

HIGH PERFORMANCE MOTOR DRIVE USING MATRIX CONVERTER

*Abdelkader DJAHBAR, **Benyounés MAZARI

*Department of Electrical Engineering, Faculty of Electrical Engineering,
University of Oran, Oran 31000, Algeria
E-mail: a_djahbar@yahoo.fr

** Laboratory of Development of Electric Drives, Department of Electrical Engineering,
Faculty of Electrical and Electronics Engineering,
University of Sciences and Technology (USTO), Oran, Algeria

SUMMARY

The presented paper deals with the designed concept of the matrix converter control system employed in the industry applications. Matrix converters provide an all-silicon solution to the problem of converting AC power from one frequency to another, offering almost all the features required of an ideal static frequency changer. They possess many advantages compared to the conventional voltage or current source inverters. A matrix converter does not require energy storage components as a bulky capacitor or an inductance in the DC-link. It enables the bi-directional power flow between the power supply and load. The most of the contemporary modulation strategies are able to provide practically sinusoidal waveforms of the input and output currents with negligible low order harmonics and to control the input displacement factor. The solution is advantageous especially with regard to the above mentioned reactive component reduction.

A new direct AC/AC converter topology based on the scalar modulation strategy with three intervals and vector control technique for matrix converter fed induction motor drive system is proposed. It is shown, through analysis and simulation, that the proposed matrix converter topology is free of any switching difficulties and gives high quality sinusoidal waveform on both sides. The primary objective of this research effort is to evaluate the utility of the matrix converter in industry applications, primarily for motor control. Closed-loop direct and indirect vector control has been presented and simulated using a 1.5 Kw induction motor. By applying this control strategy, we will be able to combine the advantages of matrix converter and the advantages of the vector control. Simulation results are shown to demonstrate the effectiveness of the proposed control scheme.

Keywords: Matrix converter, Induction Machine, Scalar Modulation, Vector control.

1. INTRODUCTION

Matrix converter fed motor drive is superior to pulse width modulation (PWM) inverter drives because it provides bidirectional power flow, sinusoidal input/output currents, and adjustable input power factor [1], [2]. Furthermore, matrix converter allows a compact design due to the lack of dc-link capacitors for energy storage. However, only a few of practical matrix converters have been applied to vector control system of induction motors (IM) for the reason: Modulation technique and commutation control are more complicated than conventional PWM inverter [2].

In order to realize high performance control of matrix converter fed induction motor drive system, the scalar modulation strategy of matrix converter is used [3]-[5]. It's based on algorithm proposed by Roy [4]. The synoptic diagram of matrix converter fed an induction machine is given to the figure 1.

The basic matrix converter configuration with high frequency control was originally introduced in 1980 [4]. Since then, matrix converters have been subject of intensive research which mostly concentrated on two aspects:

Implementation of the matrix converter switches and the matrix converter control [5].

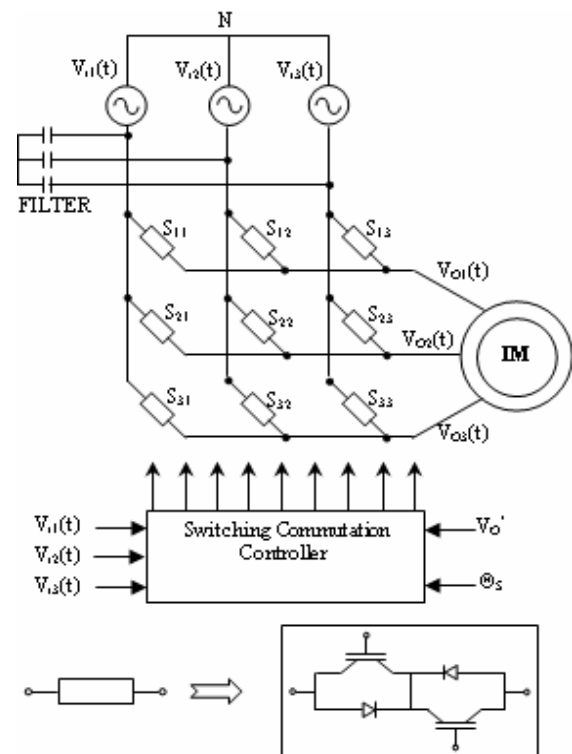


Fig. 1 The topology of three-phase to three-phase matrix converter and the configuration of bi-directional switch.

