NONLINEAR FEEDBACK APPROACH BASED ON SLIDING MODE CONTROLLER FOR AN INDUCTION MOTOR FED BY MATRIX CONVERTER

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ABSTRACT

In this paper, a nonlinear feedback linearization approach and sliding mode controller are combined to generate the reference voltages for a matrix converter controlled by DSVM strategy fed induction motor in order to preserve the robustness with respect to desired dynamic behaviors for this drive system. The transfer function and the selection approach of the parameter for the input filter are also introduced. The simulation results of the robustness testing of the drive system have been carried out to validate the advantages of the proposed control system.

MOTS CLES: Direct Space vector modulation (DSVM), induction motor (IM), LC input filter, matrix converter (MC), nonlinear feedback control, sliding mode controller (SMC).

1 INTRODUCTION

The matrix converter (MC) consists of nine bidirectional switches, arranged as three sets of three so that any of the three input phases can be connected to any of the three output lines, as shown in Figure 3 [1, 2, 3]. During recent years, the MC technology has attracted the power electronics industry and the development progress has been accelerated. The main obstacles toward realizing an industrial MC have been overcome and intial steps toward developing a commercial product have already been taken [4].

MC introduced in [2, 3, 5], have been found as interesting alternative to standard ac/dc/ac converters. The MC, has a number of attractive features such as: power converter adjustment capability and it does not involve a dc voltage link and the associated large capacitor as in the case of ac/dc/ac converter (capacitor between the rectifier and the inverter), and it allows bidirectional power flow. However, the MC presents disadvantages such as: the input grid voltage unbalances can result in unwanted output harmonic currents, which deteriorate the drive system performance, short-time voltage sag could interrupt the normal operation of load equipment, even causing some failures. In the literature, two methods of control are adapted for the control of the MC, such that the Venturini and SVM (direct and indirect) methods, the main advantage of the SVM method lies in lower switching losses compared with the Venturini method. The Venturini method, however, exhibits superior performance in terms of harmonics [3, 5].

In order to solve the filtering problem, the authors in [6, 7, 8] propose various configurations of the input filter adopted by MC. In addition, the transfer function, the characteristics, and the selection approach parameters of any configurations are also presented [7]. The conventional inductive capacitive (LC) filter is well adapted to filtering of harmonics, due to the simplicity of implementation, and satisfactory performance provided by this configuration.

The field-oriented control allows a natural decoupling between the speed and flux amplitude. However, the parameter variations and the high operation speed cause loss decoupling, [9, 10].

The nonlinear control is another alternative approach to achieve an asymptotic decoupling for induction motor while ensuring a perfect linearization some is the profiles of the reference trajectories imposed on the system. Several work [10, 11, 12, 13], showed that this technique revealed interesting properties as for decoupling between electromagnetic torque and flux amplitude in term of the response time of the torque, and with the parametric variations robustness.

Recently, in [14] the nonlinear robust auto disturbance rejection controller (ADRC) is combined by the vector control to generate the reference voltage to the MC controlled by indirect SVM strategy in order to obtain a good performance of the induction motor drive system. The dynamic performance, the stability and reliability of the proposed control strategy under external disturbances have been verified by simulation results. This study constitutes an initiative for the application of nonlinear approach to control matrix converter fed induction motor drive system.

The essential objective of this paper is to design a nonlinear feedback control approach for MC fed IM drive system. In order to obtain a high dynamic performance, a sliding mode controller is employed with the approach considered previously, taking the place of conventional proportional-integral (PI) controller. This control technique consist to generate the reference voltages for a matrix converter controlled by DSVM strategy fed induction motor in order to preserve the robustness with respect to desired dynamic behaviors for this drive system. The DSVM modulation strategy is employed to regulate the input/output sinusoidal waveforms of MC with unity input power factor operation condition. This approach has the advantage of an immediate comprehension of the switching strategies. In addition, the number of switch commutations within a cycle period can be limited utilizing an opportune commutation sequence [5]. Moreover, the transfer function and the parameter selection approach for the input filter are also introduced.

Finally, the robustness testing of the proposed drive system is verified by simulation.

2 NONLINEAR FEEDBACK CONTROL DESIGN

2.1 IM Model

Assuming linear magnetic circuits, the dynamics of a balanced non-saturated IM in a stator fixed reference frame (α-β) attached to the stator, are given by [9, 10, 11, 15].

