Digital Vector Control of a Six-Phase Series-Connected Two-Motor Drive.

R. N. Andriamalala, H. Razik, L. Baghli
Université Henri Poincaré - Nancy 1
GREEN-UHP - UMR-7037, BP 239
F - 54506 Vandœuvre-lès-Nancy, Cedex, France
E-mail: [Rijaniaina.Andriamalala, Hubert.Razik, Lofti.Baghli]@green.uhp-nancy.fr

F.M. Sargos
GREEN-ENSEM - UMR-7037
2 avenue de la Foret de Haye
F - 54516 Vandœuvre-lès-Nancy, Cedex, France
E-mail: Francois.Sargos@ensem.inpl-nancy.fr

***Abstract*—This paper deals with a decoupled vector control a six-phase series-connected two-motor drive using the RST controller. This kind of controller is a various of digital controller. The first step is the choice of a reference model of the current and speed loops. Then, the three polynomials R, S, T constituting the controllers will be synthesized by utilizing the Diophante equation after introducing some specifications on each polynomial. Experimental results show the effectiveness of the study.**

I. Introduction

For many industry applications, there are space limitations like in a railway traction drive. Then, researchers and engineers have suggested solutions such as using the mono-converter multi-machine structures [1]–[4]. The motors can be connected in parallel or in series. Many years ago, series connections of DC machines have often been released in railway traction systems [3]. For instance, the parallel connection between induction motors was utilized in Japan for the Shinkansen traction system [4]. The mono-converter multi-machine systems are also an economic solution because the number of the required power electronic devices is reduced [5], [6]. In reference [7], the grid is made in series with the six-phase induction machine and the inverter. This system simultaneously allows to control the six-phase induction motor and to optimize the power factor with consideration of the fault tolerance. The mono-converter multi-machine system is an extension of the multi-phasing operation which also enables to realize the fault tolerance aspect and to improve the power quality [8].

Several years ago, people have improved the investigation of the concept of the decoupled control of series-connected multi-phase motors using a single inverter system [2], [9]. Further studies were performed according to the phase numbers of the motors distinguishing if it is even or odd [10], [11].

Nowadays, there are various kinds of advanced digital controllers applications upon electric-power systems need studies and anlysis. Among them, one proposes the RST controller [12], [13] that will be used to perform the vector control of a six-phase series-connected two-motor drive. This kind of controller allows to get more degree of freedom on the choice of the dynamic of the system. On the next paragraph, one will review the decoupled-Park models of the both motors fed by inverters [2] (the first motor is a six-phase induction machine while the second is a three-phase induction machine).

After that, one will show how to perform the calculation of each controller. The first step is the design of a reference model in closed loop. This one allows to synthesize the controller by solving the Diophante equation after the determination the degrees of three polynomials R, S and T. Experimental tests are developed to show effectiveness of the theoretical study.

II. Modeling

The electrical equations of the system are:

$$\left\{\begin{array}{l} \mathbf{v} = \left(\mathbf{R_{s_1}} + \mathbf{R'_{s_2}}\right)\mathbf{i} + \frac{d\left[\left(\mathbf{L_{s_1}}+\mathbf{L'_{s_2}}\right)\mathbf{i}\right]}{dt} \\ + \frac{d}{dt}\left(\mathbf{L_{sr_1}}\mathbf{i_{r_1}}\right) + \frac{d}{dt}\left(\mathbf{L'_{sr_2}}\mathbf{i_{r_2}}\right) \\ \mathbf{0} = \mathbf{R_{r_1}}\mathbf{i_{r_1}} + \frac{d}{dt}\left(\mathbf{L_{rs_1}}\mathbf{i}\right) + \frac{d}{dt}\left(\mathbf{L_{r_1}}\mathbf{i_{r_1}}\right) \\ \mathbf{0} = \mathbf{R_{r_2}}\mathbf{i_{r_2}} + \frac{d}{dt}\left(\mathbf{L'_{rs_2}}\mathbf{i}\right) + \frac{d}{dt}\left(\mathbf{L_{r_2}}\mathbf{i_{r_2}}\right) \end{array}\right. \quad (1)$$

