A Matrix Converter for an Induction Motor Drive System

A. DJAHBAR 1                 B. MAZARI 1               N. MANSOUR 2
a_djahbar@yahoo.fr   mazari_dz@yahoo.fr   nmansour@eng.uob.bh

1 Laboratoire de Développement des Entraînements Electriques LDEE
University of Sciences and Technology
Faculty of Electrical Engineering
Oran, Algeria
2 University of Bahrain, Faculty of Engineering
Bahrain

Abstract
In this paper, a control strategy of scalar modulation with three intervals and vector control technique for matrix converter fed induction motor drive system is proposed. 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, three-phase Induction Machine, Modulation of VENTURINI, Scalar Modulation, Vector control.

1. Introduction
Matrix converter (MC) is a modern energy conversion device that has been developed over the last two decades [1], [2]. 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. 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 some well-known reasons: 1) implementation of switch devices in matrix converter is difficult; 2) 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, three strategies of modulation of matrix converter are used [3],[4]:

The first is based on algorithm proposed by VENTURINI and ALESINA [3], the second is based on technique of space vector modulation (SVM), and the last is the scalar modulation [5], [6]. 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].

Figure 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 [1]:

\[
\begin{align*}
\frac{di_d}{dt} &= \frac{1}{L_d} \left[ - (R_d + \frac{L_d}{L_m} R_r) i_d + w_L L_m i_q + \frac{L_m}{T_d} \phi_r + \omega_m \frac{L_m}{L_m} \phi_r + v_d \right] \\
\frac{di_q}{dt} &= \frac{1}{L_q} \left[ - w_L L_m i_d - (R_q + \frac{L_q}{L_m} R_r) i_q - \omega_m \frac{L_m}{L_m} \phi_r + \frac{L_m}{T_d} \phi_r + v_q \right] \\
\frac{d\phi_r}{dt} &= \frac{1}{T_d} i_d - \frac{1}{T_d} \phi_r + (w_q - p w_u) \phi_r \\
\frac{d\phi_q}{dt} &= \frac{1}{T_d} i_q - (w_q - p w_u) \phi_q - \frac{1}{T_d} \phi_q \\
\frac{dw}{dt} &= \frac{1}{J} (C_m - C_s - Bw_u)
\end{align*}
\]

The developed torque equation is given by:

\[
c_{em} = p \frac{L_m}{L_r} (\Phi_d r i_q s - \Phi_q r i_d s)
\]

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 \((q_\Phi = q_\Phi = 0)\). From this condition and eqns.1, the \(i_d\) reference can be computed in order to impose the flux \(\phi_r\). Furthermore, the position \(q_1\) of the rotating frame can be estimated using eqns.1.

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

\[
\begin{align*}
\frac{d\phi_d}{dt} &= \frac{R}{L_r} (L_r i_d - \Phi_d) \\
\frac{d\phi_q}{dt} &= \frac{R}{L_r} (L_r i_q - \Phi_q) \\
v_d &= R i_d + \sigma L_d \frac{di_d}{dt} + w_L \frac{L_m}{T_d} \phi_r - \sigma L_r w_i_r \\
v_q &= R i_q + \sigma L_d \frac{di_q}{dt} + (w_q - p w_u) \Phi_r + \sigma L_r w_i_q
\end{align*}
\]

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 \(i_q\) such as DC machine.

A decoupled model can be obtained by using two intermediate variables:

\[
\begin{align*}
v_{d_1} &= v_{d_2} + v_1 \\
v_{q_1} &= v_{q_2} + v_2
\end{align*}
\]

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

The stator voltages \((v_{d}, v_{q})\) are reconstructed from \((v_{d_1}, v_{q_1})\).

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 figure 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 ($s_{11}, s_{12}, ..., s_{33}$) 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 $f_i = \omega_0 / 2\pi$, through 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 $f_o = \omega_0 / 2\pi$. The input three-phase voltages of the converter are given by:

$$
\begin{bmatrix}
V_{i1} \\
V_{i2} \\
V_{i3}
\end{bmatrix} = V_{in} \begin{bmatrix}
\cos(\omega t) \\
\cos(\omega t - 2\pi / 3) \\
\cos(\omega t - 4\pi / 3)
\end{bmatrix}
$$

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

$$
\begin{bmatrix}
V_{o1} \\
V_{o2} \\
V_{o3}
\end{bmatrix} = V_{on} \begin{bmatrix}
\cos(\omega t) \\
\cos(\omega t - 2\pi / 3) \\
\cos(\omega t - 4\pi / 3)
\end{bmatrix}
$$

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 [7]:

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 ($\omega_0 = 2\pi f_s$) complies with $\omega_0 / 2\pi \gg \omega_0$ and $\omega_0$: in other words the switching frequency is much higher ($f_s \gg 20\max(\omega_0, f_0)$) than the input and output frequencies).

