STUDY TO APPLY PERPENDICULAR TRANSFORMATION IN ORDER TO BUILD MATHEMATICAL MODEL FOR AXIAL FLUX PERMANENT-MAGNET MACHINE

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SUMMARY

Axial Flux Permanent-Magnet (AFPM) machine is a multivariable object due to its multivariable mathematical model which is defined by the matrix equations: voltage equation, flux equation, torque equation and motion equation. Especially, the complex inductance matrix in the machine's mathematical description causes difficulties in analyzing to build its mathematical model. This paper proposes a method of applying perpendicular transformation to simplify the machine's model to help the design of controller easier.

Keywords: *Axial Flux Permanent, perpendicular transformation, PID controller.*

INTRODUCTION

Axial Flux Permanent - Magnet motor (AFPM) has many advantages such as: high performance, high ratio of power and size, high power density, long life, small moment of inertia, wide speed range, high ratio of torque and current, less affected by interference, and robust [1-4]. Thus, AFPM motors are used widely in high quality speed variable electrical drive systems such as industrial robots, CNC machines, medical equipment, and flywheels in energy storage systems and AFPM motors have the almost absolute advantages in electric cars. The basic differences between AFPM motor and other motors are that the electromotive force of AFPM motor is trapezoid wave form due to its centralized windings (the electromotive force of other motors are sinusoidal wave form due to distributed windings). Because of trapezoid electromotive force, AFPM motor has characteristics similar to characteristics of DC motor, high power density, high capability of torque generation, and high performance.

When the application such as electrical drive system for grinder that requires very high speed (>10.000 rpm) or liquid Helium pump system which has very low temperature (< 0^{o}C). AFPM motor is used in combination with two radial magnetic bearings arranged at the two ends of the motor sharp. If we want the rotor to rotate, it must be levitated and not in contact with the stator, so the rotor can move axially. To prevent the rotor from translating axially, an axial magnetic bearing must be added. This makes the system structure becomes bulky. Recent studies proposed the models that integrate axial magnetic bearings into the motor's stator windings in order to reduce the overall size of the system [9].

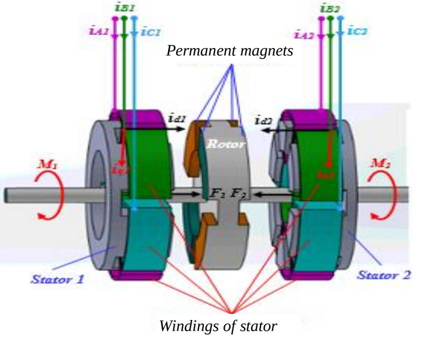

Fig.1: *3D-Drawing of an AFPM motor integrated two radial magnetic bearing*

Fig.1 presents the 3D-drawing of a AFPM motor with integrated radial magnetic bearing at the two ends of the sharp (not described in the figure) [5-7].

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Fig.2 shows the physical model of the motor in **Fig.1**.

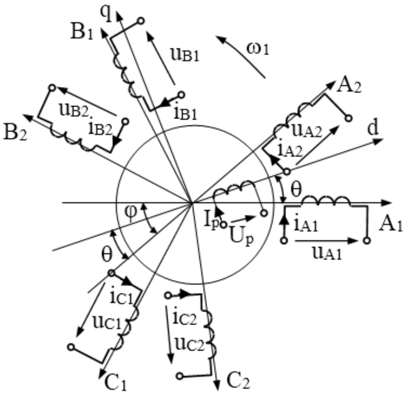

Fig.2: *Physical model of AFPM*

The six-phase system shows in **Fig.2** is split into two three-phase systems motor 1 (M1) and motor 2 (M2) shown in **Fig.3**a and **Fig.3**b.

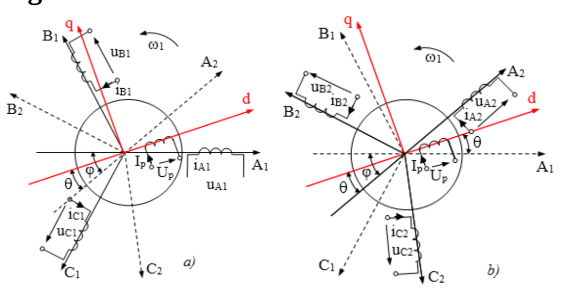

Fig.3: *a) Physical model of M1;*
b) Physical model of M2

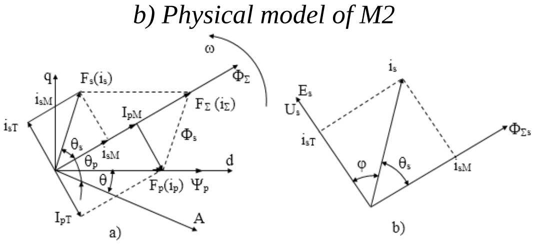

Fig.4: *a) Space vectors of the flux and magnetomotive force;*
b) Phase angle of current, voltage and flux.