where $\mathbf{i}$, $\mathbf{i_{r_1}}$, $\mathbf{i_{r_2}}$ and $\mathbf{v}$ designate the vector of the inverter currents, the vector of the rotor currents of the first motor, the vector of the rotor currents of the second motor, the vector of the of the inverter voltage, respectively. R_{s_1}, L_{s_1}, L_{sr_1}, R_{r_1}, L_{r_1} are the stator resistance matrix, the stator inductance matrix, the stator to rotor mutual inductance matrix, the rotor resistance matrix, the rotor inductance matrix of the first motor, respectively. R_{r_2}, L_{r_2} indicate, the rotor resistance matrix, the rotor inductance matrix of the second motor and the matrix R'_{s_2}, L'_{s_2}, L'_{sr_2}, respectively. These matrix are defined as:

$$R'_{s_2} = \begin{bmatrix} R_{s_2} & R_{s_2} \\ R_{s_2} & R_{s_2} \end{bmatrix} \quad (2)$$

$$L'_{s_2} = \begin{bmatrix} L_{s_2} & L_{s_2} \\ L_{s_2} & L_{s_2} \end{bmatrix} \quad (3)$$

$$L'_{sr_2} = \begin{bmatrix} L_{s_2} \\ L_{s_2} \end{bmatrix} \quad (4)$$

$$L'_{rs_2} = \begin{bmatrix} L^T_{sr_2} & L^T_{sr_2} \end{bmatrix} \quad (5)$$

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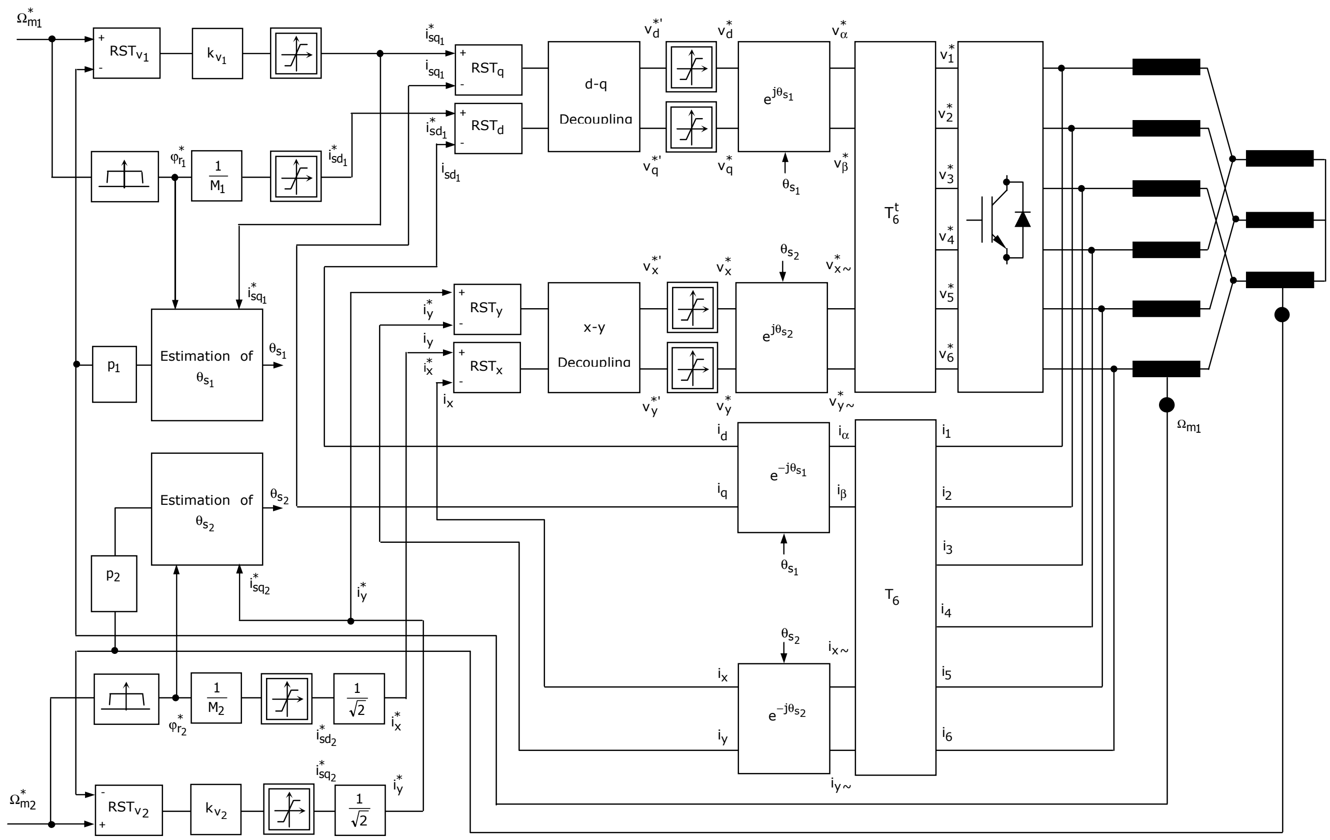