2. DYNAMIC MODELING OF INDUCTION MACHINE

It is well known that the mathematical model of an induction machine can be obtained using the two-

axis theory. By choosing the reference frame d-q, the dynamics of a squirrel cage induction machine can be represented by the following nonlinear differential equations:

$$\begin{cases} \frac{di_{ds}}{dt} = \frac{1}{L_s} \left[-\left(R_s + \frac{L_m^2}{L_r^2} R_r\right) i_{ds} + \omega_s L_s i_{qs} + \frac{L_m}{T_r L_r} \phi_{dr} + p \omega_m \frac{L_m}{L_r} \phi_{qr} + v_{ds} \right] \\ \frac{di_{qs}}{dt} = \frac{1}{L_s} \left[-\omega_s L_s i_{ds} - \left(R_s + \frac{L_m^2}{L_r^2} R_r\right) i_{qs} - p \omega_m \frac{L_m}{L_r} \phi_{dr} + \frac{L_m}{T_r L_r} \phi_{qr} + v_{qs} \right] \\ \frac{d\phi_{dr}}{dt} = \frac{L_m}{T_r} i_{ds} - \frac{1}{T_r} \phi_{dr} + (\omega_s - p \omega_m) \phi_{qr} \\ \frac{d\phi_{qr}}{dt} = \frac{L_m}{T_r} i_{qs} - (\omega_s - p \omega_m) \phi_{dr} - \frac{1}{T_r} \phi_{qr} \\ \frac{d\omega_m}{dt} = \frac{1}{J} (C_{em} - C_r - B \omega_m) \end{cases} \quad (1)$$

The developed torque equation is given by:

$$T_e = p \frac{L_m}{L_r} (\phi_{dr} i_{qs} - \phi_{qr} i_{ds})$$

$$\begin{cases} \frac{d\Phi_r}{dt} = \frac{R_r}{L_r} (L_m i_{ds} - \Phi_r) \\ \frac{d\Theta_s}{dt} = \omega_s = \omega_m + \frac{L_m i_{qs}}{T_r \Phi_{qr}} \end{cases} \quad (2)$$

3. DIRECT FIELD ORIENTED CONTROL

Direct field oriented control, published for the first time by Blaschke in his pioneering work in 1972, consists in adjusting the flux by a component of the current and the torque by the other component. For this purpose, it is necessary to choose a d-q reference frame rotating synchronously with the rotor flux space vector, in order to achieve decoupling control between the flux and the produced torque. This technique allows to obtain a dynamical model similar to the DC machine [1]. The regulation of the flux can be direct or indirect. In the direct control, the rotor flux is estimated or reconstructed fig.2. The field orientation is obtained by imposing the condition ($\phi_{dr} = \phi_r$ and $\phi_{qr} = 0$). From this condition and eqns.1, the i_{ds} reference can be computed in order to impose the flux ϕ_r . Furthermore, the position θ_s of the rotating frame can be estimated using eqns.1.

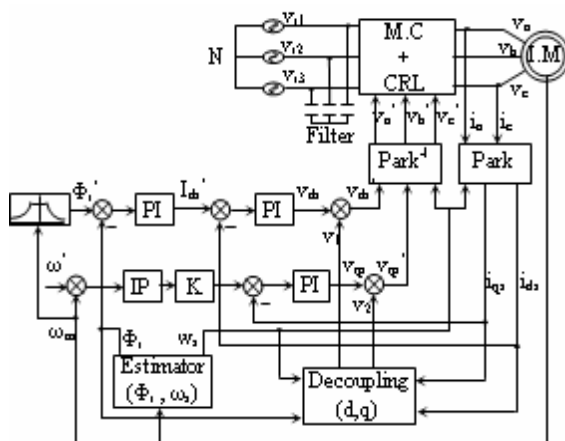


Fig. 2 Synoptic scheme of the direct field oriented control of IM. fed from a matrix converter ($k=L_r/pL_m\Phi_r^*$).

With taking into account the field orientation of the machine, the stator equations on d-q axis become:

$$\begin{cases} v_{ds} = R_s i_{ds} + \sigma L_s \frac{di_{ds}}{dt} + \frac{L_m}{L_r} \frac{d\Phi_r}{dt} - \sigma L_s \omega_s i_{qs} \\ v_{qs} = R_s i_{qs} + \sigma L_s \frac{di_{qs}}{dt} + \omega_s \frac{L_m}{L_r} \Phi_r + \sigma L_s \omega_s i_{ds} \end{cases} \quad (3)$$

Using the system given by eqn.3, we can remark the interaction of both inputs, which makes the control design more difficult. The first step of our work is to obtain a decoupled system in order to control the electromagnetic torque via stator quadrature current i_{qs} such as DC machine. A decoupled model can be obtained by using two intermediate variables:

$$\begin{cases} v_{ds}^* = v_{ds} + v_1 \\ v_{qs}^* = v_{qs} + v_2 \end{cases} \quad (4)$$

Or v_1 and v_2 are electromotive forces that introduce a coupling between the axes d and q.

The stator voltages (v_{ds} , v_{qs}) are reconstituted from (v_{ds}^* , v_{qs}^*). The speed regulation is reached by an IP regulator type, flux and currents by PI regulators Fig.2.