The parameters of two motors are shown on the graph of flux vector space and electromotive force as **Fig. 4**. The next contents build the mathematical model to bring simpler model then design and test controllers, evaluate the effectiveness of the orthogonal transformation by simulation.

MATHEMATICAL MODEL OF AFPM AND ORTHOGONAL TRANSFORMATION

The analysis of 3-phase motor is based on orthogonal transformation of the matrices to make AFPM's mathematical model similar to that of DC motor [10,11].

Multi – variable mathematical model of AFPM

Write the motor's voltage equations in matrix form and use the operator p instead of differential notation d/dt:

Voltage balance equation for M1:

$$\begin{vmatrix} u_{A1} \\ u_{B1} \\ u_{C1} \\ U_p \end{vmatrix} = \begin{vmatrix} R_1 & 0 & 0 & 0 \\ 0 & R_1 & 0 & 0 \\ 0 & 0 & R_1 & 0 \\ 0 & 0 & 0 & R_p \end{vmatrix} \times \begin{vmatrix} i_{A1} \\ i_{B1} \\ i_{C1} \\ I_p \end{vmatrix} + p \begin{vmatrix} \psi_{A1} \\ \psi_{B1} \\ \psi_{C1} \\ \Psi_p \end{vmatrix} \qquad (1)$$

Or: $$\boldsymbol{u} = \boldsymbol{Ri} + p\boldsymbol{\psi} \qquad (2)$$

Where: $u_{A1}, u_{B1}, u_{C1}, U_p$, $i_{A1}, i_{B1}, i_{C1}, I_p$, $\psi_{A1}, \psi_{B1}, \psi_{C1}, \Psi_p$: instantaneous values of voltage, current and flux of the phase windings of stator, and rotor, respectively.

Flux equation of M1:

Fluxes of 3-phase stator windings and rotor windings are expressed by matrix equation as follows:

$$\begin{vmatrix} \psi_{A1} \\ \psi_{B1} \\ \psi_{C1} \\ \Psi_p \end{vmatrix} = \begin{vmatrix} L_{A1A1} & L_{A1B1} & L_{A1C1} & L_{A1p} \\ L_{B1A1} & L_{B1B1} & L_{B1C1} & L_{B1p} \\ L_{C1A1} & L_{C1B1} & L_{C1C1} & L_{C1p} \\ L_{pA1} & L_{pB1} & L_{pC1} & L_{pp} \end{vmatrix} \times \begin{vmatrix} i_{A1} \\ i_{B1} \\ i_{C1} \\ I_p \end{vmatrix} \qquad (3)$$

Where: Elements on the principal diagonal are self-inductances of the stator windings and rotor excitation winding, other elements are mutual inductances between windings.

For shortage and simplification, (3) can be rewritten in matrix form:

Where: $\psi_s = |\psi_{A1} \quad \psi_{B1} \quad \psi_{C1}|^T$;

$i_s = |i_{A1} \quad i_{B1} \quad i_{C1}|^T$

$$L_{ss} = \begin{bmatrix} L_{ms} + L_{ts} & -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} \\ -\frac{1}{2}L_{ms} & L_{ms} + L_{ts} & -\frac{1}{2}L_{ms} \\ -\frac{1}{2}L_{ms} & -\frac{1}{2}L_{ms} & L_{ms} + L_{ts} \end{bmatrix} \qquad (4a)$$

$$L_{ps} = L_{sp}^T = $$

$$= L_{ms} \times \begin{bmatrix} \cos\theta & \cos(\theta - 120^0) & \cos(\theta + 120^0) \\ \cos(\theta + 120^0) & \cos\theta & \cos(\theta - 120^0) \\ \cos(\theta - 120^0) & \cos(\theta + 120^0) & \cos\theta \end{bmatrix} \qquad (4b)$$