Fig. 1. Diagram of the control scheme

where R_{s_2}, L_{s_2}, L_{sr_2} respectively represent the stator resistance matrix, the stator inductance matrix, the stator to rotor mutual inductance matrix of the second motor.

After applying the six-phase power-invariant transformation C_6 to the inverter currents and voltages and the three-phase Concordia transformation C_3 to the rotor currents of the two motors, Jones *et al* [2] demonstrated a decoupled of the two-motor drive system through the $\alpha - \beta$ and $x_\sim - y_\sim$ frames. The index $\sim$ indicates alternative quantities. This decoupling physically indicates that rotors of both motors could rotate with two different speeds. Consequently, the speeds of the rotating fields in both machines are independent. One omits homopolar components and applies the rotational transformations $R(\theta_{s1})$ and $R(\theta_{s2})$ to the $d-q$ and $x-y$ frames such that:

$$\theta_{s1} = \int \omega_1 dt, \qquad \theta_{s2} = \int \omega_2 dt \tag{6}$$

and:

$$R(\theta) = \begin{bmatrix} \cos\theta & -\sin\theta \\ \sin\theta & \cos\theta \end{bmatrix} \tag{7}$$

where ω_1 and ω_2 designate the speeds of the two rotating fields in both motors, respectively. One gets the Park models of the two motors in the synchronous reference frames:

$$\begin{cases} v_d = R_{s_1} i_d + \frac{d\psi_d}{dt} - \omega_1 \psi_q \\ v_q = R_{s_1} i_q + \frac{d\psi_q}{dt} + \omega_1 \psi_d \\ 0 = R_{r_1} i_{rd_1} + \frac{d\psi_{rd_1}}{dt} - (\omega_1 - p_1 \Omega_{m_1}) \psi_{rq_1} \\ 0 = R_{r_1} i_{rq_1} + \frac{d\psi_{rq_1}}{dt} + (\omega_1 - p_1 \Omega_{m_1}) \psi_{rd_1} \end{cases} \tag{8}$$

with:

$$\begin{cases} \varphi_{dq} = L_{s_1} i_{dq} + M_1 i_{rdq_1} \\ \varphi_{rdq_1} = L_{r_1} i_{rdq_1} + M_1 i_{dq} \end{cases} \tag{9}$$

where L_{s_1}, M_1, L_{r_1}, p_1, Ω_{m_1} respectively represent the stator cyclic inductance, the magnetizing inductance, the rotor cyclic inductance, the pole number and the mechanical speed of the six phase motor. And:

$$\begin{cases} v_x = \left(R_{s_1} + 2R_{s_2}\right) i_x + \frac{d\psi_x}{dt} - \omega_2 \psi_y \\ v_y = \left(R_{s_1} + 2R_{s_2}\right) i_y + \frac{d\psi_y}{dt} + \omega_2 \psi_x \\ 0 = R_{r_2} i_{rd_2} + \frac{d\psi_{rd_2}}{dt} - (\omega_2 - p_2 \Omega_{m_2}) \psi_{rq_2} \\ 0 = R_{r_2} i_{rq_2} + \frac{d\psi_{rq_2}}{dt} + (\omega_2 - p_2 \Omega_{m_2}) \psi_{rd_2} \end{cases} \tag{10}$$

with:

$$\begin{cases} \varphi_{xy} = \left(L_{sl_1} + 2L_{s_2}\right) i_{xy} + \sqrt{2} M_2 i_{rdq_2} \\ \varphi_{rdq_2} = L_{r_2} i_{rdq_2} + \sqrt{2} M_2 i_{ixy} \end{cases} \tag{11}$$

where L_{s_2}, M_2, L_{r_2}, p_2, Ω_{m_2} represent the stator cyclic inductance, the magnetizing inductance, the rotor cyclic inductance, the pole number and the mechanical speed of the second machine, respectively. Electromagnetic torques of both motors are:

$$\begin{cases} \Gamma_{em_1} = p_1 \frac{M_1}{Lr_1} \left(i_q \psi_{rd_1} - \psi_{rq_1} i_d\right) \\ \Gamma_{em_2} = p_2 \frac{\sqrt{2} M_2}{Lr_2} \left(i_y \psi_{rd_2} - \psi_{rq_2} i_x\right) \end{cases} \tag{12}$$

