©Journal of the Technical University Sofia, branch Plovdiv
"Fundamental Sciences and Applications", Vol. 16, 2011
International Conference Engineering, Technologies and Systems
TechSys 2011
BULGARIA

STRICTLY ORIENTATED STATE SPACE MODELS OF PERMANENT-MAGNET SYNCHRONOUS MOTORS FOR FEEDBACK LINEARIZATION CONTROL

RUMEN MISHKOV, VALENTIN PETROV

Abstract. *This article presents strictly orientated state space models of permanent-magnet synchronous motors based on a two-phase dq mathematical model in rotating reference coordinates synchronous with the electrical rotor speed. Park-Clarke transformations and the equivalent power principle are used. The state space models are specially suited for the application of advanced control differential geometric design approaches including multi-input multi-output feedback linearization and adaptive systems design. The models are simulated for illustration.*

Keywords: *nonlinear systems, synchronous motors, state space models, nonlinear transformations.*

1. Introduction

The interest in permanent-magnet motors during the last thirty years is due to the multiple advantages they offer. The permanent-magnet synchronous motors (PMSM) are part of this motor class. The PMSM have sinusoidal back electro-motive force and require sinusoidal stator currents to produce constant torque [8,9]. The difference between the conventional synchronous motor and PMSM is that PMSM has no brushes, commutator, and field winding. The excitation of these motors is achieved by strong rare-earth permanent magnets in the rotor providing very powerful excitation field without the need of excitation current and no electrical losses in the rotor. As a result they are very efficient energetically while generating comparatively high torque. The lack of brushes leads to longer life span of PMSM. These motors have maximal ratio between generated torque and motor inertia or total mass conditioned by the high power density of PMSM compared to induction motors or wound rotor synchronous motors. PMSM are preferred motors for high performance applications in robotics, industrial and aerospace implementations.

The mathematical models of PMSM can be derived from the dq models of the conventional synchronous motor by removing the damper windings equations and the excitation current dynamics [7]. Most frequently the three-phase PMSM are used in practice. The transformation of three-phase variables into two-phase coordinates, known also as Park-Clarke transformation, allows the two-phase mathematical model [7,5,4] to be utilized for simulation and investigation. The two-phase dq model is widely used for analysis and design with various types of control methods [10,5,1,2,3,6,4].

The paper presents a derivation of a two-phase dq mathematical model for PMSM, reflection of the three-phase variables in two-phase dq coordinates by Park-Clarke transformation and application of the equivalent power principle. Equivalent replacement scheme is proposed for the electrical processes taking place in the PMSM. The two possible state space models suitable for multi-variable feedback linearization control methods are proposed. These models are simulated in no load mode.

2. PMSM mathematical modeling

This section presents the derivation of a three-phase PMSM model based on the two-phase synchronous motor in dq coordinates. The stator of the three-phase motor is represented by two windings only. The rotor magnets are modeled as a current source or flux linkage source concentrated along one of the axis. Figure 1 shows the PMSM in dq coordinates. The stator windings are shifted at 90 electrical degrees with respect to one another while the rotor winding is rotated with respect to the d-axis stator winding at the electrical angle. The motor model is derived with the following assumptions

♦ The stator windings are balanced with sinusoidal magnetomotive force (MMF).

♦ The dependency of the inductance on the rotor position is sinusoidal.

♦ The saturation and the parameters changes are neglected.

The stator voltages along the q and d axes are

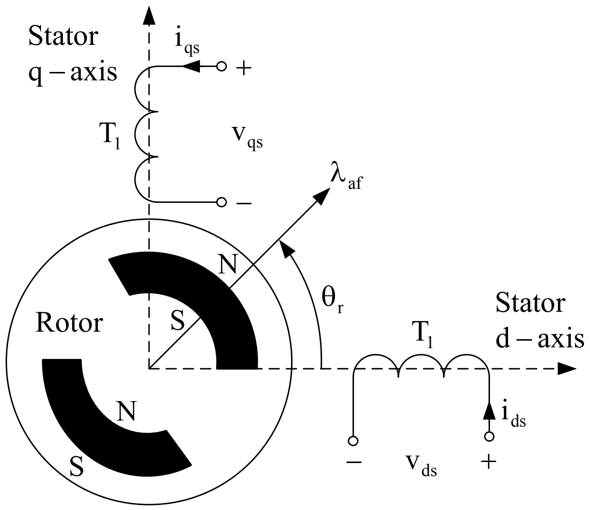

Figure 1. *Two-phase PMSM*

determined as sums of voltage drops and flux linkages derivatives in the respective windings