4. INDIRECT FIELD ORIENTED CONTROL

The indirect field oriented control (IFOC) technique is very useful for implementing high performance induction motor drive systems. In general in the IFOC technique the shaft speed, that is

usually measured, and the slip speed, that is calculated based on the machine parameters, are added to define the angular frequency of the rotor flux vector. The standard IFOC technique is essentially a feedforward scheme and has the drawback of being dependent on the motor temperature and the level of magnetic excitation of the motor.

5. MATRIX CONVERTER THEORY

5.1. Problem definition

The schematic diagram of a three-phase matrix converter is presented in fig.1. Its inputs are the phase voltages v_{i1}, v_{i2}, v_{i3} , and its outputs are the voltages v_{o1}, v_{o2}, v_{o3} , (in equation 6 this notation is used for the first harmonic of the output voltages). The matrix converter components ($S_{11}, S_{12}, \dots, S_{33}$) represent nine bi-directional switches which are capable of blocking voltage in both directions and of switching without any delays. These are nine ideal switches.

The matrix converter connects the three given inputs, with constant amplitude V_i and frequency $f_i = \omega_i / 2\pi$, through the nine switches to the output terminals in accordance with precalculated switching angles. The three-phase output voltages obtained have controllable amplitudes V_o and frequency $f_o = \omega_o / 2\pi$.

The input three-phase voltages of the converter are given by:

$$\begin{bmatrix} v_{i1} \\ v_{i2} \\ v_{i3} \end{bmatrix} = V_{im} \begin{bmatrix} \cos(\omega_i t) \\ \cos(\omega_i t - 2\pi/3) \\ \cos(\omega_i t - 4\pi/3) \end{bmatrix} \quad (5)$$

The required first harmonic of the output phase voltages of the unloaded matrix converter is [3]:

$$\begin{bmatrix} v_{o1} \\ v_{o2} \\ v_{o3} \end{bmatrix} = V_{om} \begin{bmatrix} \cos(\omega_o t) \\ \cos(\omega_o t - 2\pi/3) \\ \cos(\omega_o t - 4\pi/3) \end{bmatrix} \quad (6)$$

The problem at hand may be defined as follows: with input voltages as equation (5), the matrix converter switching angles equations will be formulated so that the first harmonic of the output voltages will be as equation (6).

5.2. The switching angles formulation

The switching angles, of the nine bidirectional switches s_{ij} ($i=1,2,3$ and $j=1,2,3$) which will be calculated, must comply with the following rules[5]:

1. At any time 't', only one switch s_{ij} ($j=1,2,3$) will be in 'ON' state. This assures that no short circuit will occur at the input terminals.
2. At any time 't', at least two of the switches s_{ij} ($i=1,2,3$) will be in 'ON' state. This condition guarantees a closed-loop path for the load current (usually this is an inductive current).
3. The switching frequency is f_s and its angular frequency ($\omega_s = 2\pi f_s$) complies with $\omega_s \gg \omega_i$ and ω_o : in other words the switching frequency is much higher ($f_s \approx 20 \times \max(f_i, f_o)$) than the input and output frequencies).

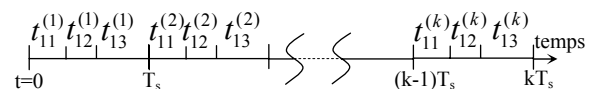


Fig. 3 Segmentation of the axis time for the consecutive orders of intervals closing of the switches.

During the k^{th} switching cycle T_s ($T_s = 1/f_s$) fig.3, the first phase output voltage is given by:

$$v_{o1} = \begin{cases} v_{i1} & 0 \leq t - (k-1)T_s < m_{11}^k T_s \\ v_{i2} & m_{11}^k T_s \leq t - (k-1)T_s < (m_{11}^k + m_{12}^k) T_s \\ v_{i3} & (m_{11}^k + m_{12}^k) T_s \leq t - (k-1)T_s < T_s \end{cases} \quad (7)$$

Where 'm's are defined by :

$$m_y^k = \frac{t_{ij}^k}{T_s} \quad (8)$$

Where t_{ij}^k : time interval when s_{ij} is in 'ON' state, during the k^{th} cycle, and k is being the switching cycle sequence number.

The 'm's have the physical meaning of duty cycle.

Also,

$$\sum_{j=1}^3 m_y^k = m_{i1}^k + m_{i2}^k + m_{i3}^k = 1 \quad \text{and} \quad 0 < m_y^k < 1 \quad (9)$$

Which means that during every cycle T_s all switches will turn on and off once.