$$\begin{cases} \dot{x} = f(x) + g(x)u \\ y = h(x) \end{cases} \tag{1}$$

Where:

$$x = [i_{s\alpha} \; i_{s\beta} \; \varphi_{r\alpha} \; \varphi_{r\beta} \; \omega_m]; \; u = \begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix}$$

$$f(x) = \begin{bmatrix} f_1(x) \\ f_2(x) \\ f_3(x) \\ f_4(x) \\ f_5(x) \end{bmatrix} = \begin{bmatrix} -\gamma i_{s\alpha} + \frac{K}{T_r}\varphi_{r\alpha} + K n_p \omega \varphi_{r\beta} \\ -\gamma i_{s\beta} + \frac{K}{T_r}\varphi_{r\beta} - K n_p \omega \varphi_{r\alpha} \\ \frac{L_m}{T_r} i_{s\alpha} - \frac{1}{T_r}\varphi_{r\alpha} - n_p \omega \varphi_{r\beta} \\ \frac{L_m}{T_r} i_{s\beta} - \frac{1}{T_r}\varphi_{r\beta} + n_p \omega \varphi_{r\alpha} \\ \frac{n_p L_m}{J L_r}(\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) - \frac{1}{J}(f\omega + T_r) \end{bmatrix};$$

$$g = [g_1 \; g_2] = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \end{bmatrix};$$

i, φ, u denote current, flux linkage and voltage; the subscripts s and r stand for stator and rotor; (α,β) denote the components of a vector with respect to a fixed stator reference frame. The positive constants, related to electrical mechanical parameters of the IM (1) are defined as:

$$T_r = \frac{L_r}{R_r}, \sigma = 1 - \frac{L_m}{L_s L_r}, \gamma = \frac{1}{\sigma L_s}\left(R_s + \frac{L_m^2}{L_r T_r}\right), K = \frac{L_m}{\sigma L_s L_r}$$

2.2 Nonlinear feedback control

In order to achieve fast torque response as well as operate in the flux weakened region and maximize the power efficiency for the IM drive, the torque (T_e) and the norm of the rotor flux linkage ($\varphi_{r\alpha}^2 + \varphi_{r\beta}^2$) are assumed to be the system outputs. Hence, on the basis of the input–output feedback linearization technique, the following variables are introduced.

$$h(x) = \begin{bmatrix} h_1(x) \\ h2(x) \end{bmatrix} = \begin{bmatrix} T_e \\ \varphi_r^2 \end{bmatrix} = \begin{bmatrix} \frac{n_p L_m}{L_r}(\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) \\ \varphi_{r\alpha}^2 + \varphi_{r\beta}^2 \end{bmatrix} \tag{2}$$

The following notation is used for the ‘‘Lie’’ derivative of a function $h(x): \Re^n \to \Re$ along a vector

$f(x) = (f_1(x), ..., f_n(x))$, [12, 17]

$$L_f h(x) = \sum_{i=1}^{n} \frac{\partial h(x)}{\partial x_i} f_i(x)$$

$$L_f^i = L_f(L_f^{i-1} h(x)) \tag{3}$$

Using the IM fifth order model (1), the derivatives of the outputs are given by:

- For the first output $y_1 = h_1(x) = T_e$

$$\dot{y}_1 = \dot{h}_1(x) = L_f h_1(x) + L_{g_1} h_1(x) v_{s\alpha} + L_{g_2} h_1(x) v_{s\beta} \quad (4)$$

Using the above notation, one can obtain that :

$$\begin{cases} L_f h_1(x) = \dfrac{n_p L_m}{L_r}[(\gamma + \dfrac{1}{T_r})(\varphi_{r\beta} i_{s\alpha} - \varphi_{r\alpha} i_{s\beta}) \\ - K n_p \omega(\varphi_{r\alpha}^2 + \varphi_{r\beta}^2) - n_p \omega(\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta})] \\ L_{g_1} h_1(x) = \dfrac{\partial h_1(x)}{\partial x_i} g_1 = -\dfrac{n_p L_m}{\sigma L_r L_s}\varphi_{r\beta} \\ L_{g_2} h_1(x) = \dfrac{\partial h_1(x)}{\partial x_i} g_2 = \dfrac{n_p L_m}{\sigma L_r L_s}\varphi_{r\alpha} \end{cases} \quad (5)$$

The relative degree corresponds the first output is: $r_1=1$.

- For the second output $y_2 = h_2(x) = \varphi_r^2$

$$\begin{cases} \dot{y}_2 = \dot{h}_2(x) = L_f h_2(x) \\ \ddot{y}_2 = \ddot{h}_2(x) = L_f^2 h_2(x) + L_{g_1}(L_f h_2(x)) v_{s\alpha} \\ + L_{g_2}(L_f h_2(x)) v_{s\beta} \end{cases} \quad (6)$$

Where :

$$\begin{cases} L_f h_2(x) = \dfrac{2L_m}{T_r}(\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) - \dfrac{2}{T_r}(\varphi_{r\alpha}^2 + \varphi_{r\beta}^2) \\ L_f^2 h_2(x) = \dfrac{-2L_m}{T_r}[(\gamma + \dfrac{3}{T_r})(\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) \\ \quad + n_p \omega(\varphi_{r\beta} i_{s\alpha} - \varphi_{r\alpha} i_{s\beta}) \\ \quad + (\dfrac{2KL_m + 4}{T_r^2})(\varphi_{r\alpha}^2 + \varphi_{r\beta}^2) + \dfrac{2L_m^2}{T_r^2}(i_{r\alpha}^2 + i_{r\beta}^2)] \\ L_{g_1}(L_f h_2(x)) = \dfrac{\partial(L_f h_2(x))}{\partial x_i} g_1 = \dfrac{2L_m}{\sigma L_s T_r}\varphi_{r\alpha} \\ L_{g_2}(L_f h_2(x)) = \dfrac{\partial(L_f h_2(x))}{\partial x_i} g_2 = \dfrac{2L_m}{\sigma L_s T_r}\varphi_{r\beta} \end{cases} \quad (7)$$

