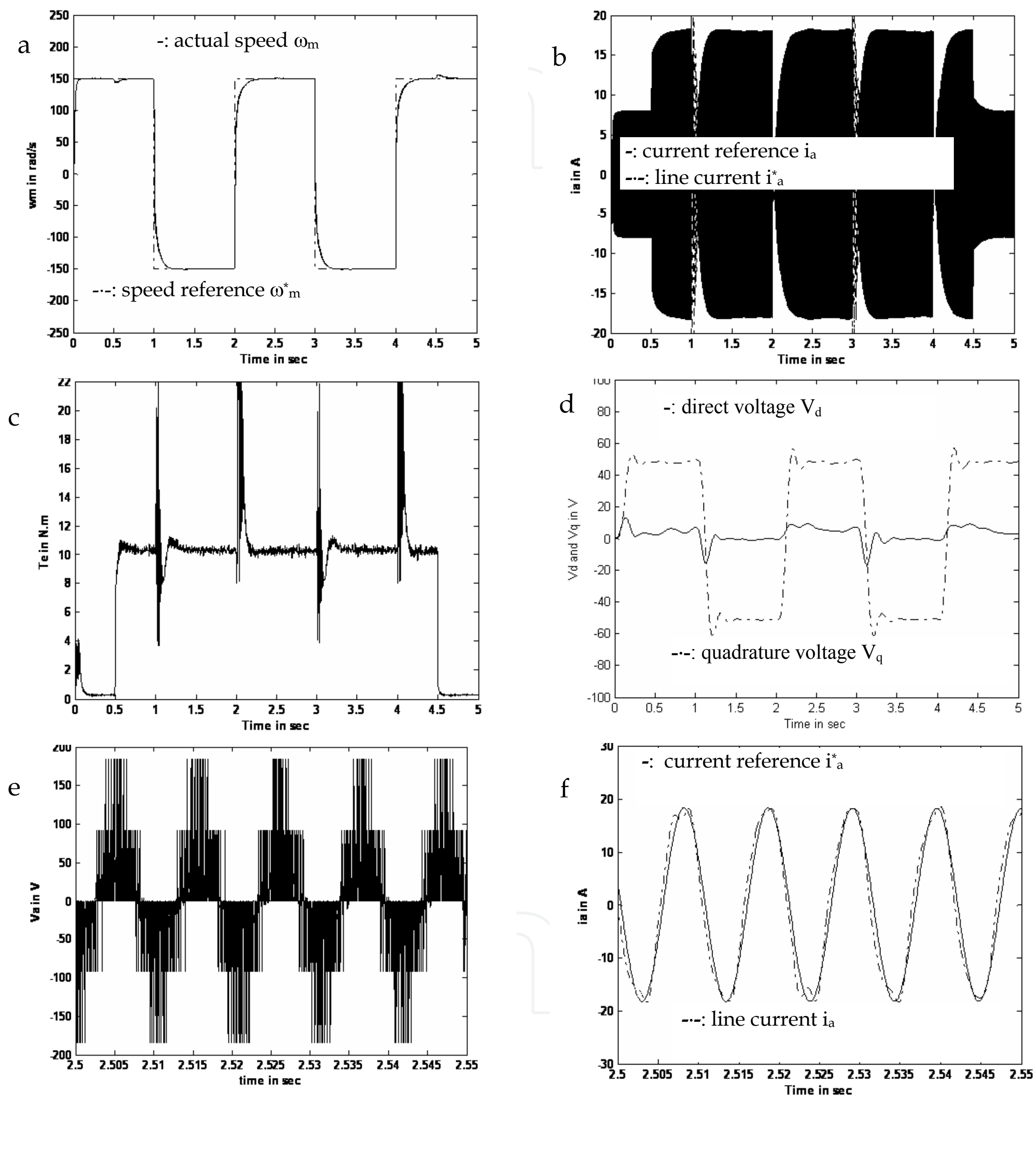


Fig. 7. Motor response to square speed reference signal with a load torque step of 10 Nm from t=0.5s to t=4.5s.