5.3. Algorithm of VENTURINI

The algorithm of VENTURINI [2], [3], allows a control of the S_{ij} switches so that the low frequency parts of the synthesized output voltages v_{oij} and the input currents i_{ij} are purely sinusoidal with the prescribed values of the output frequency, the input frequency, the displacement factor and the input amplitude. The average values of the output voltages during the k^{th} sequence are thus given by:

$$\begin{aligned}
v_{o1}^{(k)} &= v_{i1}^{(k)} \frac{t_{11}}{T_s} + v_{i2}^{(k)} \frac{t_{12}}{T_s} + v_{i3}^{(k)} \frac{t_{13}}{T_s} \\
v_{o2}^{(k)} &= v_{i1}^{(k)} \frac{t_{21}}{T_s} + v_{i2}^{(k)} \frac{t_{22}}{T_s} + v_{i3}^{(k)} \frac{t_{23}}{T_s} \\
v_{o3}^{(k)} &= v_{i1}^{(k)} \frac{t_{31}}{T_s} + v_{i2}^{(k)} \frac{t_{32}}{T_s} + v_{i3}^{(k)} \frac{t_{33}}{T_s}
\end{aligned} \quad (10)$$

If times of conduction are modulated in the shape of sinusoid with the pulsation w_m while T_s remains constant, such as $w_o = w_i + w_m$, these times are defined as follows:

1. For the 1st phase, we have:

$$\begin{aligned}
t_{11} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta)) \\
t_{12} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta - \frac{2\pi}{3})) \\
t_{13} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta - \frac{4\pi}{3}))
\end{aligned} \quad (11)$$

$$[M^{(k)}] = \frac{1}{3} \begin{bmatrix} 1+2q\cos A & 1+2q\cos(A-\frac{2\pi}{3}) & 1+2q\cos(A-\frac{4\pi}{3}) \\ 1+2q\cos(A-\frac{4\pi}{3}) & 1+2q\cos A & 1+2q\cos(A-\frac{2\pi}{3}) \\ 1+2q\cos(A-\frac{2\pi}{3}) & 1+2q\cos(A-\frac{4\pi}{3}) & 1+2q\cos A \end{bmatrix}$$

Where:
$$\begin{cases} A = w_m t + \theta \\ w_m = w_o - w_i \end{cases}$$

The running matrix converter with the VENTURINI algorithm generates at the output a three-phases sinusoidal voltage system having in that order pulsation w_m , a phase angle θ and amplitude $q \cdot V_{im}$ ($0 < q < 0.866$ with modulation of the neutral)[3].

5.4. Scalar Method of ROY

This method uses the instantaneous ratios of the input phase voltages so as to determinate the modulation matrix coefficients m_{ij} . The average value of the output voltage for the 1st phase is given by [1]:

$$v_{o1} = \frac{1}{T_s} (t_k v_k + t_l v_l + t_m v_m) = v_{om} \cos(w_o t) \quad (15)$$

With: $t_k + t_l + t_m = T_s$

2. For the 2nd phase:

$$\begin{aligned}
t_{21} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta - \frac{4\pi}{3})) \\
t_{22} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta)) \\
t_{23} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta - \frac{2\pi}{3}))
\end{aligned} \quad (12)$$

3. For the 3rd phase:

$$\begin{aligned}
t_{31} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta - \frac{2\pi}{3})) \\
t_{32} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta - \frac{4\pi}{3})) \\
t_{33} &= \frac{T_s}{3} (1 + 2q \cos(w_m t + \theta))
\end{aligned} \quad (13)$$

Where θ is initial phase angle.
The output voltage is given by [5]:

$$[V_o^{(k)}] = [M^{(k)}][V_i^{(k)}] \quad (14)$$

Where K, L and m are indexes, which can be assigned with the input voltage indexes j (j=1,2,3) according to following rules [1]:

1st rule: At any moment, the input phase voltage having the discriminate polarity from the two other input phases is assigned with L.

2nd rule: The two input phase voltages having the same polarity are assigned with M and K, smallest of both (in absolute value), being K. Times of conduction t_k and t_l are selected such as:

$$\frac{t_k}{t_l} = \frac{v_k}{v_l} = \rho \quad (16)$$

In addition: $0 \leq \rho \leq 1$

The converter chopping is only up to the scalar comparison between input phase voltages and the instantaneous desired output voltage value v_o . Times of conduction t_k , t_l and t_m during one period of switching are given as follows [1]:

$$\begin{cases} t_l = \frac{T_s(v_o - v_m)}{\rho v_k + v_l - (1 + \rho)v_m} \\ t_k = \rho t_l \\ t_m = T_s - (\rho + 1)t_l \end{cases} \quad (17)$$

The computation of conduction times t_k , t_l and t_m for a transfer ratio of $q \leq 0.57$ gives positive values similar to the method of VENTURINI. But for a ratio higher than 0.57, some of these times of conduction will have negative values because of the instantaneous input voltage limitation of the converter. However this ratio can reach value 0.866 with the use of the neutral modulation method [4]:

$$v_{oj(j=1,2,3)} = V_{om} \cos(\omega_o t + j \frac{2\pi}{3}) + \frac{V_{im}}{4} \cos(3\omega_i t) - \frac{V_{om}}{6} \cos(3\omega_o t) \quad (18)$$

6. SIMULATION RESULTS

6.1. Performances of VENTURINI strategy

With this strategy, we simulate the simple and compound voltages delivered by the matrix converter for two values of the switching frequency (Fig. 4 and Fig. 5).