Or: $\boldsymbol{\psi} = \boldsymbol{Li}$ (5)

Substitute the flux equation into the voltage balance equation, we have:

$$u = Ri + p(Li) = Ri + L\frac{di}{dt} + i\frac{dL}{dt} = Ri + L\frac{di}{dt} + \omega\frac{dL}{d\theta}i \quad (6)$$

Motion equation

In general, the motion equation of the electrical drive system as follows:

$$M_{dt} = M_c + \frac{J}{n_p}\frac{d\omega}{dt} + \frac{D}{n_p}\omega + \frac{K}{n_p}\theta \quad (7)$$

Where: M_c, J, D, K, n_p are torque load, moment of inertia, proportional coefficient of torque load versus angular speed, double of pole. For the constant torque load, then:

$$M_{dt} = M_c + \frac{J}{n_p}\frac{d\omega}{dt} \quad (8)$$

Torque equation

Based on the principle of electromechanical energy conversion, in multi-winding motor, electromagnetic energy is: $W_m = \frac{1}{2}i^T\psi = \frac{1}{2}i^T Li$ (9)

Electromagnetic torque is equal to partial derivative with respect to angular displacement θ_m of the electromagnetic energy in the motor, when the current is constant; there is only one variable that is θ_m, $\theta_m = \theta / n_p$, hence:

$$M_{dt} = \left.\frac{\partial W_m}{\partial \theta_m}\right|_{i=const} = n_p \left.\frac{\partial W_m}{\partial \theta}\right|_{i=const} \quad (10)$$

Substitute the equation (9) into (10), as well as consider the relationship between the expressions (4a) and (4b):

$$M_{dt} = \frac{1}{2}n_p i^T \frac{\partial L}{\partial \theta} i = \frac{1}{2}n_p i^T \begin{bmatrix} 0 & \frac{\partial}{\partial \theta}L_{sK} \\ \frac{\partial}{\partial \theta}L_{Ks} & 0 \end{bmatrix} i \quad (11)$$

$$i^T = \left| i_s^T \quad I_p^T \right| = \left| i_{A1} \quad i_{B1} \quad i_{C1} \quad I_p \right|$$

$$\left.\begin{aligned} & u_1 = Ri_1 + L\frac{di_1}{dt} + \omega\frac{\partial L}{\partial \theta}i_1 \\ & M_{dt1} = \frac{1}{2}n_p i^T \frac{\partial L}{\partial \theta} i_1 = M_c + \frac{J}{n_p}\frac{d\omega}{dt} \\ & \omega = \frac{d\theta}{dt} \end{aligned}\right\} \quad (12)$$

$$\begin{aligned} & \frac{di_1}{dt} = -L^{-1}(R + \omega\frac{\partial L}{\partial \theta})i_1 + L^{-1}u_1 \\ & \frac{d\omega}{dt} = \frac{n_0^T}{2J}i^T\frac{\partial L}{\partial \theta}i_1 - \frac{n_p}{J}M_c \\ & \frac{d\theta}{dt} = \omega \end{aligned} \quad (13)$$

Mathematical model of 3-phase synchronous motor

The combination of (6), (8) and (11) gives multivariable mathematical model of 3-phase synchronous motor when the motor is under constant torque load:

The set of equations (12) can be written in formal form of nonlinear state equations (13): Because the mathematical model has a complex matrix of inductances, it is difficult to use for analysis. For the convenience, coordinator transformations are usually used to simplify the model. The mathematical model of Motor no.2 is similar to Motor no.1 but notice that the index "1" should be replaced by index "2".

Orthogonal transformations and DC motor equivalent model

Orthogonal transformations

To simplify the model, we have to simplify the flux equation firstly. If the physical model of synchronous motor (**Fig.5a**) can be converted into equivalent model of DC motor (**Fig.5b**), after that apply control methods for DC motor then the problem becomes much more simple.

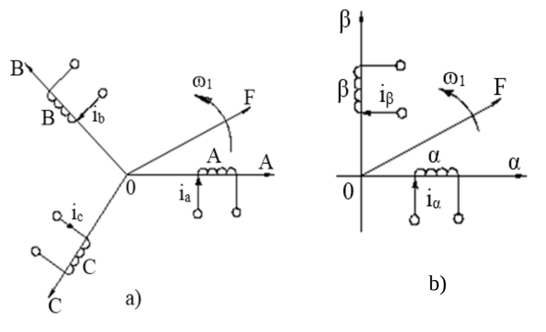

Fig.5: *a) Physical model of three-phase AC windings; b) Equivalent two-phase AC windings model.*

If we set the same dynamic magnet generated as reference, the system of 3 three-phase AC conductors in Fig.5a and the system of 2 crossed conductors in Fig.5b are equivalent. In other hand i_A, i_B, i_C in three-phase coordinate and i_α, i_β in two-phase coordinate are equivalent, they can both generate a same rotational magnetomotive forces.