$$v_{qs} = R_q i_{qs} + \frac{d\lambda_{qs}}{dt},$$

$$v_{ds} = R_d i_{ds} + \frac{d\lambda_{ds}}{dt},$$

where R_q, R_d are the stator windings resistances, i_{qs}, i_{ds} are the stator currents, and λ_{qs}, λ_{ds} are the stator windings flux linkages. The latter can be writen as a sum of flux linkages due to their own excitation and mutual flux linkages resulting from other winding current and magnet sources. The rotor flux linkages have components on the q and d axes as they are assumed to be concentrated along the axis of the instantaneous rotor position. Thus, the stator windings flux linkages are written as

$$\lambda_{qs} = L_{qq} i_{qs} + L_{qd} i_{ds} + \lambda_{af} \sin\theta_r,$$

$$\lambda_{ds} = L_{dq} i_{qs} + L_{dd} i_{ds} + \lambda_{af} \cos\theta_r,$$

where θ_r is the instantaneous rotor position and λ_{af} is the rotor flux linkage. As the windings are balanced the resistances are equal and denoted as $R_s = R_q = R_d$. Then the equations of the stator voltages take the form

$$\begin{aligned} v_{qs} &= R_s i_{qs} + \lambda_{af} \frac{d\sin\theta_r}{dt} + i_{qs} \frac{dL_{qq}}{dt} \\ &+ L_{qd} \frac{di_{ds}}{dt} + i_{ds} \frac{dL_{qd}}{dt} + L_{qq} \frac{di_{qs}}{dt} \end{aligned} \tag{1}$$

$$\begin{aligned} v_{ds} &= R_s i_{ds} + i_{qs} \frac{dL_{qd}}{dt} + \lambda_{af} \frac{d\cos\theta_r}{dt} \\ &+ L_{qd} \frac{di_{qs}}{dt} + L_{qd} \frac{di_{qs}}{dt} + i_{qs} \frac{dL_{dd}}{dt} \end{aligned} \tag{2}$$

where L_{qq}, L_{dd} are the windings self-inductances, L_{qd}, L_{dq} are the mutual inductances due to the currents i_d and i_q, respectively. The flux linkages are functions of the rotor position. The magnets are aligned with the d-axis stator winding when $\theta_r = 0$. At that position the length of the flux path in the air is increased by the magnets' thickness. The relative permeability of the magnets is almost equal to the relative permeability of the air, and therefore, the reluctance of the flux in this path is increased, hence, the winding inductance is decreased. This position corresponding to minimum inductance is denoted as L_d. When $\theta_r = 90$ electrical degrees, the magnet flux path along the d-axis does not cross the magnets at all but crosses the iron of the rotor and the air gaps on both sides. This position corresponding to maximum inductance is denoted as L_q. The windings are distributed to provide sinusoidal MMF. Hence, the self-inductances can be modeled as cosinusoidal functions of twice the rotor position according to the d-axis. Then the q and d windings self-inductances in terms of maximum windings inductances and the rotor position are

$$L_{qq} = L_1 + L_2 \cos(2\theta_r), \tag{3}$$

$$L_{dd} = L_1 - L_2 \cos(2\theta_r), \tag{4}$$

where

$$L_1 = \frac{1}{2}(L_q + L_d),\ L_2 = \frac{1}{2}(L_q - L_d).$$

The mutual inductance between the q and d windings is zero if the rotor is cylindrical and smooth. Because the saliency in the permanent-magnet machines that have magnets placed inside the rotor, the d and q windings fluxes will be linked as the uneven reluctance provides path for the flux through the q-axis winding. When θ_r is zero or 90 electrical degrees the mutual coupling is zero, but is maximum if $\theta_r = -45^o$. Therefore assuming sinusoidal variation, the mutual inductances between the q and d axes windings are given by

$$L_{qd} = L_{dq} = -L_2 \cos(2\theta_r). \tag{5}$$

In PMSM, $L_q > L_d$ always, because the permanent magnets in real machines always have an arc less than 180^o with an interpolar space of iron that provides lower reluctance path for the flux.