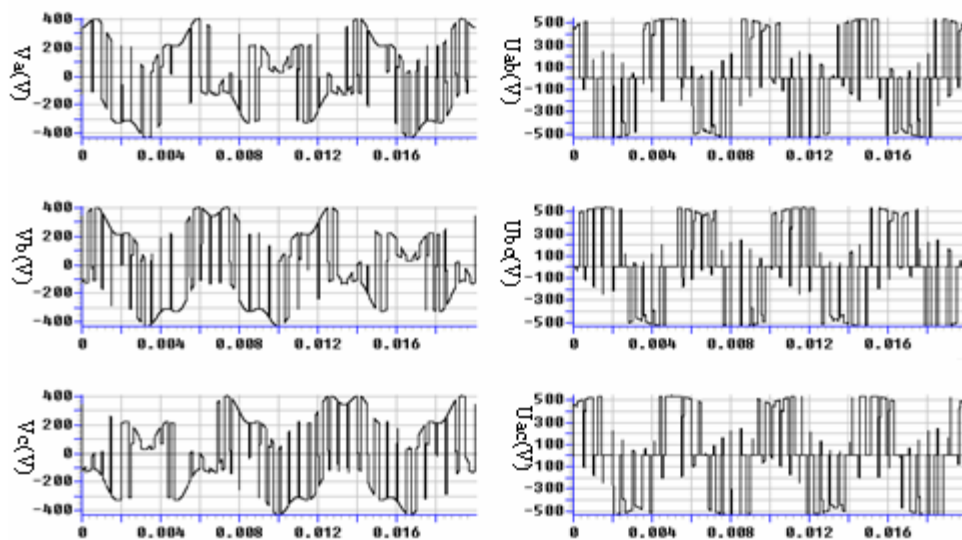


Fig. 4 voltages waveforms v_a , v_b , v_c , u_{ab} , u_{bc} and u_{ac} ,
for: $f_s=1$ kHz, $f_i=50$ Hz, $f_o=25$ Hz, $q=0.57$

Simulation results of the starting process of an induction motor fed from a matrix converter for a frequency $f_s=2$ kHz are shown on the fig.6.

6.2. Performances of Scalar strategy

The average output voltage without and with modulation of the neutral are depicted on the Fig.7, one notices that the output voltages are limited to half for $q=0.57$.

The simple and compound voltages waveforms delivered by the Matrix Converter for two values of f_s are shown on the fig.8,9.

The simulation of the vector control of an Induction Machine associated with a Matrix Converter controlled by the scalar modulation strategy is given by the figs.10,11 and 12. The fig.10 shows the decoupling carried out between the flux and the electromagnetic torque, the fig.11 represents the speed, torque, flux d-q, current, voltages d-q and simple voltage for the indirect control, the fig.11 shows the same characteristics for the direct control. The figs.11 and 12 represent step speed from 0 to 100 rd/s then application of a load torque of 10 Nm between 0.5s and 1.5s and finally an inversion of rotary motion at the moment 2s from a speed of 100 rd/s to -100 rd/s.

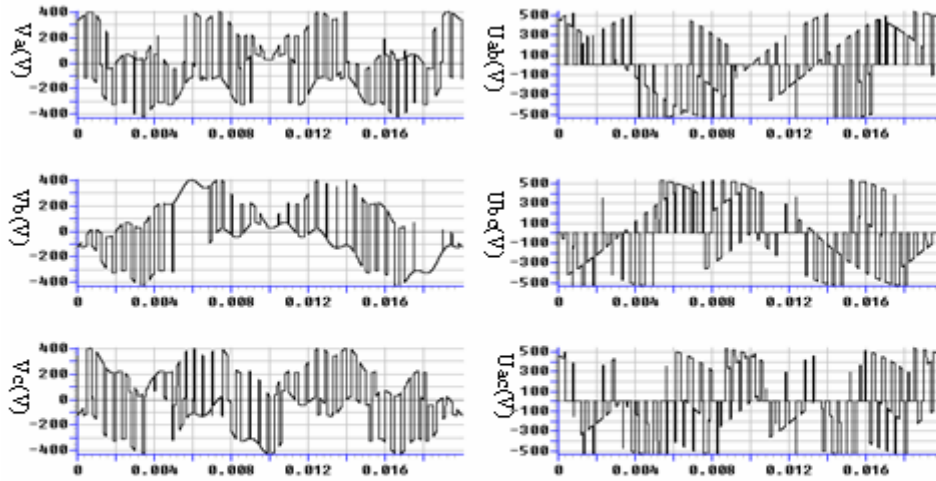


Fig. 5 voltages waveforms: V_a , V_b , V_c , u_{ab} , u_{bc} and u_{ac} , for: $f_s=2$ kHz, $f_i=50$ Hz, $f_o=200$ Hz, $q=0.866$