Assuming that u and i are voltage and current vectors in a system of coordinators,

$$u = \begin{bmatrix} u_A \\ u_B \\ u_C \end{bmatrix}; \; i = \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} \qquad (14)$$

u' and *i'* are voltage and current vectors in a new one:

$$u' = \begin{bmatrix} u'_\alpha \\ u'_\beta \end{bmatrix}; \quad i' = \begin{bmatrix} i'_\alpha \\ i'_\beta \end{bmatrix} \qquad (15)$$

The coordinator transformation is defined as follows:

$$u = A_u u' \qquad (16a)$$

và: $$i = A_i i' \qquad (16b)$$

Where: A_u, A_i are transformation matrixes of real numbers. Assuming that the power is invariable, then:

$$P = u_A i_A + u_B i_B + u_C i_C = u^T i \; = u'_\alpha i'_\alpha + u'_\beta i'_\beta = u'^T i'$$

(17) Substitute (16a), (16b) into (17):

$$i^T u = (A_i i')^T A_u u' = i'^T A_i^T A_u u' = i'^T u'$$

Yields: $$A_i^T A_u = I \qquad (18)$$

I is identity matrix. The expression (18) is relationship between the transformation matrices under the condition of invariable power. In general, for simplification:

$$A_i = A_u = A$$

Then (8) becomes: $A^T A = I$ or:

$$A^T = A^{-1} \qquad (19)$$

When the condition (19) is satisfied, the matrix A is called the orthogonal matrix.

From that, we can find out the transformation matrix for current as the expression (20) and (21). In fact, they are also the transformation matrix for voltage and flux:

$$C_{3/2} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \qquad (20)$$

Contrary: $$C_{2/3} = C_{3/2}^{-1} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 0 & \frac{1}{2} \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} & \frac{1}{\sqrt{2}} \end{bmatrix} \qquad (21)$$

According to the **Fig.3,** the motor M1 is fed by the system A_1, B_1, C_1 and the motor M2 is fed by the system A_2, B_2, C_2. Both systems are transformed into two 2-phase system $\alpha_1\beta_1$ and $\alpha_2\beta_2$. Performing vectors addition we have a system $\alpha_\Sigma \beta_\Sigma$ shown in **Fig.6**.

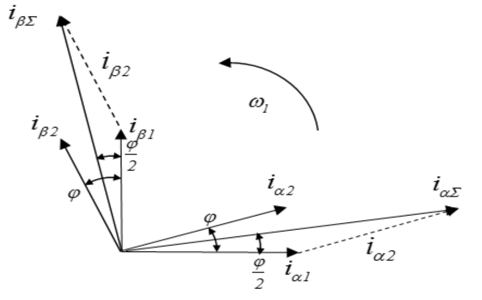

Fig.6: *Vector graph of 2-phase current of AFPM*

Next, assuming φ is the angle between d axis and α axis and applies the 2-phase/2-phase transformation $C_{2r/2s}$. From which we can deduce:

$$\begin{bmatrix} i_d \\ i_q \\ i_0 \end{bmatrix} = \begin{bmatrix} \cos\frac{\varphi}{2} & \sin\frac{\varphi}{2} & 0 \\ -\sin\frac{\varphi}{2} & \cos\frac{\varphi}{2} & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \\ i_0 \end{bmatrix} \qquad (22)$$

$$\begin{bmatrix} i_\alpha \\ i_\beta \\ i_0 \end{bmatrix} = C_{3/2} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} i_A \\ i_B \\ i_C \end{bmatrix} \qquad (23)$$

From (20), can be rewritten as follows to (23): Combining two above expressions, it is possible to obtain the transmitted matrix from 3-phase coordinate ABC to 2-phase rotational coordinate dq_0 as

$$C_{3s/2r} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos\frac{\varphi}{2} & \cos\left(\frac{\varphi}{2} - 120^0\right) & \cos\left(\frac{\varphi}{2} + 120^0\right) \\ -\sin\frac{\varphi}{2} & -\sin\left(\frac{\varphi}{2} - 120^0\right) & -\sin\left(\frac{\varphi}{2} + 120^0\right) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \qquad (24)$$