Substituting the self-inductances (3) and (4) and the mutual inductances (5) into the equations of the stator voltages (1) and (2) results in

$$\begin{bmatrix} v_{qs} \\ v_{ds} \end{bmatrix} = R_s \begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} + 2\omega_r L_2 \begin{bmatrix} -\sin 2\theta_r & -\cos 2\theta_r \\ -\cos 2\theta_r & \sin 2\theta_r \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} + \begin{bmatrix} L_1 + L_2 \cos 2\theta_r & -L_2 \sin 2\theta_r \\ -L_2 \sin 2\theta_r & L_1 - L_2 \cos 2\theta_r \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} + \lambda_{af} \omega_r \begin{bmatrix} \cos \theta_r \\ -\sin \theta_r \end{bmatrix}.$$

In surface mount magnet machines $L_q = L_d$, hence, $L_2 = 0$. Then, the stator voltages equations for surface mount magnet PMSM are

$$\begin{bmatrix} v_{qs} \\ v_{ds} \end{bmatrix} = R_s \begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} + \begin{bmatrix} L_1 & 0 \\ 0 & L_1 \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} + \lambda_{af} \omega_r \begin{bmatrix} \cos \theta_r \\ -\sin \theta_r \end{bmatrix}$$

The algebraic relation between the flux linkages and the currents is written in compact form as

$$\begin{bmatrix} \lambda_{qs} \\ \lambda_{ds} \end{bmatrix} = \begin{bmatrix} L_{qq} & L_{qd} \\ L_{dq} & L_{dd} \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} + \lambda_{af} \begin{bmatrix} \sin \theta_r \\ \cos \theta_r \end{bmatrix}.$$

In the salient pole PMSMs, the inductances are rotor position dependent. The solution of such equations is cumbersome for it requires a greater computational resource. If these dependencies are eliminated via a transformation then the equations will become manageable in the sense of equivalent circuit and phasor diagram building, and finding the steady-state equations. They are very important for investigation of the machine performance both in steady state and dynamic mode.

2.1 Transformation to rotor reference frames

The relationships between the q and d axes of the stationary reference frames, and the q^r and d^r axes of the rotor reference frames are shown in figure 2.

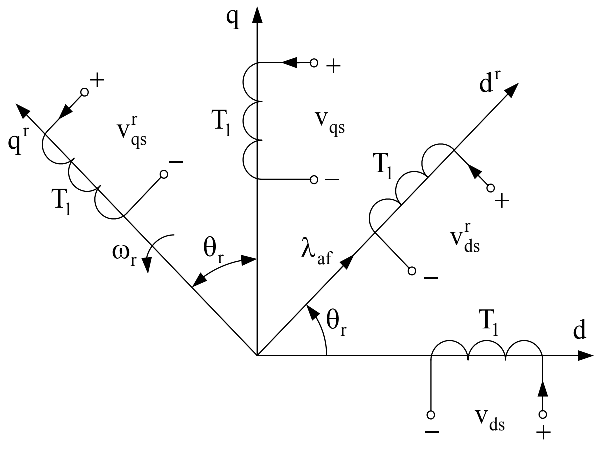

Figure 2. *Stationary and rotor reference frames*

The reference frames rotating at the rotor speed are hereafter referred to as rotor reference frames. The transformation leading to constant inductance is achieved by replacement of the real stator windings with fictitious stator windings which are placed along the q^r and d^r axes. After this operation the real and the fictitious stator windings have equal turns for each phase and should produce equivalent MMF. The actual stator MMF in any axis is the product of the number of turns and the current in the respective axis winding. It is equal to the MMF of the fictitious stator windings. It is established that the actual stator windings MMFs are obtained by projecting these fictitious MMFs on the q and d axes of the actual stator windings. This leads to cancellation of the number of turns from both sides of the q and d axes stator MMF equations, resulting in a relationship between the actual and the fictitious stator currents. The relationship between the currents in the stationary reference frames and the currents in the rotor reference frames is