These two subsystem show that its possible to apply the indirect rotor field orientation control (IRFOC) method upon the two motors through respectively the $d-q$ and $x-y$ axis. This IRFOC of the two motor mean:

$$\begin{cases} \psi_{rd_1} = \varphi_{r_1}, & \psi_{rq_1} = 0 \\ \psi_{rd_2} = \varphi_{r_2}, & \psi_{rq_2} = 0 \end{cases} \tag{13}$$

This conditions involve the following relations between reference quantities of the two motor:

$$\begin{cases} v^*_d = R_{s_1} i^*_d + \sigma_1 L_{s_1} \frac{di^*_d}{dt} + \frac{M_1}{L_{r_1}} \frac{d\psi^*_{r_1}}{dt} - \sigma_1 L_{s_1} \omega_1 i^*_q \\ v^*_q = R_{s_1} i^*_q + \sigma_1 L_{s_1} \frac{di^*_q}{dt} + \omega_1 \frac{M_1}{L_{r_1}} \psi^*_{r_1} + \sigma_1 L_{s_1} \omega_1 i^*_d \end{cases} \tag{14}$$

where:

$$\sigma_1 = 1 - \frac{M_1^2}{L_{s_1} L_{r_1}} \tag{15}$$

σ_1 is the leakage factor of the first motor.

$$\left\{\begin{array}{l} v_x^* = \left(R_{s_1} + 2R_{s_2}\right) i_x^* + \sigma_2 \left(L_{sl_1} + 2L_{s_2}\right) \frac{di_x^*}{dt} \\ + \frac{\sqrt{2}M_2}{L_{r_2}} \frac{d\psi_{r_2}^*}{dt} - \sigma_2 \left(L_{sl_1} + 2L_{s_2}\right) \omega_2 i_y^* \\ v_y^* = \left(R_{s_1} + 2R_{s_2}\right) i_y^* + \sigma_2 \left(L_{sl_1} + 2L_{s_2}\right) \frac{di_y^*}{dt} \\ + \omega_2 \frac{\sqrt{2}M_2}{L_{r_2}} \psi_{r_2}^* + \sigma_2 \left(L_{sl_1} + 2L_{s_2}\right) \omega_2 i_x^* \end{array}\right. \quad (16)$$

$$\sigma_2 = 1 - \frac{\left(\sqrt{2}M_2\right)^2}{\left(L_{sl_1} + 2L_{s_2}\right) L_{r_2}} \quad (17)$$

σ_2 is close to the leakage factor of the second motor.

The drive scheme is summarized by the figure 1.

III. Controller designs

A. Current Controllers

Since the $d-q$ and $x-y$ are independent, one only shows how to determine the RST controllers for the currents as for the speed of the six-phase motor. The second motor controllers will be deduced by the same manner.

From the expression (14), we can consider the transfer function of the $d-q$ axis currents of the inverter which also represent the $d-q$ axis currents of the first motor:

$$G_c'(s) = \frac{1}{R_{s_1}} \frac{1}{1 + \sigma_1 \tau_{s_1} s} \quad (18)$$

In practice, one considers that the inverter introduces delay due to the PWM [14]. This one can be considered as equivalent to a time constant τ_d which is approximately equal to $300\mu s$ for our test bench. Usually, τ_d is very small with respect to $\sigma\tau_{s_1}$. Hence, an approximation of the effective continuous transfer function of the $d-q$ currents is expressed as follow:

$$G_c(s) = \frac{1}{R_{s_1}} \frac{1}{1 + \left(\sigma_1 \tau_{s_1} + \tau_d\right) s} \quad (19)$$