During the $k^{th}$ switching cycle $T_k$: (fig.3, the first phase output voltage is given by:

$$
v_{ol} = \begin{cases}
0 & \text{if } t < -Ts \\
m_1^k Ts & \text{if } t \geq -Ts \text{ and } t < -Ts + Ts \\
m_2^k Ts & \text{if } t \geq -Ts + Ts \text{ and } t < -Ts + 2Ts \\
m_3^k Ts & \text{if } t \geq -Ts + 2Ts \text{ and } t < -Ts + 3Ts \\
0 & \text{if } t \geq -Ts + 3Ts
\end{cases}
$$

Where $m^k_s$ are defined by:

$$
m_s^k = \frac{t_s}{T_s}
$$

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

The $m_s^k$ have the physical meaning of duty cycle. Also,

$$
\sum_{m_{ij}} = m_{i}^k + m_{j}^k + m_{k}^k = 1 \quad \text{and} \quad 0 < m_{ij}^k < 1
$$

Which means that during every cycle $T_k$ 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 $v_{ol}$ and the input currents $ij$ 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:

$$
\begin{align*}
V_{ol}^{(k)} &= V_{o1}^{(k)} \frac{f_{11}}{Ts} + V_{o2}^{(k)} \frac{f_{12}}{Ts} + V_{o3}^{(k)} \frac{f_{13}}{Ts} \\
V_{o2}^{(k)} &= V_{o1}^{(k)} \frac{f_{21}}{Ts} + V_{o2}^{(k)} \frac{f_{22}}{Ts} + V_{o3}^{(k)} \frac{f_{23}}{Ts} \\
V_{o3}^{(k)} &= V_{o1}^{(k)} \frac{f_{31}}{Ts} + V_{o2}^{(k)} \frac{f_{32}}{Ts} + V_{o3}^{(k)} \frac{f_{33}}{Ts} 
\end{align*}
$$

If times of conduction are modelled in the shape of sinusoid with the pulsation $\omega_o$ while $T_s$ remains constant, such as $w_o=2\pi w_n$, these times are defined as follows:

1. For the 1st phase, we have:

$$
\begin{align*}
t_{11} &= \frac{T}{3} (1 + 2q \cos(w_n t + \theta)) \\
t_{12} &= \frac{T}{3} (1 + 2q \cos(w_n t + \theta - \frac{2\pi}{3})) \\
t_{13} &= \frac{T}{3} (1 + 2q \cos(w_n t + \theta - \frac{4\pi}{3}))
\end{align*}
$$
2. For the 2nd phase:

\[
t_{21} = \frac{T}{3}(1 + 2q \cos(w_m t + \theta - \frac{4\pi}{3})) \\
t_{22} = \frac{T}{3}(1 + 2q \cos(w_m t + \theta)) \\
t_{23} = \frac{T}{3}(1 + 2q \cos(w_m t + \theta - \frac{2\pi}{3}))
\] (12)

3. For the 3rd phase:

\[
t_{31} = \frac{T}{3}(1 + 2q \cos(w_m t + \theta - \frac{2\pi}{3})) \\
t_{32} = \frac{T}{3}(1 + 2q \cos(w_m t + \theta - \frac{4\pi}{3})) \\
t_{33} = \frac{T}{3}(1 + 2q \cos(w_m t + \theta))
\] (13)

Where \( \theta \) is initial phase angle.

The output voltage is given by [7]:

\[
V_{o(i)} = [M^{(i)}]V_{i(i)}
\] (14)

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

Where:

\[
\begin{align*}
V_o & = V_o(m) + jV_o(l) + jV_o(m) \\
V_i & = V_i(o) - V_i(i)
\end{align*}
\]

The computation of conduction times \( t_k, t_l, t_m \) for a transfer ratio \( q \leq 0.57 \) 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 [3], [7]:

\[
[1 + \rho \cos w_i t + \rho \cos (3w_i t)] - [\rho \cos (3w_o t)] = 0
\] (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 \( t_k, t_l, t_m \) are selected such as:

\[
\frac{t_k - t_l}{t_l} = \frac{V_o - V_m}{\rho V_l - (1 + \rho)V_m}
\] (17)

The 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 \( f_s \).
Figure 5. Voltages waveforms: \( v_a, v_b, v_c, u_{ab}, u_{bc} \) and \( u_{ac} \) for: \( f_s=2 \, \text{kHz}, f_i=50 \, \text{Hz}, f_o=200 \, \text{Hz}, q=0.866 \)

Simulation results of the starting process of an induction motor fed from a matrix converter for a frequency \( f_s=2 \, \text{kHz} \) are shown on the figure below.

6.2. Performances of Scalar strategy

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

Figure 7. Input simple voltages and desired output waveforms for: \( f_s=2 \, \text{kHz}, f_i=50 \, \text{Hz}, f_o=200 \, \text{Hz} \)

(a): \( q=0.57 \), (b): \( q=0.866 \)

The simulation of the vector control of an IM associated with a MC, 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.

Figure 10. Decoupling flux-torque at \( f_s=2 \, \text{kHz} \).
7. Conclusion

Vector control of induction machine fed from a three phase 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.

8. 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=0.258H
J=0.031 kg.m², k_f=0.001136 kg.m²/s.

9. References