$$\begin{bmatrix} i_{qs} \\ i_{ds} \end{bmatrix} = \begin{bmatrix} \cos \theta_r & \sin \theta_r \\ -\sin \theta_r & \cos \theta_r \end{bmatrix} \begin{bmatrix} i^r_{qs} \\ i^r_{ds} \end{bmatrix}. \qquad (6)$$

The speed of the rotor reference frames is

$$\dot{\theta}_r = \omega_r ,$$

where $\dot{\theta}_r$ is the time derivative of the electrical rotor angle in rad/s. Likewise the relationship between the voltages is

$$\begin{bmatrix} v_{qs} \\ v_{ds} \end{bmatrix} = \begin{bmatrix} \cos \theta_r & \sin \theta_r \\ -\sin \theta_r & \cos \theta_r \end{bmatrix} \begin{bmatrix} v^r_{qs} \\ v^r_{ds} \end{bmatrix}. \qquad (7)$$

Substituting equations (6) and (7) into the equations of the stator voltages (1) and (2) results in the PMSM model in the rotor reference frames

$$\begin{bmatrix} v^r_{qs} \\ v^r_{ds} \end{bmatrix} = \begin{bmatrix} R_s + L_q \frac{d}{dt} & \omega_r L_d \\ -\omega_r L_q & R_s + L_d \frac{d}{dt} \end{bmatrix} \begin{bmatrix} i^r_{qs} \\ i^r_{ds} \end{bmatrix} + \begin{bmatrix} \omega_r \lambda_{af} \\ 0 \end{bmatrix} \qquad (8)$$

where ω_r is the electrical rotor speed.

2.2 Phase coordinate transformations

Usually three-phase PMSMs are used in practice but the above two-phase model derived is much more convenient. That is why a transformation

$$\mathbf{x}^r_{qd} = \mathbf{T}^r_{abc} \mathbf{x}_{abc} ,$$

where

$$\mathbf{x}^r_{qd} = [x^r_q, x^r_d]^T , \ \mathbf{x}_{abc} = [x_a, x_b, x_c]^T ,$$

$$\mathbf{T}^r_{abc} = \frac{2}{3} \begin{bmatrix} \cos \theta_r & \cos(\theta_r - \frac{2\pi}{3}) & \cos(\theta_r + \frac{2\pi}{3}) \\ \sin \theta_r & \sin(\theta_r - \frac{2\pi}{3}) & \sin(\theta_r + \frac{2\pi}{3}) \end{bmatrix},$$

between the stationary three-phase variables of the real motor and the two phase variables of the model in rotor reference frames is needed to implement the two-phase model. This is the well known Park-Clarke transformation. The transformation of the three-phase stator currents in two-phase rotor reference frames currents is

$$\mathbf{i}^r_{qd} = \mathbf{T}^r_{abc} \mathbf{i}_{abc} ,$$

with

$$\mathbf{i}^r_{qd} = [i^r_q, i^r_d]^T, \quad \mathbf{i}_{abc} = [i_a, i_b, i_c]^T.$$

Likewise the transformation of the three-phase stator voltages is

$$\mathbf{v}^r_{qd} = \mathbf{T}^r_{abc}\mathbf{v}_{abc}, \tag{9}$$

where

$$\mathbf{v}^r_{qd} = [v^r_q, v^r_d]^T, \ \mathbf{v}_{abc} = [v_a, v_b, v_c]^T.$$