The relative degree corresponds the second output is: $r_2=1$. Consequently, the total relative degree $r=r_1+r_2=3$. In this case, the full linearization is not realized. For this reason, the following change of coordinates is indispensable [9]:

$$\begin{cases} z_1 = y_1 \\ z_2 = y_2 \\ z_3 = \dot{z}_2 = L_f h_2(x) \\ z_4 = \arctan\left(\dfrac{\varphi_{r\beta}}{\varphi_{r\alpha}}\right) \\ z_5 = \omega \end{cases} \quad (8)$$

Thus, the derivatives of the outputs are given by:

$$\begin{cases} \dot{z}_1 = v_1 \\ \dot{z}_2 = z_3 \\ \dot{z}_3 = v_2 \\ \dot{z}_4 = n_p \omega + \dfrac{R_r z_1}{n_p z_2} \\ \dot{z}_5 = \dfrac{1}{J}(z_1 - T_r - f z_5) \end{cases} \quad (9)$$

From (9), (8), (6) and (4), the new inputs v_1 and v_2 of system are designed as :

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} L_f h_1 \\ L_f^2 h_2 \end{bmatrix} + D(x)\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix} \quad (10)$$

Where $D(x)$ is the decoupled matrix defined as :

$$D(x) = \begin{bmatrix} L_{g_1} h_1 & L_{g_2} h_1 \\ L_{g_1} L_f h_2 & L_{g_2} L_f h_2 \end{bmatrix} = \begin{bmatrix} -\dfrac{n_p L_m}{\sigma L_r L_s}\varphi_{r\beta} & \dfrac{n_p L_m}{\sigma L_r L_s}\varphi_{r\alpha} \\ \dfrac{2L_m}{\sigma L_s T_r}\varphi_{r\alpha} & \dfrac{2L_m}{\sigma L_s T_r}\varphi_{r\beta} \end{bmatrix} \quad (11)$$

Since $\det(D(x)) \neq 0$, $D(x)$ is nonsingular everywhere in ω. The input-output linearizing feedback for system (9) is given by :

$$\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix} = D^{-1}(x)\begin{bmatrix} v_1 - L_f h_1 \\ v_2 - L_f^2 h_2 \end{bmatrix} \quad (12)$$

In order to track desired smooth reference signals T_{ref} and φ_{rref}^2 for the torque $z_1 = T_e$ and the square of the rotor flux $z_2 = \varphi_r^2$, the input signals v_1 and v_2 in (12) are designed as [9] .

$$\begin{cases} v_1 = k_{11}(z_{1ref} - z) + \dot{z}_{ref} \\ v_2 = k_{21}(z_{2ref} - z_2) + k_{22}(\dot{z}_{2ref} - \dot{z}_2) + \ddot{z}_{2ref} \end{cases} \quad (13)$$

Where k_{11}, k_{21} and k_{22} are positive constant design parameters can be determined by imposing a dynamics on the error ($e_1 = (z_{1ref} - z_1)$, $e_2 = z_{2ref} - z_2$) according to :

$$\begin{cases} k_{11}e_1 + \dot{e}_1 = 0 \\ k_{21}e_1 + k_{22}\dot{e}_1 + \ddot{e}_1 = 0 \end{cases} \tag{14}$$

The above dynamics will be stable if the two polynomials of e_1 and e_2 have their roots on the left side of complex plan [9, 10, 17].

In order to track desired smooth reference signals ω_{ref} for the speed $z_5 = \omega$, a sliding mode controller is designed, taking the place of conventional proportional-integral (PI) controller, as follows:

The slip surface is selected as :

$$S(z_5) = e(z_5) = z_{5ref} - z_5 \tag{15}$$

Its time derivate is expressed by :

$$\dot{S}(z_5) = \dot{z}_{5ref} - \dot{z}_5 \tag{16}$$

Under sliding mode condition ($\dot{S}(z_5) = 0$), and from the model (1), the equivalent component of the reference torque is given by :

$$(z_{1ref})_{eq} = \dot{z}_{5ref} + \frac{1}{J}(fz_{5ref} + T_r) \tag{17}$$

In order to reduce the chattering phenomenon due to the discontinuous nature of the SM controller, a smooth function is used for the nonlinear component.

$$(z_{1ref})_n = k_\omega \frac{S(z_5)}{|S(z_5)| + \varepsilon_\omega} \tag{18}$$

$k_\omega, \varepsilon_\omega$: are positive constants.