Its discrete transfer function is as follows:

$$H_c(z) = \frac{B_c(z)}{A_c(z)} = \frac{b_{0c}}{z + a_{0c}} \quad (20)$$

such that:

$$\left\{\begin{array}{l} a_{0_c} = -e^{-\frac{T_e}{\sigma_1 \tau_s + \tau_d}} \\ b_{0_c} = \frac{1}{R_{s_1}} \left(1 + a_{0_c}\right) \end{array}\right. \quad (21)$$

where, T_e designates the electrical sampling time which is equal to $200\mu s$ for the test bench. Along this study, the index c means a symbol related to the current of the $d-q$ axis. $H_{mc}(z)$ is the reference model of the dynamic of

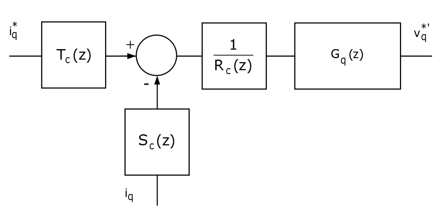

Fig. 2. Example of RST controller for q axis current of the fist motor

the $d-q$ axis currents in closed-loop:

$$H_{mc}(z) = \frac{B_{mc}(z)}{A_{mc}(z)} \quad (22)$$

The index m signifies a symbol related to the reference model. The polynomial $A_{mc}(z)$ is monic [12]. Consider respectively $d^o A_c(z)$, $d^o B_c(z)$, $d^o A_{mc}(z)$, $d^o B_{mc}(z)$ the degrees of the previous polynomials, then, they satisfy to the following condition:

$$d^o A_{m_c} - d^o B_{m_c} \geqslant d^o A_c - d^o B_c \quad (23)$$

To fixe $H_{mc}(z)$, we can consider:

$$d^o A_{mc} = 2 \quad (24)$$

i.e:

$$A_{mc}(z) = z^2 + a_{m_{1c}} z + a_{m_{0c}} \quad (25)$$

The choice of coefficients $a_{m_{1c}}$, $a_{m_{0c}}$ depends of the continuous characteristic polynomial $s^2 + 2\xi_c \omega_{n_c} s + \omega_{n_c}^2$. Usually, the damping ratio $\xi_c \geqslant 0.7$. With a good knowledge of the current dynamic, one can fixe the settling time and deduce the natural frequency. Hence:

$$\left\{\begin{array}{l} a_{m_{0c}} = e^{-2\xi_c \omega_{n_c} T_e} \\ a_{m_{1c}} = -2e^{-\xi_c \omega_{n_c} T_e} \cos\left(\omega_{n_c} \sqrt{1 - \xi_c^2}\right) \end{array}\right. \quad (26)$$

To make the overall gain equal to the unity at $z = 1$:

$$B_{mc}(z) = \frac{A_{mc}(1)}{B_c(1)} B_c(z) = 1 + a_{m_{1c}} + a_{m_{0c}} \quad (27)$$

On the controller shown by the figure 2, it is advantageous to introduce an integral component in the polynomial R(z):

$$R_c(z) = (z-1)^l R_c'(z) \quad (28)$$

l is the integral order. Four our study, we will always take $l = 1$. Since $R(z)$ is monic [12], $R_c'(z)$ is also. The polynomial $T(z)$ is constructed by the following manner:

$$T_c(z) = \frac{A_{mc}(1)}{B_c(1)} A_{0c}(z) \quad (29)$$

where:

$$A_{0c}(z) = z^{d^o A_{0c}} \quad (30)$$

is named the observer polynomial. Its degree can be calculated by the relation:

$$d^o A_{0c} = 2d^o A_c - d^o A_{mc} + l - 1 = 0 \quad (31)$$

So:

$$A_{0c}(z) = 1 \quad (32)$$

and:

$$T_c(z) = \frac{1 + a_{m_{1c}} + a_{m_{0c}}}{b_{0c}} \quad (33)$$

$R_c(z)$ is monic, $R_c'(z)$ is also. Mathematical consideration leads us to find the degrees of $R_c'(z)$ and $S(z)$ by the expressions:

$$\left\{\begin{array}{l} d^o R_c' = d^o A_{mc} + d^o A_{0c} - d^o A_c - l = 0 \\ d^o S_c = d^o A_c + l - 1 = 1 \end{array}\right. \quad (34)$$

Consequently:

$$\begin{cases} R'_c(z) = 1 \Longrightarrow R_c(z) = z - 1 \\ S_c(z) = s_{0c} + s_{1c} z \end{cases} \tag{35}$$