The inverse transformation from two-phase rotor reference frames to stationary three-phase reference frames for the stator currents and voltages is

$$\mathbf{i}_{abc} = [\mathbf{T}^r_{abc}]^{-1}\mathbf{i}^r_{qd}, \tag{10}$$

$$\mathbf{v}_{abc} = [\mathbf{T}^r_{abc}]^{-1}\mathbf{v}^r_{qd}. \tag{11}$$

2.3 Power equivalence

The input power of the three-phase motor has to be equal to the input power of the two-phase motor in order to have meaningful interpretation in the modeling, analysis and simulation. Such equality is presented in this section. The three-phase instantaneous input power is

$$p_i = v^T_{abc} i_{abc} = v_{as} i_{as} + v_{bd} i_{bs} + v_{cs} i_{cs} \tag{12}$$

Its two-phase equvalence is derived by substitution of transformations (10) and (11) into equation (12)

$$p_i = 1.5(v^r_{qs} i^r_{qs} + v^r_{ds} i^r_{ds}).$$

2.4 Torques dynamic balance

The electromagnetic torque is the most important output variable that influences the mechanical dynamics of the motor. It is derived from the matrix-vector equation of the machine by considering the input power and its components such as resistive losses, mechanical power, and the rate of change of the stored magnetic energy. In steady state the rate of change of the stored magnetic energy is zero, hence, the output power is the difference between the input power and the resistive losses. This condition is not fulfilled dynamically. The derivation of the electromagnetic torque is based on these peculiarities. The dynamic equations of PMSM in a vector-matrix form is

$$\mathbf{V} = \mathbf{R}\,\mathbf{i} + \mathbf{L}\frac{d\mathbf{i}}{dt} + \mathbf{G}\,\mathbf{i}\,\omega_r, \tag{13}$$

where **R** and **L** are diagonal matrices of resistances and inductances, **G** contains the remaining coefficients connected with the electrical rotor speed ω_r. The instantaneous input power is derived by multiplying equation (13) with the transpose of the current vector

$$p_i = \mathbf{i}^T\mathbf{V} = \mathbf{i}^T\mathbf{R}\,\mathbf{i} + \mathbf{i}^T\mathbf{L}\frac{d\mathbf{i}}{dt} + \mathbf{i}^T\mathbf{G}\,\mathbf{i}\,\omega_r.$$

The term $\mathbf{i}^T\mathbf{R}\,\mathbf{i}$ gives the stator and rotor resistive losses, $\mathbf{i}^T\mathbf{L}\,d\mathbf{i}/dt$ is the rate of change of the stored magnetic energy, and the term $\mathbf{i}^T\mathbf{G}\,\mathbf{i}\,\omega_r$ is the air gap power P_a. It is known that the air gap power is the product of the mechanical rotor speed and the electromagnetic torque. Then

$$\omega_m T_e = P_a = \mathbf{i}^T\mathbf{G}\,\mathbf{i}\,\omega_r = \mathbf{i}^T\mathbf{G}\mathbf{i}P\omega_m.$$

where P is the number of pole pairs and ω_m is the mechanical rotor speed. Canceling the speed and substituting the matrix **G** yields the electromagnetic torque as

$$T_e = 1.5P[\lambda_{af} + (L_d - L_q)i^r_{ds}]i^r_{qs} \tag{14}$$

The equation for the motor dynamics is

$$T_e = T_L + B\omega_m + J\frac{d\omega_m}{dt} \tag{15}$$

where T_L is the load torque, B is the friction coefficient, and J is the moment of inertia.

2.5 Equivalent circuits

The PMSM equivalent circuits of the q and d axes can be derived from the stator voltage equations (8) and are shown in figures 3a and 3b, respectively.