The SM controller for the speed is defined by :

$$z_{1ref} = (z_{1ref})_{eq} + (z_{1ref})_n \tag{19}$$

3 DIRECT SPACE VECTOR MODULATION

The DSVM represents the three-phase input currents and output line-to-line voltages as space vectors by $\vec{i}_i$ and $\vec{v}_o$ (figure 1). It is based on the concept of approximating a rotating reference voltage vector with those voltages physically realisable on a MC. For nine bidirectional switches, there are 27 possible switching configurations available in which only 21 are commonly utilized to generate the desired space vectors in the DSVM control method. These 21 configurations (table I) in which may be divided into four groups [4, 5, 6].

The first three groups (±1, ±2, ±3, ±4, ±5, ±6, ±7, ±8, ±9) have two common features; namely, each of them consists of six vectors holding constant angular positions and each of them formulates a six-sextant hexagon as shown in Fig. 1.

The general formulae to calculate the on-time durations, have been given as [16].

$$\begin{cases} \delta_1 = \frac{2}{\sqrt{3}} q \sin\left[\alpha_o - (k_v - 1)\frac{\pi}{3}\right] \sin\left[\frac{\pi}{6} - \left(\alpha_i - (k_i - 1)\frac{\pi}{3}\right)\right] \\ \delta_2 = \frac{2}{\sqrt{3}} q \sin\left[\alpha_o - (k_v - 1)\frac{\pi}{3}\right] \sin\left[\frac{\pi}{6} + \left(\alpha_i - (k_i - 1)\frac{\pi}{3}\right)\right] \\ \delta_3 = \frac{2}{\sqrt{3}} q \sin\left[k_v \frac{\pi}{3} - \alpha_o\right] \sin\left[\frac{\pi}{6} - \left(\alpha_i - (k_i - 1)\frac{\pi}{3}\right)\right] \\ \delta_4 = \frac{2}{\sqrt{3}} q \sin\left[k_v \frac{\pi}{3} - \alpha_o\right] \sin\left[\frac{\pi}{6} + \left(\alpha_i - (k_i - 1)\frac{\pi}{3}\right)\right] \\ \delta_0 = 1 - (\delta_1 + \delta_2 + \delta_3 + \delta_4) \end{cases} \tag{21}$$

Where $q = V_O/V_i$ is the voltage transfer ratio, α_o and α_i are the phase angles of the output voltage and input current vectors, respectively, and k_v, k_i are the output voltage sector and input current vector sector.