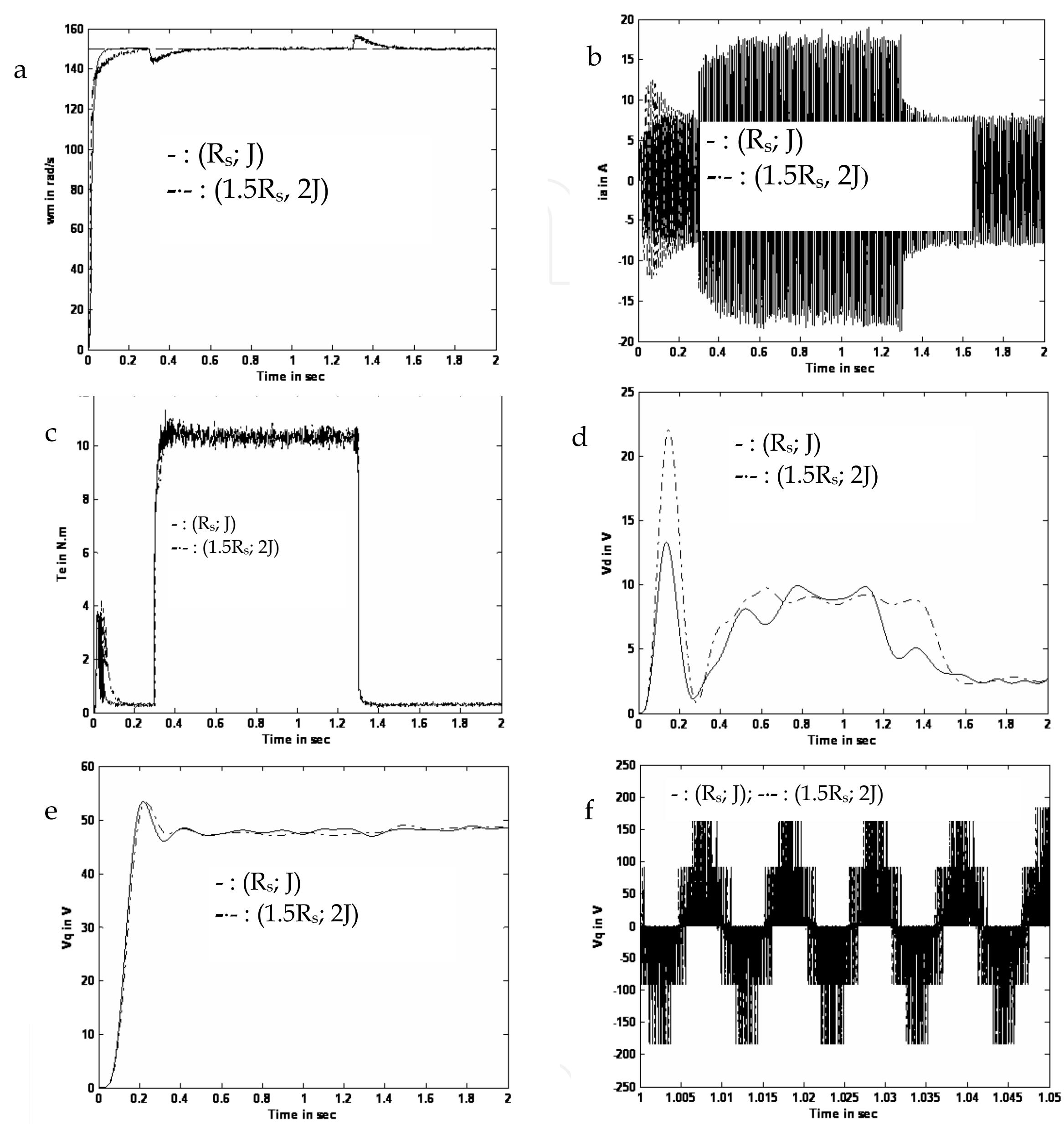


Fig. 8. Motor response to step reference with a change of +50% of the stator winding resistance R_s and a change of +100% of the inertia moment J.

A test of robustness at parameter changes has been performed. As presented in Fig. 8, a simultaneous change of +50% of the stator winding resistance R_s and +100% of the moment inertia J. The change of the stator winding resistance, R_s, affects slightly the dynamic motor response. This is due to the fact that the electrical time constant ρ_f of closed-loop system appearing in Equation (80) is compensated by the imposed damping gain, K_f, from Equation (71). However, a change of +100% inertia moment J increases the mechanical time constant and hence the rotor speed settling time (see Fig. 2.5). The designed PBCC is based only on the electrical part of the PMSM and has no direct compensation effect on the mechanical part.