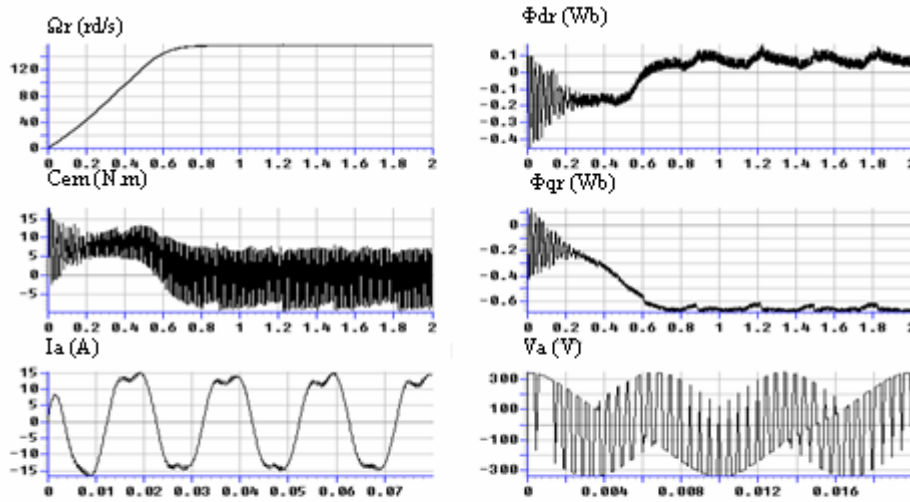


Fig. 6 Response of rotor speed, developed torque, phase current, simple voltage and rotor flux for: $f_o=50$ Hz, $q=0.866$

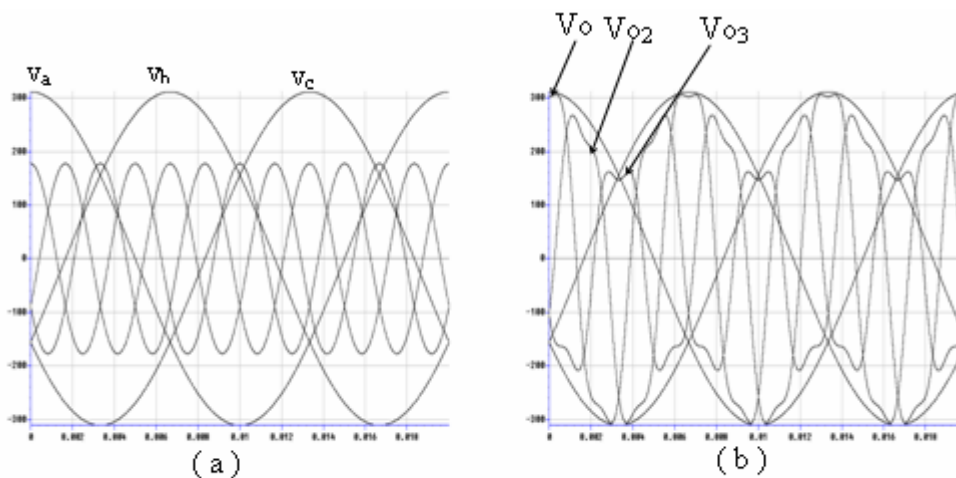


Fig. 7 Input simple voltages and desired output waveforms for : $f_s=2$ KHz, $f_o=200$ Hz
 (a): $q=0.57$
 (b): $q=0.866$

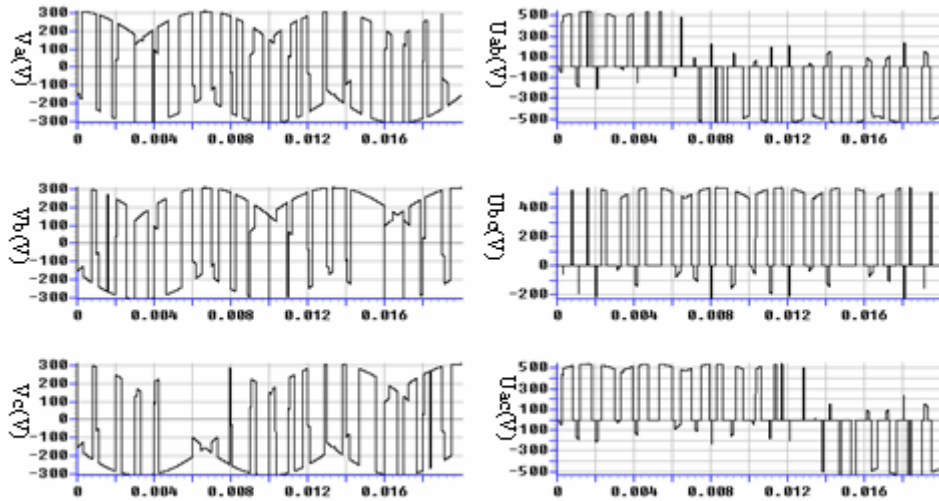


Fig. 8 voltages waveforms, v_a , v_b , v_c , u_{ab} , u_{bc} and u_{ac} for:
 $f_s=1\text{kHz}$, $f_i=50\text{ Hz}$, $f_o=25\text{ Hz}$, $q=0.57$