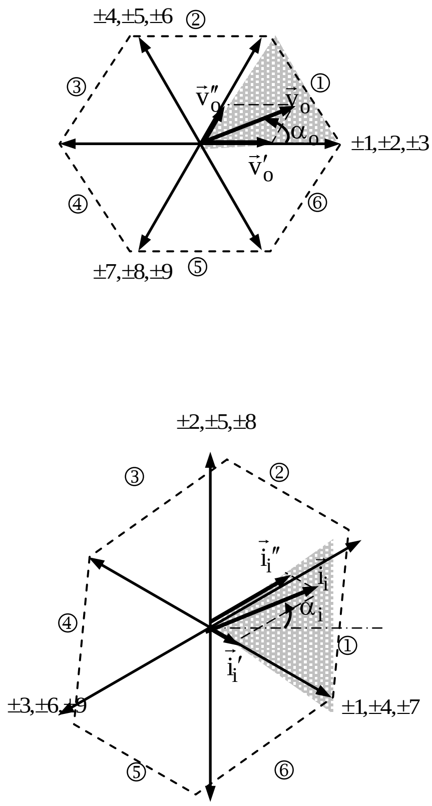

Figure1: Output voltage and input current space vectors

Table 1: Switching Configurations

Swit. Conf.	Switches on	V_o	α_o	I_i	α_i
+1	S_{aA} S_{bB} S_{bC}	$2/3V_{ab}$	0	$2/\sqrt{3}\, i_A$	$-\pi/6$
-1	S_{bA} S_{aB} S_{aC}	$-2/3V_{ab}$	0	$-2/\sqrt{3}\, i_A$	$-\pi/6$
+2	S_{bA} S_{cB} S_{cC}	$2/3V_{bc}$	0	$2/\sqrt{3}\, i_A$	$\pi/2$
-2	S_{cA} S_{bB} S_{bC}	$-2/3V_{bc}$	0	$-2/\sqrt{3}\, i_A$	$\pi/2$
+3	S_{cA} S_{aB} S_{aC}	$2/3V_{ca}$	0	$2/\sqrt{3}\, i_A$	$7\pi/6$
-3	S_{aA} S_{cB} S_{cC}	$-2/3V_{ca}$	0	$-2/\sqrt{3}\, i_A$	$7\pi/6$
+4	S_{bA} S_{aB} S_{bC}	$2/3V_{ab}$	$2\pi/3$	$2/\sqrt{3}\, i_B$	$-\pi/6$
-4	S_{aA} S_{bB} S_{aC}	$-2/3V_{ab}$	$\pi/3$	$-2/\sqrt{3}\, i_B$	$-\pi/6$
+5	S_{cA} S_{bB} S_{cC}	$2/3V_{bc}$	$\pi/3$	$2/\sqrt{3}\, i_B$	$\pi/2$
-5	S_{bA} S_{cB} S_{bC}	$-2/3V_{bc}$	$2\pi/3$	$-2/\sqrt{3}\, i_B$	$\pi/2$
+6	S_{aA} S_{cB} S_{aC}	$2/3V_{ca}$	$2\pi/3$	$2/\sqrt{3}\, i_B$	$7\pi/6$
-6	S_{cA} S_{aB} S_{cC}	$-2/3V_{ca}$	$2\pi/3$	$-2/\sqrt{3}\, i_B$	$7\pi/6$
+7	S_{bA} S_{bB} S_{aC}	$2/3V_{ab}$	$4\pi/3$	$2/\sqrt{3}\, i_C$	$-\pi/6$
-7	S_{aA} S_{aB} S_{bC}	$-2/3V_{ab}$	$4\pi/3$	$-2/\sqrt{3}\, i_C$	$-\pi/6$
+8	S_{cA} S_{cB} S_{bC}	$2/3V_{bc}$	$4\pi/3$	$2/\sqrt{3}\, i_C$	$\pi/2$
-8	S_{bA} S_{bB} S_{cC}	$-2/3V_{bc}$	$4\pi/3$	$-2/\sqrt{3}\, i_C$	$\pi/2$
+9	S_{aA} S_{aB} S_{cC}	$2/3V_{ca}$	$4\pi/3$	$-2/\sqrt{3}\, i_C$	$7\pi/6$
-9	S_{cA} S_{cB} S_{aC}	$-2/3V_{ca}$	$4\pi/3$	$2/\sqrt{3}\, i_C$	$7\pi/6$
0_a	S_{aA} S_{aB} S_{aC}	0	-	0	-
0_b	S_{bA} S_{bB} S_{bC}	0	-	0	-
0_c	S_{cA} S_{cB} S_{cC}	0	-	0	-

4 LC INPUT FILTER

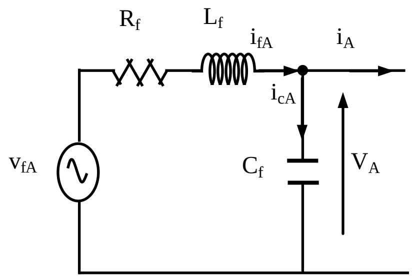

Figure2: Single phase equivalent circuit of input filter.

The input filter can be modelled by the following equations [6, 7]:

$$\begin{cases} V_A(s) = \dfrac{V_{fA}(s) - (L_f s + R_f) I_A(s)}{L_f C_f s^2 + R_f C_f s + 1} \\ I_{fA}(s) = \dfrac{1}{L_f C_f s^2 + R_f C_f s + 1} [I_A(s) + sC_f V_{fA}(s)] \end{cases} \quad (21)$$

Where *x(s)* denotes the Laplace transfer function of *x(t)*.

From (22), the characteristic frequency ω_n and damping factor ζ of the transfer functions are given by (23).

$$\begin{cases} \omega_n = 1/\sqrt{L_f C_f} \\ \zeta = \dfrac{1}{2} R_f \sqrt{C_f / L_f} \end{cases} \quad (22)$$

The design of the input filter has to accomplish the following [6, 7, 8]:

- the value of Cf/Lf is usually less than 1;
- the damping factor ζ is usually very small and approaches zero;
- the cut-off frequency of the input filter is lower the switching frequency (If near 10 KHz switching frequencies are considered, the filter should have a cut-off frequency of lkHZ to 2kHz).

5 SIMULATION RESULTS AND DISCISSIONS

In order to verify the robustness of the proposed control scheme of the induction motor fed by matrix converter (fig.3), the numerical simulations have been carried out using MATLAB, with the parameters of drive system are reported in Appendix.

The simulation test involves the following operating sequences (fig. 4): at the beginning, the motor operate in unloaded mode, with speed reference 100 rd/s and square rotor flux reference 1Wb. At t=0.5s, a 20 Nm load torque is applied, then, this torque is reversed (-20 Nm) with t=1s (fig.4.b), it is interesting to take into account that the torque inversion test is suitable only for one bidirectional converter like the MC, since in this case the motor is a function in generating mode. At t=1.5s, the motor is unloaded again. At t=2, the square of rotor flux reference takes 0.5 Wb as value, and then it returns to its initial value 1 Wb for t=2.5s (fig.4.c). At t=3s, the reference speed is reversed from 100 rd/s to -100 rd/s (fig.4.a). Where, the reference signals for torque and square rotor flux, consist of step functions.