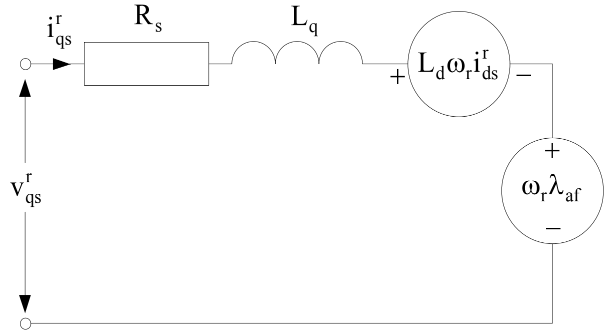

Figure. 3a. *Dynamic stator q-axis equivalent circuit*

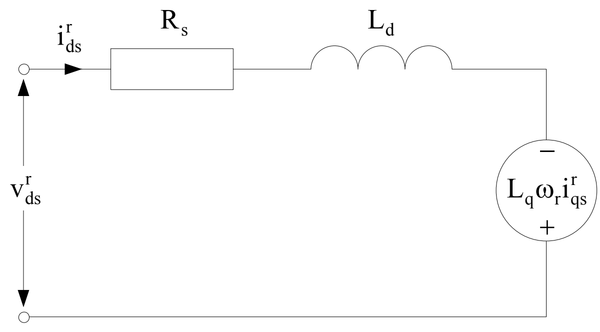

Figure 3b. *Dynamic stator d-axis equivalent circuit*

2.6 State space dynamic models

The PMSM dynamic model according to equations (8) and (15) with the additional equation of the mechanical angle $d\theta_m/dt$ are expressed as follows

$$\frac{d\theta_m}{dt} = \omega_m \tag{16a}$$

$$\frac{d\omega_m}{dt} = \frac{3P[\lambda_{af} + (L_d - L_q) i_{ds}^r] i_{qs}^r - T_l - B\omega_m}{2J} \qquad (16b)$$

$$\frac{di_{qs}^r}{dt} = -\frac{R_s}{L_q} i_{qs}^r - \frac{\omega_r}{L_q}(L_d i_{ds}^r + \lambda_{af}) + \frac{1}{L_q} v_{qs}^r \qquad (16c)$$

$$\frac{di_{ds}^r}{dt} = -\frac{R_s}{L_d} i_{qs}^r + \frac{\omega_r}{L_d} L_q i_{qs}^r + \frac{1}{L_d} v_{qs}^r \qquad (16d)$$

Introducing the state space vector

$$\mathbf{x} = [x_1, x_2, x_3, x_4]^T = [\theta_m, \omega_m, i_{qs}^r, i_{ds}^r]^T,$$

leads to the first state space model defined with respect to the stator currents

$$\begin{aligned} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= c_1 x_3 + c_2 x_3 x_4 - c_3 - c_4 x_2 \\ \dot{x}_3 &= -c_5 x_3 - c_6 x_2 x_4 - c_7 x_2 + c_8 u_1 \\ \dot{x}_4 &= -c_9 x_4 + c_{10} x_2 x_3 + c_{11} u_2 \end{aligned} \qquad (17)$$

with control vector

$$\mathbf{u} = [u_1, u_2]^T = [v_{qs}^r, v_{ds}^r]^T, \qquad (18)$$

and coefficients

$$c_1 = \frac{3P\lambda_{af}}{2J}, c_2 = \frac{3P(L_d - L_q)}{2J}, c_3 = \frac{T_L}{J}, c_4 = \frac{B}{J},$$

$$c_5 = \frac{R}{L_q}, c_6 = \frac{L_d P}{L_q}, c_7 = \frac{\lambda_{af} P}{L_q}, c_8 = \frac{1}{L_q}, c_9 = \frac{R}{L_d},$$

$$c_{10} = \frac{L_q P}{L_d}, c_{11} = \frac{1}{L_d}.$$

The second state space model, defined with respect to the flux linkages is obtained from (16) by considering the relations

$$\lambda_{qs}^r = L_q i_{qs}^r, \ \lambda_{ds}^r = L_d i_{ds}^r + \lambda_{af} \qquad (19)$$

and introducing the new state space vector

$$\mathbf{x} = [x_1, x_2, x_3, x_4]^T = [\theta_m, \omega_m, \lambda_{qs}^r, \lambda_{ds}^r]^T,$$

with the same control vector (18) in the form

$$\begin{aligned} \dot{x}_1 &= x_2 \\ \dot{x}_2 &= c_1 x_3 + c_2 x_3 x_4 - c_3 - c_4 x_2 \\ \dot{x}_3 &= -c_5 x_3 - c_6 x_2 x_4 + u_1 \\ \dot{x}_4 &= c_7 - c_8 x_4 + c_6 x_2 x_3 + u_2 \end{aligned}$$