To achieve the determination of $S_c(z)$, one solves the Diophante equation:

$$(z-1)R'_c(z)A_c(z)+B_c(z)S_c(z) = A_{mc}(z)A_{0c}(z) \tag{36}$$

Finally:

$$\begin{cases} s_{0c} = \frac{a_{m_{0c}}+a_{0c}}{b_{0c}} \\ s_{1c} = \frac{1+a_{m_{1c}}-a_{0c}}{b_{0c}} \end{cases} \tag{37}$$

B. Speed Controller

The speed discrete transfer function of the first motor is:

$$H_v(z) = \frac{b_{0v}}{(z + a_{0v})} \tag{38}$$

where:

$$\begin{cases} a_{0v} = -e^{-f_v \frac{T_m}{J}} \\ \\ b_{0v} = \frac{1}{f_v}\left(1 + a_{0v}\right) \end{cases} \tag{39}$$

f_v and J respectively designate the viscous friction coefficient and the inertia. T_m indicates the mechanic sampling time. On this paper, the index v means a symbol related to the speed. Since the q-axis current loop is often very fast than the speed one, one can also fixe a second order reference model $H_{mv}(z)$.

$$H_{mv}(z) = \frac{B_{mv}(z)}{A_{mv}(z)} = \frac{1 + a_{m_{1v}} + a_{m_{0v}}}{z^2 + a_{m_{1v}} z + a_{m_{0v}}} \tag{40}$$

Like in current loop, the choice of $a_{m_{0v}}$ $a_{m_{1v}}$ depends on the damping ratio ξ_v and the natural frequency ω_{n_v} fixed by mean of the continuous second order characteristic polynomial. With the similar steps of calculations, the three polynomials $R_v(z)$, $S_v(z)$, $T_v(z)$ of the speed controller are:

$$\begin{cases} R_v(z) = z - 1 \\ S_v(z) = s_{0v} + s_{1v} z \\ T_v(z) = t_{0v} \end{cases} \tag{41}$$

such that:

$$\begin{cases} s_{0v} = \frac{a_{m_{0v}}+a_{0v}}{b_{0v}} \\ s_{1v} = \frac{1+a_{m_{1v}}-a_{0v}}{b_{0v}} \\ t_{0v} = \frac{1+a_{m_{1v}}+a_{m_{0v}}}{b_{0v}} \end{cases} \tag{42}$$

IV. Experimental Results

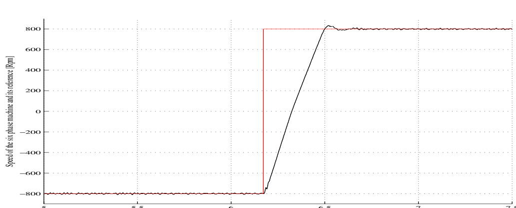

Fig. 3. Speed of the six-phase machine (black) and its reference (red) around the instant of its speed reversal.

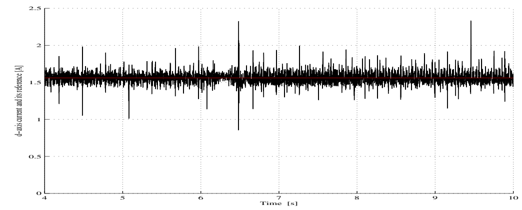

Fig. 4. Axis current i_d (black) and its reference i_d^* (red) around the instant of speed reversal of the six-phase machine.

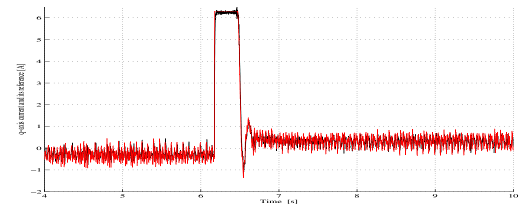

Fig. 5. Axis current i_q (black) and its reference i_q^* (red) around the instant of speed reversal of the six-phase machine.

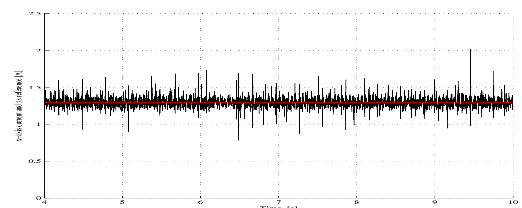

Fig. 6. Axis current i_x (black) and its reference i_x^* (red) around the instant of speed reversal of the six-phase machine.