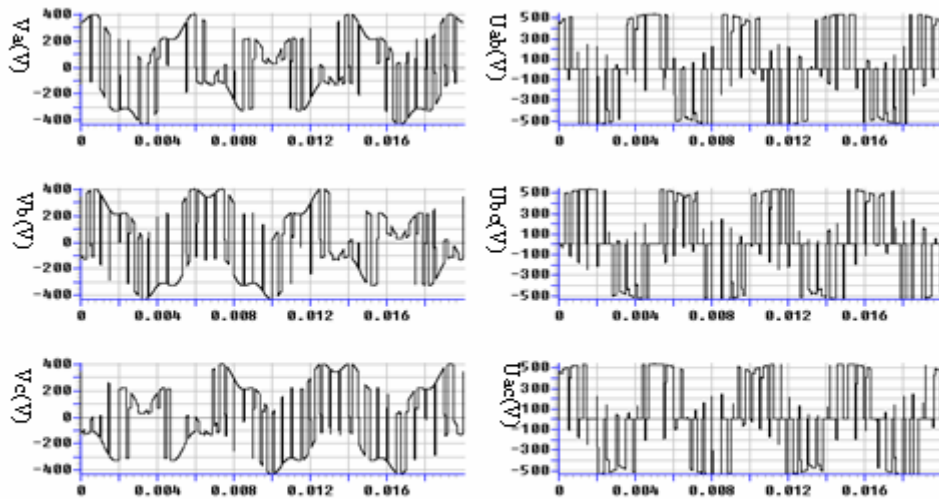


Fig. 9 voltages waveforms, v_a , v_b , v_c , u_{ab} , u_{bc} and u_{ac} ,
 for : $f_s=2\text{kHz}$, $f_i=50\text{ Hz}$, $f_o=200\text{ Hz}$, $q=0.866$

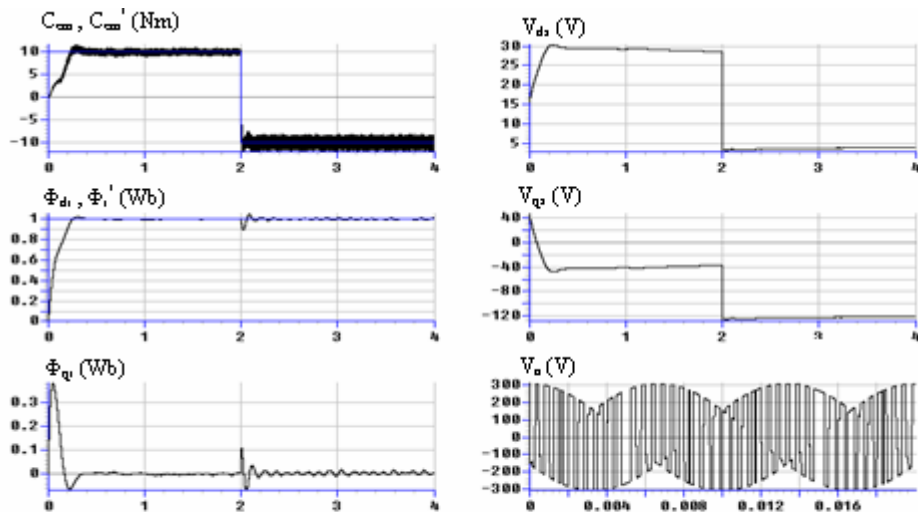


Fig.10 Decoupling flux - torque at $f_s=2\text{ kHz}$.

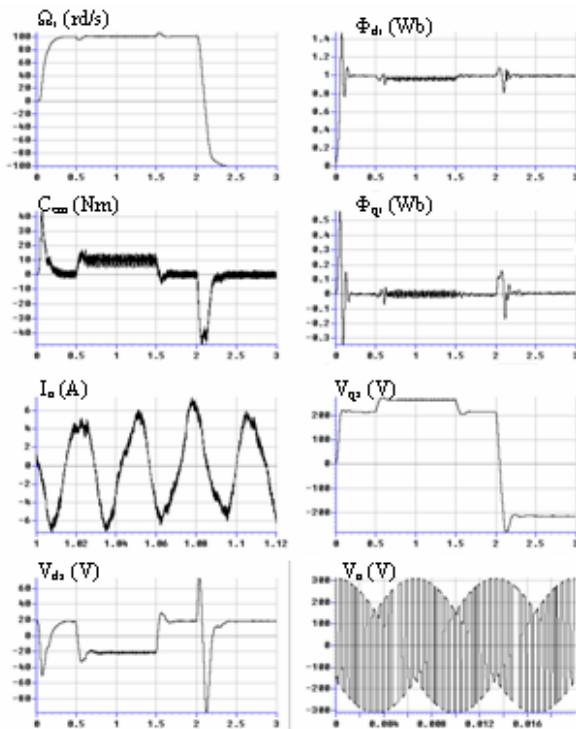


Fig. 11 Indirect vector control response ($f_s=2$ kHz).