Its coefficients read

$$c_1 = \frac{3P\rho\lambda_f}{2L_q J}, c_2 = \frac{3P(1-\rho)}{2L_q J}, c_3 = \frac{T_L}{J}, \ c_4 = \frac{B}{J},$$

$$c_5 = \frac{R}{L_q}, c_6 = P, \ c_7 = \frac{R\lambda_f}{L_d}, \ c_8 = \frac{R}{L_d}, \rho = \frac{L_q}{L_d}.$$

3. Simulation results

The first PMSM state space mathematical model (17) is simulated to check its working capacity. The simulation is carried out with zero initial conditions

$$\mathbf{x}_0 = [x_{10}, x_{20}, x_{30}, x_{40}]^T = [0,0,0,0]^T,$$

control vector

$$\mathbf{u} = [u_1, u_2]^T = [v_{qs}^r, v_{ds}^r]^T = \mathbf{T}_{abc}^r \mathbf{v}_{abc},$$

the friction considered in the model via the c_4 coefficient, and load torque $T_L = 10\,\mathrm{Nm}$. Figure 4 shows the transient responses along x_2, x_3, x_4, u_1, and u_2 for the current state space model

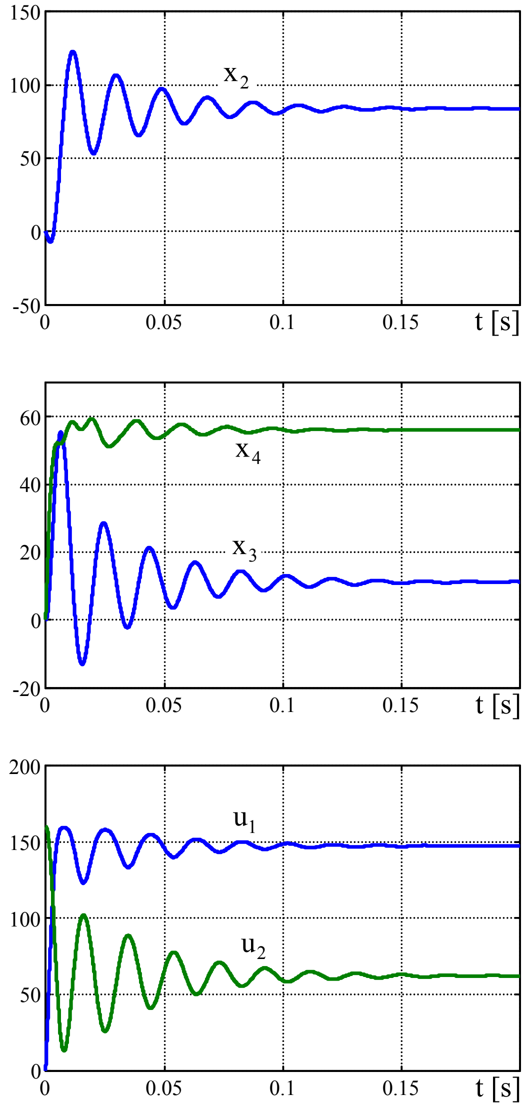

Figure 4. *Dynamic response of current SS model*

while figure 5 displays the electromagnetic torque evolution. The two models differ only in the variables x_3, x_4 being currents in the first model