4. Conclusion

In the section 2, a strategy for designing PMSM control system that requires only rotor position and stator voltage measurements was presented. To this end, the passivity approach to design a controller-observer is adopted. It was shown that this strategy can provide asymptotically stabilizing solutions to the output feedback motor tracking problem. It is shown from simulation results that the robustness of the combined controller-observer with respect to the load and model uncertainties. This is mainly due to the fact that both of the controller and observer exploit the physical structure of the PMSM system and the injection of the high damping.

A new passivity-based speed control law for a PMSM has been developed in the section 3. The proposed control law does not compensate the model workless force terms as they have no effect on the system energy balance. Therefore, the identification of these terms is a key issue in the associated control design. Another feature is that the cancellation of the plant primary dynamics is not done by exact zeroing but by imposing a desired damped transient. The design avoids the using of the Euler-Lagrange model and destructuring (singularities effect) since it uses a flux-based dq-modelling, independent of the rotor angular position. The inner current control loops which have been built using classical PI controllers preserve the passivity property of the current-controlled synchronous machine.

Unlike the majority of the nonlinear control methods used in the PMSM field, this control loop compensates the nonlinearities by means of a damped transient. Its computation aims at imposing the currents set-points based on the flux references in the dq-frame. These latter variables are computed based on the load torque estimation by imposing maximum torque operation conditions.

The speed control law contains a damping term ensuring the system stability and the adjustment of the tracking error convergence speed. The obtained closed-loop system allows exponential zeroing of the speed error, also preserving the passivity property.

Simulation studies show the feasibility and the efficiency of the proposed controller. This controller can be easily included into control structures developed for current-fed induction motor commonly used in industrial applications. Its relatively simple structure should not involve significant hardware and software implementation constraints.

Appendix 1

R_a= 2 Ω; R_m ; 0.00019 Nm/rd/s; λ_m =0.2 Wb ; n_p=2 ; L_d=3.1 mH; L_q= 3.1 mH; D_m=0.024 Kgm2; I_n=15 A; V_n=250 V; P_n=3.75 KW; N=4000 r n/mn.

Appendix 2

Rated power = 6 Kw; Rated speed = 3000 rpm; Stator winding resistance = 173.77 e-3 Ω ; Stator winding direct inductance = 0.8524 e-3 H; Stator winding quadrate inductance = 0.9515 e-3 H; Rotor flux = 0.1112 Wb; Viscous friction = 0.0085 Nm/rad/s; Inertia = 48 e-4 kg.m^2; Pairs pole number = 4; Nominal current line = 31 A; Nominal voltage line = 310 V and the machine type is Siemens 1FT6084-8SK71-1TGO.

5. References

Achour, AY.; Mendil, B. (2007). Commande basée sur la passivité associée aux modes de glissements d'un moteur synchrone à aimants permanents. *JESA*, vol.41, No3-4, April 2007), pp 311-332, ISBN 978-2-7462-1854-3

Ailon, A.; Ortega, R. (1993). An observer-based set-point controller for robot manipulators with flexible joints. *System Control Literature* 1993, vol.21, No.4, (October 1993), pp 329-335

Arnold, V I. (1989). *Mathematical Methods of Classical mechanics*, Springer, ISBN 0-387-96890-3, New York, USA, 1989

Berghis, H.; Nijmeijer, H. (1993). A passivity approach to controller-observer design for robots. *IEEE Transaction on robotic and automatic,* vol.9, No.6, (December 1993), pp-754, ISSN 1042-296X

Gökder, LU.; Simaan, MA. (1997). A passivity-based control method for Induction motor control. *IEEE Transactions on Industrial Electrical*. Vol.44, No.5, (October 1997), pp 688-695, ISSN 0278-0046

Kim, KC.; Ortega, R.; Charara, A.; Vilain, JP. (1997). Theoretical and experimental Comparison of two nonlinear controllers for current-fed induction motors. *IEEE Transactions on Control System Techniques*, vol.5, No.5, (May 1997), pp 338-348, ISSN 1063-6536

