Predictive Direct Torque and Flux Control of an Induction Motor Drive fed by a Direct Matrix Converter with Reactive Power Minimization

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***Abstract*— The paper presents a Predictive Direct Torque and Flux Control (PDTFC) of an induction machine fed by a Direct Matrix Converter (DMC). The method combines the merits of Finite States Model Predictive Control (FSMPC) with the ones of DTC control. The proposed control algorithm selects the switching state of the DMC that minimizes the error between torque and flux predictions to their computed values for all different voltage vectors. The optimal voltage vector that minimizes a cost function is then applied to the terminal of the induction machine. Moreover, the proposed predictive control is easily extended to minimize the reactive power in the voltage source side. The control method uses only one sample time and it is very intuitive since it is simple, multi-objective and provides best performances compared to other control laws (fast dynamic response, simple implementation).**

***Index Terms*—Induction motor, predictive control, finite states-space model, direct matrix converter, cost function, reactive power.**

I. INTRODUCTION

Model Predictive Control (MPC) is a powerful control strategy that uses the model of the system to precalculate the behavior of the system for a predefined horizon in the future. A cost function evaluates the precalculated results and determines the optimal future control actions. Generalized Predictive Control (GPC) is the most popular method of MPC family methods since it can be applied to a great variety of systems where dead times can be easily compensated, the concept is intuitive and easy to understand, the multivariable case can be easily considered, easy inclusion of non-linearities in the model. The main disadvantages of the GPC control is the large amount of calculations, compared to classic controllers and the direct influence of the model on the quality of the resulting controller.

Since power converters have a discrete nature, Finite-States Model Predictive Control (FS-MPC) appears as an attractive alternative and offers a completely different and powerful approach to control power converters due to its fast dynamic response, no need for linear controllers in inner loops, no need for modulator (as in PWM or SVM modulation), completely different approach compared to PWM and SVM modulations, extremely simple, very good performance and can be implemented with standard commercial microprocessors. The method is based on the fact that a finite number of possible switching states can be generated by power converter (7 states for a two-level three- phase inverter, 27 states for a three-level, 64 states for a four-level VSI,…) and that the model of the system can be used to predict the behavior of the variables for each switching state.

For the selection of the appropriate switching state to be applied to the system a quality function must be defined. The cost function is then evaluated for the predicted values on each sampling interval and the optimal switching state that minimizes the quality function is selected to apply during the next sampling interval [1][2][3][6].

Firstly introduced in 1976, the Matrix Converter is a direct AC–AC converter that uses an array of (m x n) controlled bi-directional switches to directly connect m-phase inputs to n-phase outputs. Direct Matrix Converters (DMC) have recently received a considerable attention these last years because of their numerous merits on the traditional AC-DC-AC converters such as no DC-link capacitor, the bi-directional power flow control (the capability of regeneration), the sinusoidal input-output waveforms and adjustable input power factor, but the biggest drawback of this technology is the high control complexity [3].

The present paper presents an improved method combining the benefits of (FSMPC) and those of (DMC) converter to control torque and flux of an induction motor with reactive power minimization in the input system of the converter. The method provides best performances in term of rapid and precise dynamic torque response and reduces ripples on torque and currents than in DTC control with only simple hardware (no need for hysteresis controllers as in DTC control, no need for modulators as in PWM or SVM).

II. DIRECT MATRIX CONVERTER

Despite some drawbacks such as high number of power semiconductor devices, the limitation of maximum load voltage to 86% of the supply voltage, no need for energy storage element, the matrix converters have received recently a wide attention especially in motion control. The three-phase to three-phase matrix converter has been extensively researched due to its potential as a replacement for the traditional AC-DC-AC converter in AC motor drives for the following benefits [3]:

- Adjustable input displacement factor, irrespective of the load
- The capability of regeneration (Four-quadrant operation)
- High quality input and output waveforms
- The lack of bulky and limited lifetime energy storage components, such as electrolytic capacitors.

Matrix converter topologies can be divided into two types: the direct matrix converter and the indirect matrix converter. As shown in Fig.1, the Indirect Matrix Converter IMC consists of a four-quadrant current source rectifier and a two-level voltage source inverter. This IMC converter topology is preferred in some

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applications due to its simpler and safer commutation of switches, also the control is simpler and less complex than in DMC converter.