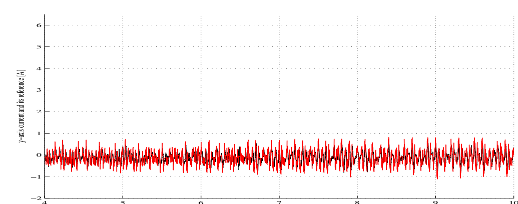

Fig. 7. Axis current i_y (black) and its reference i_y^* (red) around the instant of speed reversal of the six-phase machine.

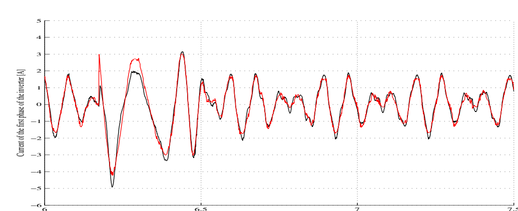

Fig. 8. Inverter current i_1 (black) and its reference i_1^* (red) around the instant of speed reversal of the six-phase machine.

The experimental test bench is composed by two three-phase inverters, a 50Hz, 2-pole, 3kW, 127V six-phase

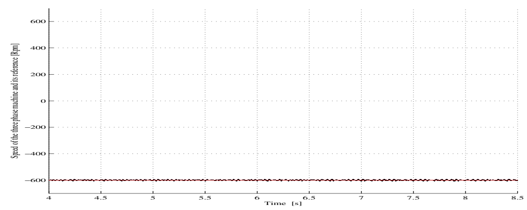

Fig. 9. Speed of the three-phase machine (black) and its reference (red) around the instant of speed reversal of the six-phase machine.

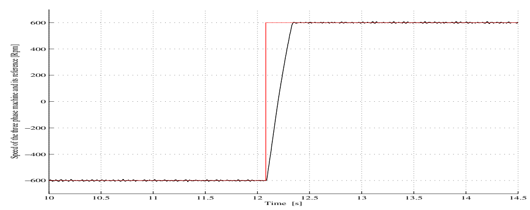

Fig. 10. Speed of the three-phase machine (black) and its reference (red) around the instant of its speed reversal.

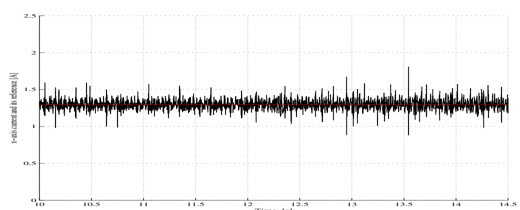

Fig. 11. Axis current i_x (black) and its reference i_x^* (red) around the instant of speed reversal of the three-phase machine.

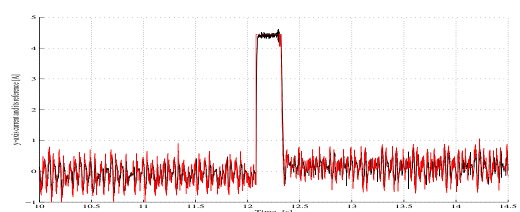

Fig. 12. Axis current i_y (black) and its reference i_y^* (red) around the instant of speed reversal of the three-phase machine.

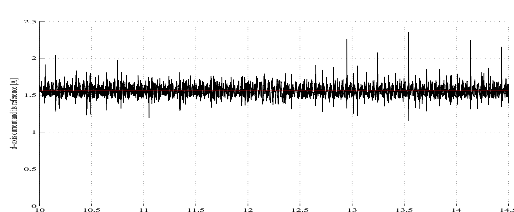

Fig. 13. Axis current i_d (black) and its reference i_d^* (red) around the instant of speed reversal of the three-phase machine.

machine and 50Hz, 2-pole, 3kW, 230V, three-phase one. Both inverters are connected to the same DC-link.

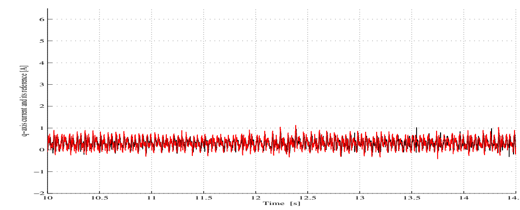

Fig. 14. Axis current i_q (black) and its reference i_q^* (red) around the instant of speed reversal of the three-phase machine.