Inverse transformation matrix:

$$C_{2r/3s} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos\frac{\varphi}{2} & -\sin\frac{\varphi}{2} & \frac{1}{\sqrt{2}} \\ \cos\left(\frac{\varphi}{2}-120^0\right) & -\sin\left(\frac{\varphi}{2}-120^0\right) & \frac{1}{\sqrt{2}} \\ \cos\left(\frac{\varphi}{2}+120^0\right) & -\sin\left(\frac{\varphi}{2}+120^0\right) & \frac{1}{\sqrt{2}} \end{bmatrix} \quad (25)$$

The formulas (24) and (25) are also used for voltage and flux transformation.

Mathematical model of AFPM in dq coordinate

Equivalent mathematical model of AFPM motor in 3-phase synchronous motor in the synchronous rotating rotor field oriented dq [1, 5, 6, 7, 8, 13] as follows:

$$\begin{cases} \begin{vmatrix} u_d \\ u_q \\ U_p \end{vmatrix} = \begin{vmatrix} R_1 + L_s p & -\omega L_q & L_{mp} \\ \omega L_d & R_1 + L_s p & \omega L_p \\ L_{mp} & 0 & R_p + L_p p \end{vmatrix} . \begin{vmatrix} i_d \\ i_q \\ I_p \end{vmatrix} \\ M_{dt} = n_p \frac{L_m}{L_p} \psi_p i_q \end{cases} \quad (26)$$

This relationship is relatively simple and similar to torque equation of DC motor.

State equation of AFPM motor is:

$$\begin{cases} \dfrac{di_{sd}}{dt} = -\dfrac{1}{T_{sd}} i_{sd} + \omega_s \dfrac{L_{sq}}{L_{sd}} i_{sq} + \dfrac{1}{L_{sd}} u_{sd} \\ \dfrac{di_{sq}}{dt} = -\omega_s \dfrac{L_{sd}}{L_{sq}} i_{sd} - \dfrac{1}{T_{sq}} i_{sq} + \dfrac{1}{L_{sq}} u_{sq} - \dfrac{\psi_p}{L_{sq}} \omega_s \end{cases} \quad (27)$$

The mathematical model (27) is represented in **Fig.7**.

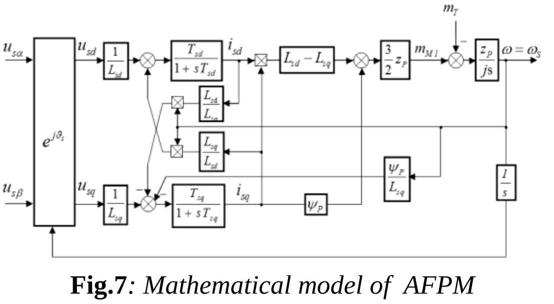

Fig.7: *Mathematical model of AFPM*

CONTROL DESIGN FOR AFPM [8,13]

After using the orthogonal matrices and transformations the mathematical model of AFPM motor shown in **Fig.1** becomes equivalent mathematical model of a motor that has one stator and one rotor. This has two advantages:

- Easy to design control of current loop and speed loop, eliminate interactions between the two motors through controlling current components $i_{d1}, i_{q1}, i_{d2}, i_{q2}$ (in **Fig.3**).
- Axial attractive forces F_1, F_2 become internal problem or the motor that we do not need to care.

In special case, when two 3-phase voltage systems of two inverters provide to M1 and M2 having the same frequency and phase, we have a corresponding motor with double-time moment. In general case, we consider that the amplifiers are equally and phase shift φ.

Base on mathematical model as **Fig.7**, we design Deadbeat controller to control the current for AFPM and PID controller for voltage control. The simulated structure of system is described in **Fig.8**. The simulation results presents the speed characteristic as showed in **Fig.9** and moment characteristic showed in **Fig.10**.