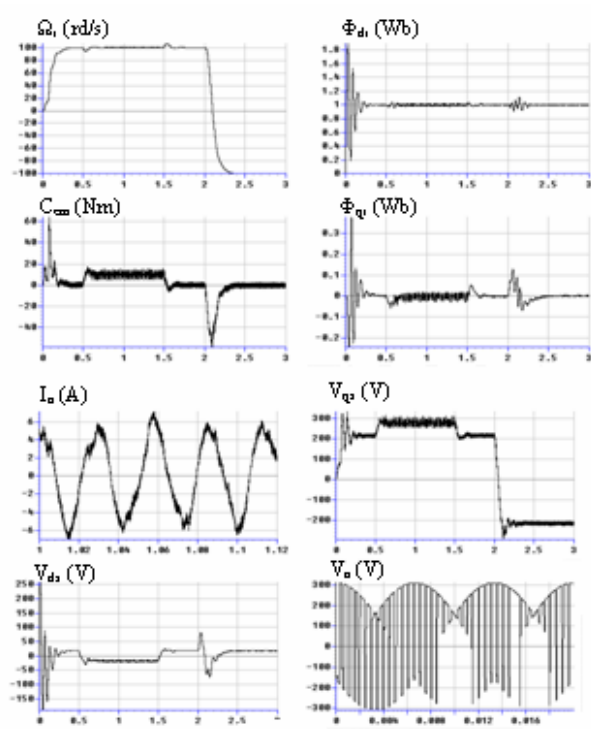


Fig. 12 Direct vector control response ($f_s=2$ kHz).

7. CONCLUSION

Vector control of induction machine fed from a matrix converter modeling and simulation have been described. The main topics discussed in the paper were: Review of matrix converter; switching angles calculation; converter modeling and simulation; direct and indirect field oriented control modeling and simulation. To our knowledge, this is the first time that a direct and indirect vector control of induction machine supplied from a matrix converter has been simulated, this being the main contribution of the paper. High performances of the direct vector control are achieved with the use of matrix converter. According to the simulation results obtained, the control algorithm presented is advisable for the establishment in the industrial drives. The next step of this research will be the realization of the motor drive.

APPENDIX

The parameters of the induction machine used are:
 $N_n=1420$ r/min , $f_n=50$ Hz, $P_n=1.5$ kW, 220/380 V,
 $3.64/6.31$ A, $R_s=4.850$ Ω , $R_r=3.805$ Ω , $L_s=L_r=0.274$
 H, $L_m=0.258$ H, $J=0.031$ kg.m², $k_f=0.001$ kg.m²/s.

REFERENCES

- [1] C. KLUMNER, I. BOLDEA: Artificial loading of induction machine using a matrix converter, IEE Power electronics, N° 475, pp. 40-45, 2000.
- [2] A. ALESINA, M. VENTURINI: Solid-state power conversion : a Fourier analysis approach to generalized transformer synthesis, IEEE

Trans. on circuit and systems, vol.cas-28, N°4, pp 319-330, 1981

- [3] A. ALESINA, M. VENTURINI: Intrinsic amplitude limits and optimum design of 9 switches direct PWM ac-ac converter, Proc.of PSEC con. pp.1284-1290, Rec.april 1988.
- [4] D.G. HOLMES, T.L. LIPO: Implementation of a controlled rectifier using ac-ac matrix converter thory, IEEE Trans. on power electronics, vol.7 N°1, pp.240-250, Jan.1992
- [5] ALESINA, M. VENTURINI: Analysis and design of optimum amplitude nine switch direct ac-ac converters, IEEE Trans. On Power Electronics, vol.4, N°1, pp.101-112, Jan. 1989.

BIOGRAPHIES

Abdelkader DJAHBAR was born in Chlef, Algeria, in 1970. He received the Master degree, with honor, in electrical engineering from the Polytechnical School of Algiers, Algeria, in 1998. In 1999, he joined the department of electrical engineering, university of chlef, where, from 2000 to 2005, he was a research assistant. His scientific work is related to electrical machines and drives and Power Electronics.

Benyounés MAZARI (1953) received the state engineer degree in electrical engineering in 1978 from the University of Sciences and Technology of Oran USTO Algeria, the M.Sc degree from the University of Colorado Boulder USA in 1981 and the Doctorat degree from the Institut National Polytechnique de Lorraine INPL Nancy France in 1992. Since 1982, he was at the USTO-Oran and from 1987-1992 he was on leave as a researcher at INPL (France). His scientific work is related to electrical machines and drives and Power Electronics.