According to figures 4 and 5, one can see that the dynamic performances of nonlinear control law are satisfactory, in both cases (PI and SMC): the steady-state errors in the responses of speed and torque are eliminated by the use a sliding mode controller (fig.5.a,b); in both cases (PI and SMC), one can note that: the decoupling between torque and square of rotor flux, is established, excepting for a small flux error occurs; the responses of speed and square of rotor flux quickly converge to their reference values. In addition, a sinusoidal waveform is obtained for the phase stator current.

From figure 6, in both cases (PI and SMC), one can note that the output MC voltage (stator phase voltage) which is reconstructed by the chopped three-phase input voltage; the input MC voltage is highly sinusoidal. The almost sinusoidal nature of the output MC current (stator phase current) and the current of input filter has also been observed.

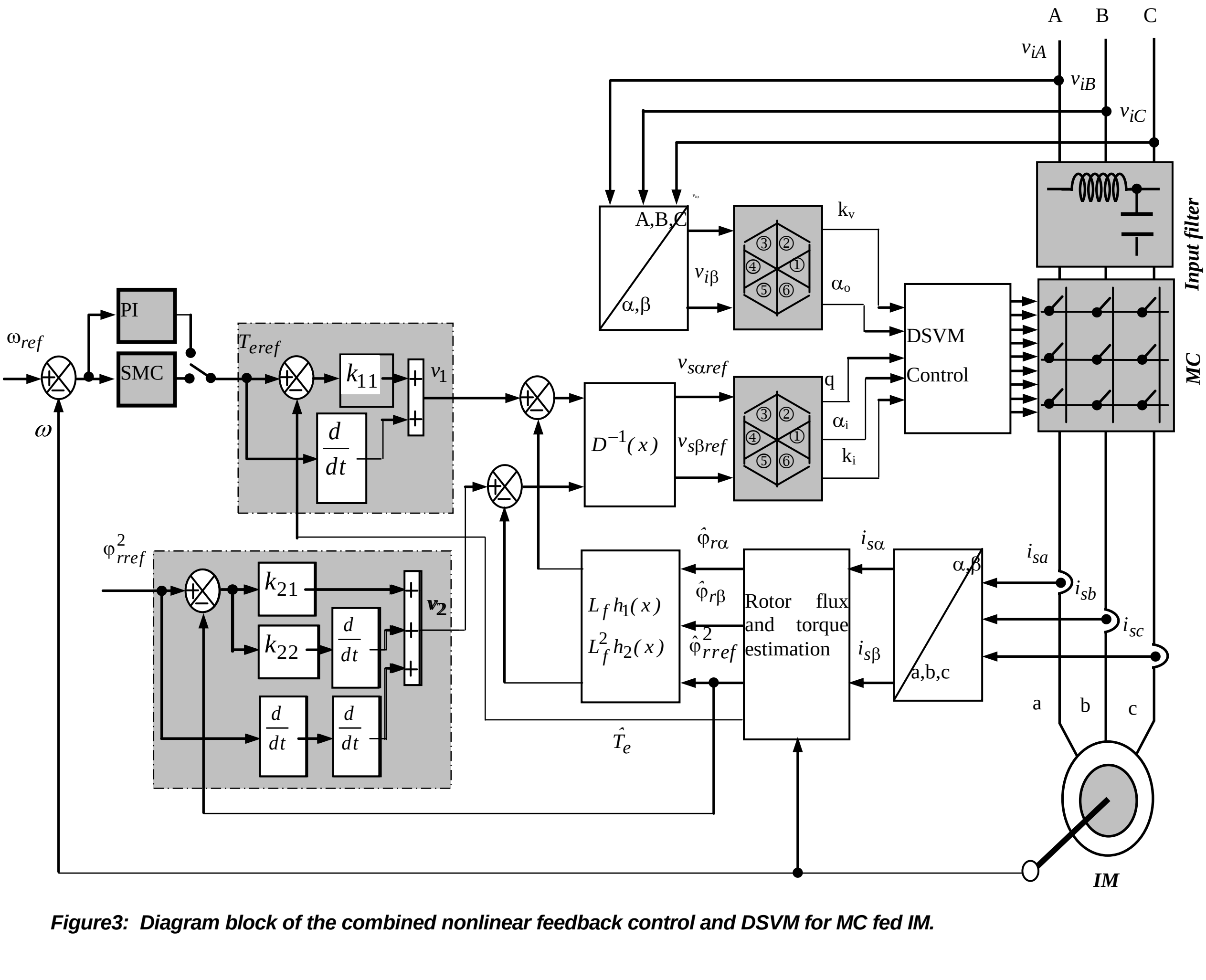

Figure3: Diagram block of the combined nonlinear feedback control and DSVM for MC fed IM.

