Feedback linearization associated to a sliding mode controller for an operating with unity power factor of induction motor fed by matrix converter

Abdelhakim Dendouga¹

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Abstract

In order to achieve a high performance and to ensure an operation with a unit power factor with respect to imposed reference signals, the input–output feedback linearization technique with a sliding mode controller was proposed to control the induction motor fed by matrix converter drive system. In addition, the direct space vector modulation strategy is adopted to control the matrix converter to minimize the switching losses. On the other hand, a damped input filter has been used to reduce the harmonic total distortion rate injected to the grid. Moreover, the topology and the optimization method for parameter calculation of the input filter are also clarified. The robustness of the proposed control strategy for the drive system was tested and validated by a simulation study.

Keywords Damped RLC input filter · Direct space vector modulation · Feedback linearization · Matrix converter fed induction motor drive system · Sliding mode controller · Unity power factor operation

1 Introduction

In recent decades and thanks to the developments in the fields of microelectronics and computer technology, the matrix converter (MC) is becoming a more attractive topic for several researchers, because of its competitive advantages compared to conventional AC-DC-AC (back-to-back) converters such as: compact size, bidirectional power flow, power adjustment capability, unity power factor operation, does not need a DC-link capacitor [1–5]. The Venturini and space vector modulation (SVM) methods are usually used in the literature to solve the MC control problem. The Venturini method presents a low harmonic rate. However, reduced switching losses is the major advantage of the SVM method [6–8].

The use of a passive filter at the input of MC is necessary to avoid the injection of current harmonics into the grid. In this context, several types of the input filter to solve

the filtering problem in the MC have been proposed. Consequently, the inductive capacitive filter with a damping resistor considered a suitable solution thanks to its satisfactory performance provided [9–11].

Recently, several methods have emerged for the control of induction motor fed by MC. The field-oriented control (FOC) allows a natural decoupling between the speed and flux amplitude. However, the parameter variations and the high operation speed cause loss decoupling [12–14].

The direct torque control (DTC) has found an interesting solution in terms to reduce the dependence on parameters [15, 16]. In spite of its advantages, the variation of stator resistor due to the temperature degrades the performance of this control method [17, 18]. Moreover, an improved model predictive control method (MPC) is proposed for the IM drives fed by an MC [19, 20]. Although the vector selected from MPC is more accurate and effective than the one which selected for DTC, the steady state

✉ Abdelhakim Dendouga, hakimdendouga@yahoo.fr | ¹I3CUB Research Laboratory, Department of Electrical Engineering, University of Biskra, Biskra, Algeria.

performance is still not satisfactory and inferior to that of FOC. The nonlinear control is another alternative strategy to ensure an excellent decoupling for induction motor while ensuring a perfect linearization some is the profiles of the reference trajectories imposed on the system [21, 22]. The authors in [22] propose a second order nonlinear robust auto disturbance rejection controller (ADRC) for control of IM fed by MC. This study presents an interesting initiative for nonlinear approach application to control the considered drive system.

In this work, the input–output feedback linearization technique is used with a sliding mode controller to generate the reference voltages for a MC controlled by direct SVM (DSVM) strategy to achieve a high performance and to ensure an operation with a unit power factor of IM drive system. Moreover, the design procedure of the input filter with damping resistor concerning the transfer function that describes its topology as well the optimal method of parameter selection have been clearly presented. In order to evaluate the robustness of the proposed drive system a simulation study has been carried out.

2 Control strategy design

2.1 Induction motor model

The model of IM in a fixed reference frame (α - β), is given by

$$\begin{aligned}\dot{x} &= f(x) + gu \\ y &= h(x)\end{aligned}\quad (1)$$

where:

$$\begin{aligned}x &= [i_{s\alpha} \ i_{s\beta} \ \varphi_{r\alpha} \ \varphi_{r\beta} \ \omega] \quad u = \begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix}; \\ f(x) &= \begin{bmatrix} -\gamma i_{s\alpha} + \frac{K}{T_r} \varphi_{r\alpha} + Kn_p \omega \varphi_{r\beta} \\ -\gamma i_{s\beta} + \frac{K}{T_r} \varphi_{r\beta} - Kn_p \omega \varphi_{r\alpha} \\ \frac{L_m}{T_r} i_{s\alpha} - \frac{1}{T_r} \varphi_{r\alpha} - n_p \omega \varphi_{r\beta} \\ \frac{L_m}{T_r} i_{s\beta} - \frac{1}{T_r} \varphi_{r\beta} + n_p \omega \varphi_{r\alpha} \\ \frac{n_p L_m}{J L_r} (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) - \frac{1}{J} (B\omega + T_L) \end{bmatrix}; \\ g &= \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \end{bmatrix}.\end{aligned}$$