Ortega, R.; Spong, M. (1989). Adaptive motion control of rigid robots: A tutorial, *Automatica*, vol.25, No.6, (November 1989), pp 877-888

Ortega, R.; Nicklasson, PJ.; Espinoza–Pérez, G. (1996). On speed control of induction motors, *Automatica*, vol.3, No.3, (March 1996), pp 455-466

Ortega, R.; Nicklasson, PJ.; Espinoza–Pérez, G. (1997). Passivity-based controller of a Class of Blondel-Park transformable electric machines, *IEEE Transactions on Automatic Control*, vol.42, No.5, (May 1997), pp 629-647, ISSN 0018-9286

Ortega, R.; Loria, A.; Nicklasson, PJ. (1998). *Passivity-based control of Euler-Lagrange systems*, , ISBN 1-85233-016-3, New York, USA, 1998

Ortega, R.; García-Canseco, E. (2004). Interconnection and damping assignment passivity-Based control: Towards a constructive procedure-Part I, *Proceedings of 43rd IEEE conference on Decision and Control*, pp 3412-3417, ISBN 0-7803-8682-5, Atlantis, Island, Bahamas, December 14-17, 2004

Ortega, R.; García-Canseco, E. (2004). Interconnection and damping assignment passivity-Based control: Towards a constructive procedure-Part II, *Proceedings of 43rd IEEE conference on Decision and Control*, pp 3418-3423, ISBN 0-7803-8682-5, Atlantis, Paradise Island, Bahamas, December 14-17, 2004

Petrović, V.; Ortega, R.; Stanković, AM. (2001). Interconnection and damping assignment approach to control of Pm synchronous motors, *IEEE Transactions on Control System Techniques*, vol.9, No.6, (November 2001), pp 811-820, ISSN 1063-6536

Qiu, J.; Zhao, G. (2006). PMSM control with port-controlled Hamiltonian theory, *Proceedings of 1st International Conference on Innovative Computing, Information and Control (ICICIC'06)*, pp 275-278, ISBN 0-7695-2616-0, Ville, Pay, August 30-31, 2006

Sira-Ramirez, H.; Ortega, R.; Perez-Moreno, R. ; Garcia-Esteban, M. (1995). A sliding mode controller-observer for DC-to-DC power converters: a passivity approach, *Proceedings* 34th IEEE conference on Decision and Control, pp 3379-3384, ISBN 0-7803-2685-7, New Orleans, LA, 13-15 December, 1995

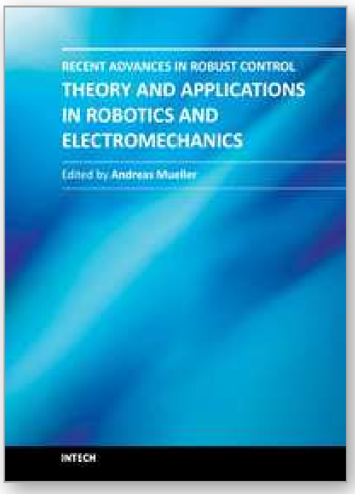
Sudhoff, S.; Wasynczuk, O.; SD Krause, PC. (2002). *Analysis of electric machinery and drive Systems*, Wiley-IEEE Press, ISBN 9780470544167, New York, USA, 2002.

Takegaki, M.; Arimoto, S. (1981). A new feedback for dynamic control of manipulators, *Transaction of the ASME, Journal of Dynamic Systems Measurements Control*, vol. 103,.2, (June 1981), pp 119-125

Travieso-Torres, JC.; Duarte Mermoud, MA. ; Estrada, JL. (2006). Tracking control of cascade Systems based on passivity: The non-adaptive and adaptive cases, *ISA Transactions*, Vol.45, No.3, (July 2006), pp 435-445

Travieso-Torres, JC.; Duarte Mermoud, MA. (2008). Two simple and novel SISO controllers for induction motors based on adaptive passivity, *ISA Transactions*, vol.47, No.1, (January 2008), pp 60-79

Van der Schaft, A. (2000). *L_2-Gain and Passivity Techniques in Nonlinear Control*, Springer, ISBN 1-85233-073-2, London, King Doom, 2000



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