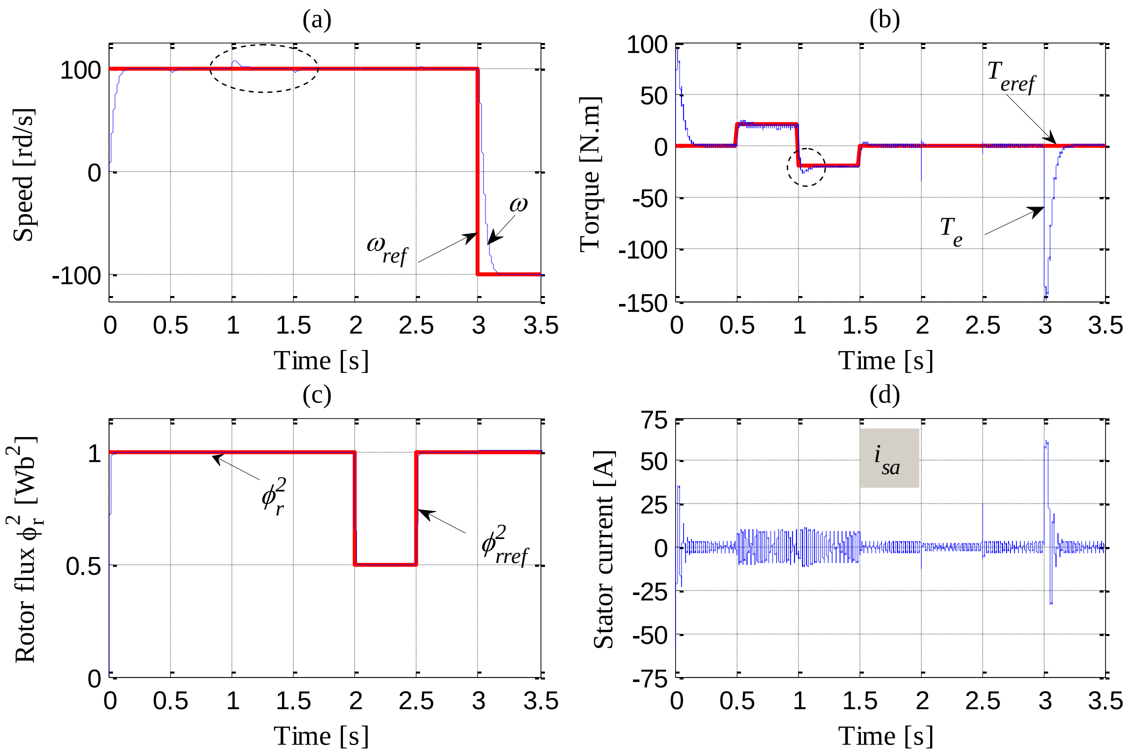

Figure4: Simulation results obtained by using a PI speed controller.

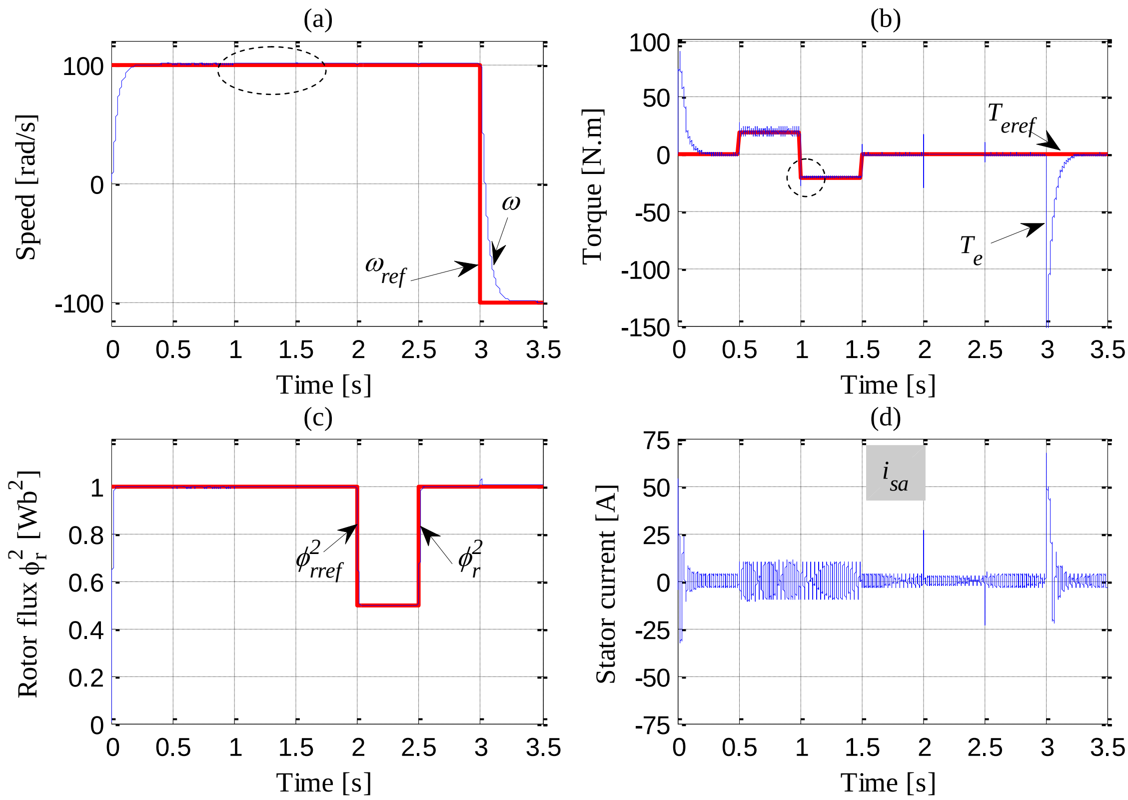

Figure 5: Simulation results obtained by using a sliding mode speed controller.