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

2.2 Feedback linearization

The torque (T_e) and the norm of the rotor flux ($\varphi_{r\alpha}^2 + \varphi_{r\beta}^2$) are defined as the outputs of the system [23], [24].

$$h(x) = \begin{bmatrix} h_1(x) \\ h_2(x) \end{bmatrix} = \begin{bmatrix} T_e \\ \varphi_r^2 \end{bmatrix} = \begin{bmatrix} \frac{n_p L_m}{L_r} (\varphi_{r\alpha} i_{s\beta} - \varphi_{r\beta} i_{s\alpha}) \\ \varphi_{r\alpha}^2 + \varphi_{r\beta}^2 \end{bmatrix} \quad (2)$$

The Lie derivatives of the outputs are given by:

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

$$\dot{y}_1 = \dot{h}_1(x) = L_f h_1(x) + L_{g_1} h_1(x) v_{s\alpha} + L_{g_2} h_1(x) v_{s\beta} \quad (3)$$

where

$$\begin{cases} L_f h_1(x) = \frac{n_p L_m}{L_r} \left[\left(\gamma + \frac{1}{T_r} \right) (\varphi_{r\beta} i_{s\alpha} - \varphi_{r\alpha} i_{s\beta}) - Kn_p \omega (\varphi_{r\alpha}^2 + \varphi_{r\beta}^2) - n_p \omega (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) \right] \\ L_{g_1} h_1(x) = \frac{\partial h_1(x)}{\partial x_i} g_1 = -\frac{n_p L_m}{\sigma L_r L_s} \varphi_{r\beta} \\ L_{g_2} h_1(x) = \frac{\partial h_1(x)}{\partial x_i} g_2 = \frac{n_p L_m}{\sigma L_r L_s} \varphi_{r\alpha} \end{cases} \quad (4)$$

The relative degree related to first output is: $r_1 = 1$.

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

$$\begin{cases} \dot{y}_2 = \dot{h}_2(x) = L_f h_2(x) \\ \ddot{y}_2 = \ddot{h}_2(x) = L_f^2 h_2(x) + L_{g_1} (L_f h_2(x)) v_{s\alpha} + L_{g_2} (L_f h_2(x)) v_{s\beta} \end{cases} \quad (5)$$

where

$$\begin{cases} L_f h_2(x) = \frac{2L_m}{T_r} (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) - \frac{2}{T_r} (\varphi_{r\alpha}^2 + \varphi_{r\beta}^2) \\ L_f^2 h_2(x) = \frac{-2L_m}{T_r} \left[\left(\gamma + \frac{3}{T_r} \right) (\varphi_{r\alpha} i_{s\alpha} + \varphi_{r\beta} i_{s\beta}) + n_p \omega (\varphi_{r\beta} i_{s\alpha} - \varphi_{r\alpha} i_{s\beta}) \right. \\ \quad \left. + \left(\frac{2KL_m + 4}{T_r^2} \right) (\varphi_{r\alpha}^2 + \varphi_{r\beta}^2) + \frac{2L_m^2}{T_r^2} (i_{r\alpha}^2 + i_{r\beta}^2) \right] \\ L_{g_1} (L_f h_2(x)) = \frac{\partial (L_f h_2(x))}{\partial x_i} g_1 = \frac{2L_m}{\sigma L_s T_r} \varphi_{r\alpha} \\ L_{g_2} (L_f h_2(x)) = \frac{\partial (L_f h_2(x))}{\partial x_i} g_2 = \frac{2L_m}{\sigma L_s T_r} \varphi_{r\beta} \end{cases} \quad (6)$$

In this case: $r_2 = 1$. The total relative degree is $r = r_1 + r_2 = 3$, in this case, the full linearization is not

achieved. To solve this problem the following change of variables is necessary [25]:

$$\begin{cases} z_1 = y_1 \\ z_2 = y_2 \\ z_3 = \dot{z}_2 = L_f h_2(x) \\ z_4 = \arctan\left(\frac{\varphi_{r\beta}}{\varphi_{r\alpha}}\right) \\ z_5 = \omega \end{cases} \quad (7)$$