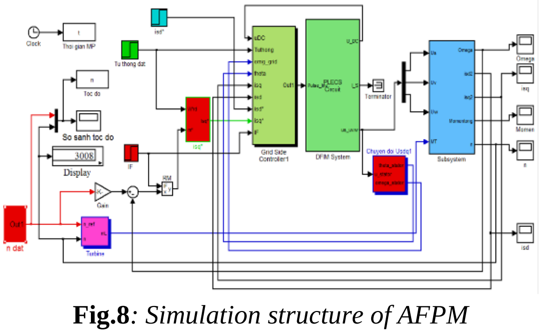

Fig.8: *Simulation structure of AFPM*

AFPM parameters

Rated power $P_{đm}$	350 W
Rated voltage $U_{đm}$	400 V
Rated frequency $U_{đm}$	20KHz
Number of pole pairs n_p	1
Residual flux density	1,45T
Stator resistance R_s	2,3Ω
Stator self-inductance L_s	$11,3.10^{-3}$H
Stator leakage inductance L_{sl}	5.10^{-3}H
Rotor self-inductance L_f	$11,3.10^{-3}$H
Basic direct-axis inductance L_{sd}	$8,2.10^{-3}$H
Basic quadrature-axis inductance L_{sq}	$9,6.10^{-3}$H
Mutual inductance L_m	$9,43.10^{-3}$H
Rotor inertia moment J_r	$8,6.10^{-6}$H
Rotor flux (permanent magnet) Ψ_p	0,0126Wb
Air gap between rotor and stator at the equilibrium position z_0	$1,75.10^{-3}$kgm^2

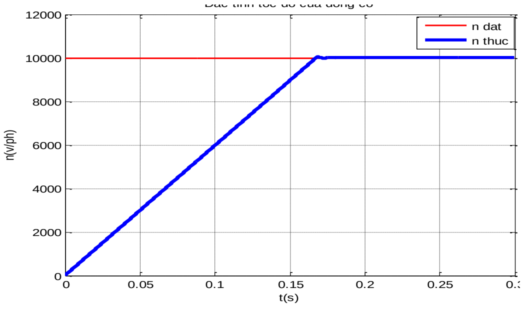

Fig.9: *Speed characteristic*

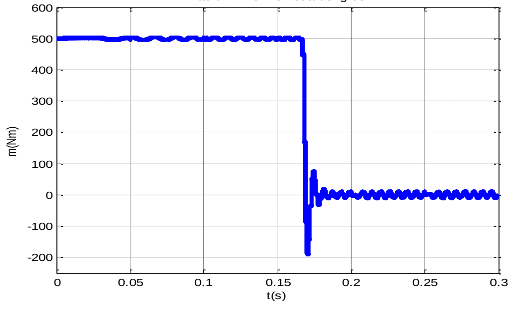

Fig.10: *Moment characteristic*

Reviews: The simulation results proved orthogonal transformation, which has been applied for the AFPM that is accurate. Comparing the simulation results with the study [9] that do not perform alternative equivalent is absolutely identical.

CONCLUSION

AFPM motor with integrated radial magnetic bearing at the two ends of the sharp, in working process its rotor does not only rotate but also moves axially. To prevent the rotor from moving axially and keep the motor compact, an axial magnetic bearing is integrated.

By using the orthogonal transformations to transform coordinator systems the physical model of the motor as shown in **Fig.3** is changed into that of an equivalent AFPM motor. That makes the control design for the motor simple and easy because the motor has only one degree of freedom as conventional motors.

In this paper, the mathematical model of the motor has been built, the suitable controller is chosen and the simulation gives good results.

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TÓM TẮT

NGHIÊN CỨU ỨNG DỤNG PHÉP BIẾN ĐỔI TRỰC GIAO ĐỂ XÂY DỰNG MÔ HÌNH TOÁN HỌC CHO AXIAL FLUX PERMANENT-MAGNET MACHINE

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AFPM - Động cơ đồng bộ từ thông dọc trục kích từ nam châm vĩnh cửu là đối tượng đa biến, do mô hình toán học nhiều biến số của nó được hình thành bởi các phương trình: ma trận điện áp, ma trận từ thông, mô men và chuyển động. Đặc biệt, trong mô tả toán học động cơ loại này có ma trận điện cảm tương đối phức tạp, khó sử dụng để phân tích xây dựng mô hình toán học. Bài báo đưa ra phương pháp nghiên cứu ứng dụng phép biến đổi trực giao để thay đổi mô hình, nhằm có được mô hình toán học thuận lợi cho thiết kế điều khiển động cơ.

Từ khóa: *Động cơ đồng bộ từ thông dọc trục, biến đổi trực giao, bộ điều khiển PID*

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