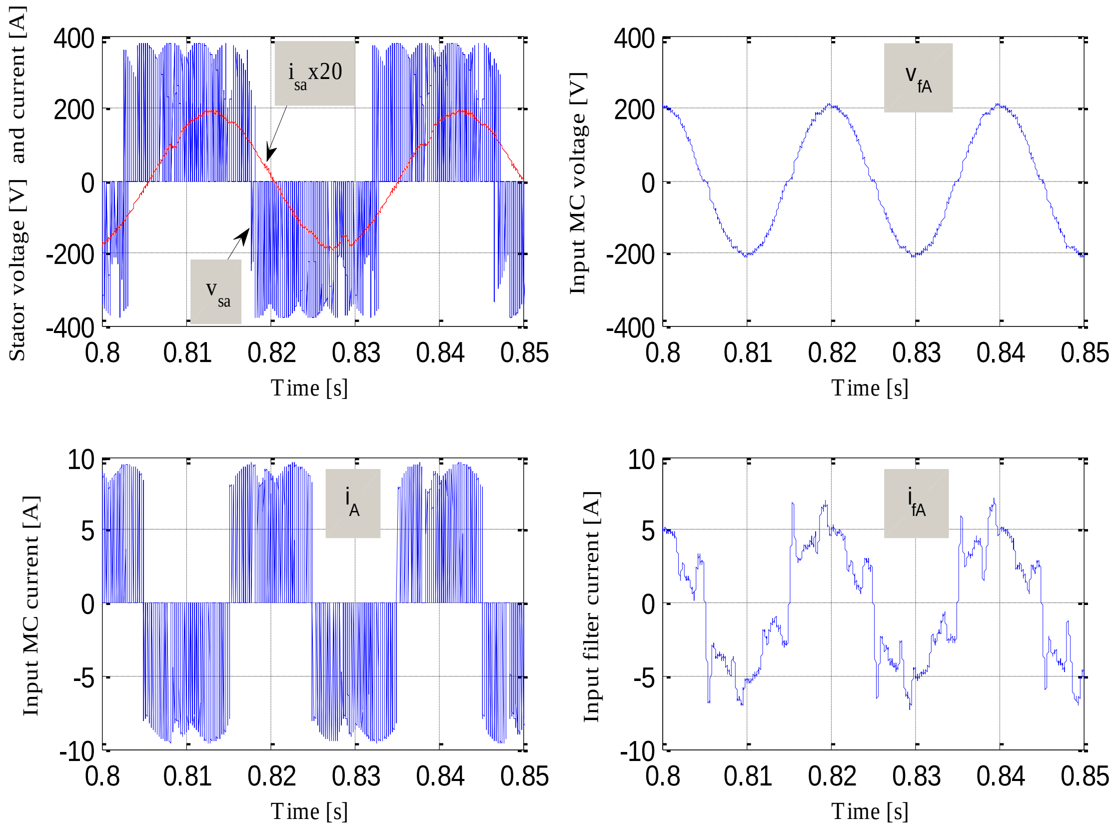

Figure6: Simulation results (PI or SMC): input voltage of the MC, the stator phase voltage and current, the input voltage and current, and the current in input of the filter.

6 CONCLUSION

In this study, a high performance of the proposed control scheme was presented. Furthermore, the decoupling between speed and rotor flux, stability, robustness, and reliability of the drive system have been ensured under external disturbances (load torque, reference speed and rotor flux variations). In order to eliminate the steady-state error of speed, a SMC is employed, taking the place of PI controller.

The proposed solution is suitable for any high dynamic performance applications. The use of the DSVM control technique for matrix converter, make possible an operation of induction motor in the four quadrants. The conventional input LC filter is well adapted to filtering of harmonics, due to the simplicity of implementation, and satisfactory performance provided.

A reading of these results, the practical implementation of the proposed drive system is considered as a prospect for the continuation of this work.

Appendix

- Induction motor parameters:

R_s	stator resistance	2.47 Ω
R_r	rotor resistance	1.24 Ω
L_s	stator inductance	0.236 H
L_r	rotor inductance	0.236 H
L_m	magnetizing inductance	0.2269 H
J	total rotor inertia	0.05 kg.m^2
f	viscous friction coefficient	0.00065 N.m./rd
n_p	pole pairs	2

- Input filter parameters:

C_f=0.0000252 F, L_f=0.0007; R_f=3 Ω.

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