The derivation of the system of Eqs. (7) is:

$$\begin{cases} \dot{z}_1 = v_1 \\ \dot{z}_2 = z_3 \\ \dot{z}_3 = v_2 \\ \dot{z}_4 = n_p \omega + \frac{R_r z_1}{n_p z_2} \\ \dot{z}_5 = \frac{1}{J}(z_1 - T_L - f z_5) \end{cases} \quad (8)$$

v_1 and v_2 of system are defined by

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} L_f h_1 \\ L_f^2 h_2 \end{bmatrix} + D(x) \begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix} \quad (9)$$

where $D(x)$ is the decoupled matrix defined as

$$D(x) = \begin{bmatrix} L_{g_1} h_1 & L_{g_2} h_1 \\ L_{g_1} L_f h_2 & L_{g_2} L_f h_2 \end{bmatrix} = \begin{bmatrix} -\frac{n_p L_m}{\sigma L_r L_s} \varphi_{r\beta} & \frac{n_p L_m}{\sigma L_r L_s} \varphi_{r\alpha} \\ \frac{2L_m}{\sigma L_s T_r} \varphi_{r\alpha} & \frac{2L_m}{\sigma L_s T_r} \varphi_{r\beta} \end{bmatrix} \quad (10)$$

Since $\det(D(x)) \neq 0$, $D(x)$ is nonsingular everywhere in ω . After the arranging of Eq. (9) we obtain

$$\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \end{bmatrix} = D^{-1}(x) \begin{bmatrix} v_1 - L_f h_1 \\ v_2 - L_f^2 h_2 \end{bmatrix} \quad (11)$$

The input signals v_1 and v_2 in (11) are designed as [25]

$$\begin{cases} v_1 = k_{11}(z_{1ref} - z_1) + \dot{z}_{1ref} \\ v_2 = k_{21}(z_{2ref} - z_2) + k_{22}(\dot{z}_{2ref} - \dot{z}_2) + \ddot{z}_{2ref} \end{cases} \quad (12)$$

where k_{11} , k_{21} and k_{22} are positive factors parameters can be calculated by imposing a dynamics on the error according to

$$\begin{cases} k_{11}e_1 + \dot{e}_1 = 0 \\ k_{21}e_2 + k_{22}\dot{e}_2 + \ddot{e}_2 = 0 \end{cases} \quad (13)$$

where $e_1 = (z_{1ref} - z_1)$, $e_2 = (z_{2ref} - z_2)$

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 [25, 26].

2.3 Sliding mode controller design

The control objective in this section is to design a sliding-mode speed controller for the IM system. In order

to guarantee an excellent speed ($z_5 = \omega$) tracking with respect to the variation of its reference, a SMC is designed taking the place of conventional PI controller, as follows [27]:

$$z_{1ref} = (z_{1ref})_{eq} + (z_{1ref})_n \quad (14)$$

For the speed, the slip surface is selected as

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

its time derivate is expressed by

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

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

$$(z_{1ref})_{eq} = \dot{z}_{5ref} + \frac{1}{J}(f z_{5ref} + T_L) \quad (17)$$

In order to minimize the chattering phenomenon (high frequency oscillations) due to the discontinuous nature of the SMC, a smooth function is used for the nonlinear component. In our case, we take the following function.

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

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

3 Direct space vector modulation

From 27 possible configurations of the MC, there are only 21 configurations containing 18 stationary vectors (from ± 1 to ± 9) and 3 null vectors, can be used by the DSVM technique to form the output voltage and the input current for the MC (Fig. 1). The representation of the 18 stationary vectors for the voltage and for the current in the fixed frame of reference (α, β), makes it possible to form two regular hexagons, each two consecutive vectors forming one sector. In this context, in each modulation period four stationary vectors are selected by the DSVM technique to calculate the switching times for MC. Moreover, the modulation period is completed by the application of the null vector. The expressions that allow the determination of switching times are given by [5, 7, 12]:

$$\delta_1 = \frac{2}{\sqrt{3}} q \sin \left[\alpha_o - (k_v - 1) \frac{\pi}{3} \right] \sin \left[\frac{\pi}{6} - \left(\alpha_i - (k_i - 1) \frac{\pi}{3} \right) \right] \quad (19)$$

$$\delta_2 = \frac{2}{\sqrt{3}} q \sin \left[\alpha_o - (k_v - 1) \frac{\pi}{3} \right] \sin \left[\frac{\pi}{6} + \left(\alpha_i - (k_i - 1) \frac{\pi}{3} \right) \right] \quad (20)$$

$$\delta_3 = \frac{2}{\sqrt{3}} q \sin \left[k_v \frac{\pi}{3} - \alpha_o \right] \sin \left[\frac{\pi}{6} - \left(\alpha_i - (k_i - 1) \frac{\pi}{3} \right) \right] \quad (21)$$

$$\delta_4 = \frac{2}{\sqrt{3}} q \sin \left[k_v \frac{\pi}{3} - \alpha_o \right] \sin \left[\frac{\pi}{6} + \left(\alpha_i - (k_i - 1) \frac{\pi}{3} \right) \right] \quad (22)$$

$$\delta_0 = 1 - (\delta_1 + \delta_2 + \delta_3 + \delta_4) \quad (23)$$

where $q = V_o/V_i$ is the voltage ratio; α_o and α_i are the phase angles of the output voltage and input current vectors, respectively; k_v, k_i are the sector number of output voltage and input current vectors.