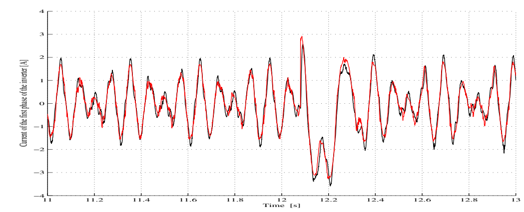

Fig. 15. Inverter current i_1 (black) and its reference i_1^* (red) around the instant of speed reversal of the three-phase machine.

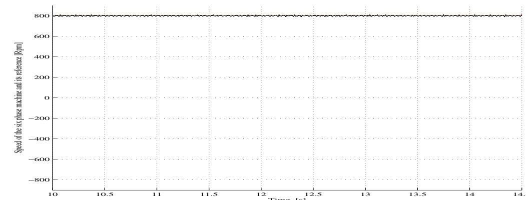

Fig. 16. Speed of the six-phase machine (black) and its reference (red) of speed reversal of the three-phase machine.

At the beginning, the six-phase induction machine (the first motor) is running at $-800rpm$ while the three phase machine (the second one) is rotating at $-600rpm$.

One fixes the direct current references of both motor at $i_{sd_1}^* = 1.56A$ and $i_{sd_2}^* = 1.83A$, respectively.

At $t = 6.17s$, one reverses the speed reference of the first motor $800rpm$. As it can be verified in the figure (3), the speed of first the motor presents a good tracking. The figures (4) and (5) indicate the currents i_d and i_q of voltage inverters and their references i_d^* and i_q^*. They represent the direct and quadrature currents of the six-phase machine. Through these figures, one notes a complete decoupling between both axis currents during the transient operation. It is also observed in the figure (5) a good response of the quadrature current with respect to its reference after a quick perturbation. The current controllers achieve a good tracking. The figures (6), (7) indicates the currents i_x and i_y of inverters. They are the image of the direct and quadrature currents of the second motor ($i_{d_2} = i_x\sqrt{2}$ and $i_{q_2} = i_y\sqrt{2}$). According to the figures (6) and (7), the controllers of the $x-y$ axis currents present correct tracking. They also show that there is any perturbation in the axis currents of the second motor during speed reversal of the first machine. Therefore both motors are electrically decoupled. The first phase current of inverters i_1 and its reference i_1^* are depicted in the figure (8) which confirms again simultaneously good reactions of current controllers after little disturbance due to the same transient operation in q-axis current. As one imposes two different speed references and both machine do not have exactly the same parameters, one observes that the current has two fundamental frequencies.

The figure (9) depicts the speed of the three-phase machine and its reference around the same transient operation. It informs that its speed controller achieves a good tracking to the reference $-600rpm$. It also confirms that the mechanic transient of the first motor does not affect those of the second one.

Now the steady state operation is established for the system. The six-phase machine is running at $800rpm$ while the three-phase one is running at $-600rpm$. At $t = 12.01$, one sets a test of reversal speed of the second machine to $600rpm$. According to the figure (10), the RST controller reacts well to perform a correct tracking for the speed of the three-phase machine. Figures (11) and (12) illustrate the x and y-axis currents of inverters. As it is observed, there is perfect decoupling between d and q-axis currents of the second motor during this second transient. Figures (13) and (14) sketching the direct and quadrature currents of the first machine reveal that they are not affected by the transient behavior of the second machine. The figure (15) sketches behaviors of the first phase current of inverters i_1 and its reference i_1^*. After little transient caused by the speed reversal of the second motor, one obtains a agreement between them. Figure (16) shows a good tracking of the speed of the first motor to the reference $800rpm$. The transient of the second machine does not disturb the speed of the first one. Controls of both machines are fully decoupled.

V. Conclusion

In this paper, the application of the RST controller on vector control of a six-phase series-connected two-motor drive is treated. To synthesize these controllers, one uses the decoupled-Park models of two motors fed by a six-phase inverter. From these models, one deduces the discrete transfer functions of currents and speeds. The first step is to choose the transfer function of the reference model of each loop. The controllers are calculated after introducing some specifications one each polynomials R, S, T and solving the Diophante equation. Experimental tests show that RST controllers give good and correct tracking in current and speed loops. The fully-decoupled controls of both motors are also confirmed.

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