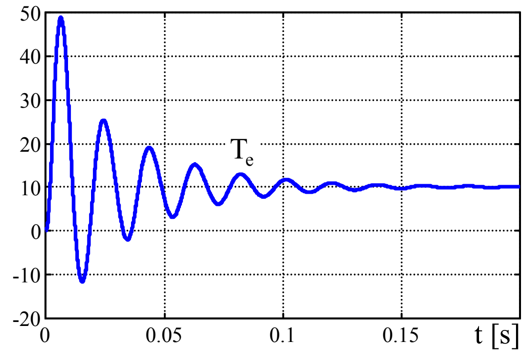

Figure 5. *Electromagnetic torque response*

and flux linkages, displayed on figure 6, for the

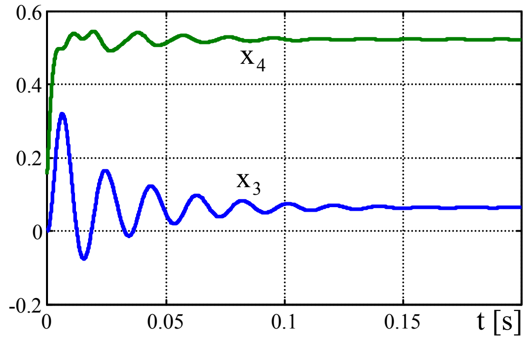

Figure 6. *Flux linkages dynamic response*

second state space model. The PMSM motor parameters used in the simulation are as follows

$$R = 1.4\,\Omega,\ L_d = 6.6\text{ mH},\ L_q = 5.8\text{ mH},$$

$$\lambda_f = 0.1546\text{ Vs},\ P = 3,\ J = 0.00176\text{ kgm}^2,$$

$$B = 0.00038818\text{ Nms}.$$

The two state space models derived are equivalent at considering the relations (19) between the flux linkages and the currents.

4. Conclusions

The paper has presented a two-phase dq mathematical model derivation for PMSM, reflection of the three-phase variables in two-phase dq coordinates by Park-Clarke transformation and application of the equivalent power principle. The model derived is in rotating reference coordinates synchronous with the electrical rotor speed which leads to constant inductances of the stator windings. Equivalent replacement scheme is proposed for the electrical processes taking place in the PMSM. The electromagnetic torque expression is derived as function of the currents and the flux linkages. The two possible state space models are derived based on the two-phase dq PMSM mathematical model. The two introduced strictly orientated state space models are specially designed for the application of advanced control theory differential geometric design approaches including multi-input multi-output feedback linearization and adaptive systems design. These models are simulated for illustrating their working capacity.

Acknowledgement

The support of Research Fund Project No: BY-TH-108/2005 from the Ministry of Education and Science of Bulgaria for this research work is gratefully acknowledged.

REFERENCES

1. Chiasson J., Modeling and High Performance Control of Electric Machines, Wiley-IEEE Press, 2005.

2. Ichikawa S., M. Tomita, S. Doki, S. Okuma, Sensorless control of permanent-magnet synchronous motors using online parameter identification based on system identification theory, IEEE Transactions on Industrial Electronics, Vol. 53, No. 2, 2006, pp. 363-372.

3. Kaewjinda W., M. Konghirun, Vector Control Drive of Permanent Magnet Synchronous Motor Using Resolver Sensor, ECTI Transactions on Electrical Eng., Electronics, and Communication, 5(1), 2007, pp. 134-138.

4. Krishnan R., Permanent Magnet Synchronous and Brushless DC Motor Drives, Taylor and Francis Group, 2010.

5. Krishnan R., Electric Motor Drives: Modeling, Analysis and Control, Prentice Hall, 2001.

6. Merzoug M. S., H. Benalla, Nonlinear Backstepping Control of Permanent Magnet Synchronous Motor (PMSM), International Journal of System Control, 2010, pp. 30-34.

7. Ong Chee-Mun, Dynamic Simulation of Electric Machinery: Using Matlab/Simulink, Prentice Hall, 1998.

8. Pillay P., R. Krishnan, Modeling of Permanent-magnet Motor Drives, IEEE Transactions on Industrial Electronics, 35(4), 1988, pp. 537-541.

9. Pillay P., R. Krishnan, Modeling, Simulation, and Analysis of Permanent-magnet Motor Drives. I. The Permanent-magnet Synchronous Motor Drive, IEEE Transactions on Industrial Applications, 25(2), 1989, pp. 265-273.

10. Solsona J., M. I. Valla, C. Muravchik, Nonlinear Control of a Permanent Magnet Synchronous Motor with Disturbance Torque Estimation, IEEE Transactions on Energy Conversion, pp. 15(2), 2000, 163-168.

Control Systems Department
Technical University–Sofia, Branch Plovdiv
25 Tsanko Diustabanov St.
4000 Plovdiv
BULGARIA
E-mail: r.mishkov@gmail.com
E-mail: v.spasovv@gmail.com