4 Damped RLC input filter design

Figure 2 shows a more suitable approach for damped input filter design, which uses a resistor connected in parallel with the capacitor of input filter to increase the damping factor of the LC filter. In order to improve the current waveform at the input of MC, and therefore to reduce harmonic injection to the grid [10, 11]. The voltage and current transfer function are given by Eqs. (24) and (25).

$$V_A(s) = \frac{R_d \cdot V_{gA}(s) - R_d \cdot (L_f \cdot s + R_f) I_A(s)}{R_d \cdot L_f \cdot C_f \cdot s^2 + (R_d \cdot R_f \cdot C_f + L_f) \cdot s + (R_d + R_f)} \quad (24)$$

Fig. 1 Output voltage and input current space vectors

$$I_{gA}(s) = \frac{(R_d \cdot C_f \cdot s + 1) \cdot V_{gA}(s) + R_d \cdot I_A(s)}{R_d \cdot L_f \cdot C_f \cdot s^2 + (R_d \cdot R_f \cdot C_f + L_f) \cdot s + (R_d + R_f)} \quad (25)$$

The parameter selection of C and L must satisfy the following condition:

- $10f_i < \frac{1}{2\pi\sqrt{L_f C_f}} < 10f_s$, f_i the input supply frequency and f_s is the switching frequency;
- Considering that the value of R_d sufficiently large and R_f is usually few ohms. The characteristic frequency ω_n and damping factor ζ can be given by

$$\omega_n = \frac{1}{\sqrt{L_f C_f}}, \quad \zeta = \frac{1}{2R_d} \sqrt{\frac{L_f}{C_f}}$$

5 Simulation and results

The unit power factor operating of drive system and the robustness against the imposed reference of the proposed control strategy have been tested and evaluated by simulation. In this case, the robustness test involves the following operating sequences: at the beginning, the motor operates in unloaded mode. At $t = 0.5$ s, a load torque is applied, then the motor is unloaded again at $t = 1.5$ s. At $t = 2$, the speed reference takes 50 rd/s as value, and then it is reversed at $t = 2.5$ s. At time $t = 3$ s, the reference speed takes its initial value (100rd/s). The sequences and modes of operation are summarized in Table 1.

After the simulation using the parameters shown in "Appendix" section, we obtain the plots of speed, torque, grid voltage and current (input filter voltage and current),

Fig. 2 Single phase RLC input filter with a damping resistor

Table 1 Operation modes

Time	0s	0.5s	1.5s	2s	2.5s	3s	3.5s
$\omega_{mref}(\text{rd/s})$		100		50	-100	100	
$T_{eref}(\text{N.m})$	0	25		0

harmonic spectrum of grid side current, stator voltage (or output mc voltage). These results allow us to evaluate the robustness and performance achieved by the proposed drive system (Fig. 3).

The simulation results shown by Fig. 4 exhibit that the dynamic performances of feedback linearization based on SMC are very acceptable in regards to the good pursuit, short response time, no steady-state errors occur.

In addition, a sinusoidal waveform is obtained for the current, in which has been confirmed by the harmonics spectrum (THD=3.12). One can note also from this figure that the output MC voltage (stator phase voltage) which is formed by the chopped three-phase input voltage.

This work clearly confirms the high performance of the proposed drive system. This is proved by the good tracked against the reference signals, unit power factor operation, and robustness of the drive system have been ensured regarding to the external disturbances. The SMC presents a high performance, especially in terms of the good pursuit, short response time, no steady-state errors occur.

Moreover, the RLC input filter with a damping resistor makes it possible to obtain a near-sinusoidal waveform for the current in the grid side with a low harmonic rate.

The proposed solution is satisfactory for any high dynamic performance applications. In this context, the practical implementation of the proposed drive system is considered as a prospect for the continuation of this work.

Fig. 4 Test by simulation of operating with an unit power factor and robustness evaluation of the proposed drive system

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Compliance with ethical standards

Conflict of interest The author declare that he has no conflict of interest.

Appendix

See Table 2.

- Input filter parameters:

$$R_d = 30 \, \Omega, L_f = 700 \, \mu\text{H}, C_f = 1.1 \, \mu\text{F}.$$

Table 2 Induction motor parameters

Symbol	Nomenclature	Value
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 Nm/rd
n_p	Pole pairs	2

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