The DMC converter (see fig.2.) is based on bi-directional switches and replaces the rectifier, inverter and energy storage element of an AC-DC-AC converter in only one stage thereby reducing the size of the conversion chain but increasing the control complexity.

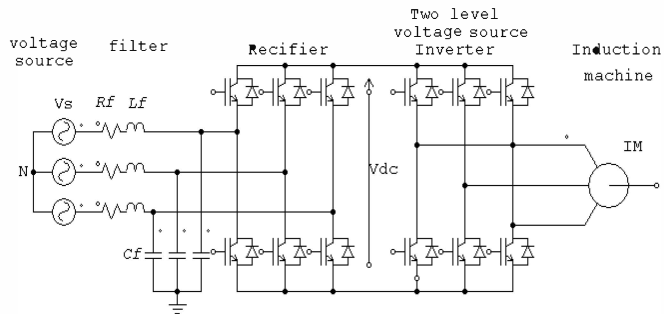

Fig 1.: Indirect Matrix Converter (IMC) power topology.

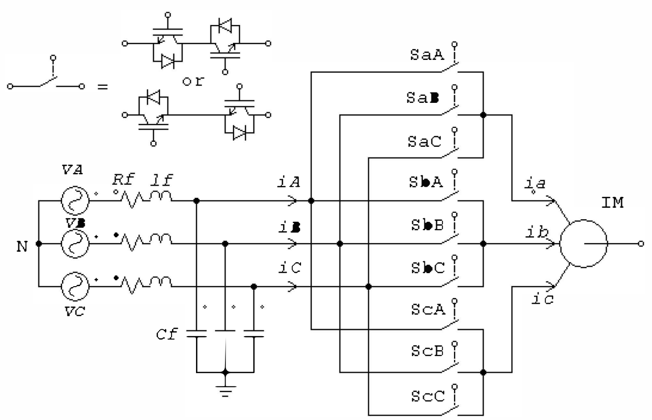

Fig 2.: Direct Matrix Converter (DMC) power topology.

As shown in Fig. 2., the matrix converter allows any input line to be connected to any output line for any given length of time. Due to the direct connection with voltage sources, the input lines must never be shorted. If the switches cause a short circuit between the input voltage sources, infinite current flows through the switches and damages the circuit. Also, due to the inductive nature of typical loads, the output terminals must not be open-circuited. If any output terminal is open-circuited, the voltage across the inductor (and consequently across the switches) is infinite and switches will be damaged due to the over-voltage. As a result, switches for each output phase must be controlled based on the following expression [3]:

$$S_{jA} + S_{jB} + S_{jC} = 1 \quad , \qquad j \in \{a,b,c\} \tag{1}$$

where S_{jk} is the switching function of a bi-directional switch, which is defined as:

$$S_{jk} = \begin{cases} 1 & S_{jk} \ \ closed \\ 0 & S_{jk} \ \ opened \end{cases} \qquad k \in \{A,B,C\} \tag{2}$$

For a three-phase to three-phase matrix converter there are twenty-seven valid switching combinations available for generating specific input phase currents and output voltages that are presented in Table 1. The inverter output voltage vector is kept constant during the switching period, so the inverter current and, hence, the motor currents can be controlled by choosing the appropriate voltage vector. The instantaneous values of the output voltages and the input currents generated by each switching combination can be determined using the instantaneous transfer matrices, where T_{LL} is the instantaneous input phase to output line-to-line transfer matrix and T_{Ph} is the instantaneous input phase to output phase matrix:

$$T_{LL} = \begin{pmatrix} S_{aA}-S_{bA} & S_{aB}-S_{bB} & S_{aC}-S_{bC} \\ S_{bA}-S_{cA} & S_{bB}-S_{cB} & S_{bC}-S_{cC} \\ S_{cA}-S_{aA} & S_{cB}-S_{aB} & S_{cC}-S_{aC} \end{pmatrix}, T_{ph} = \begin{pmatrix} S_{aA} & S_{aB} & S_{aC} \\ S_{bA} & S_{bB} & S_{bC} \\ S_{cA} & S_{cB} & S_{cC} \end{pmatrix} \tag{3}$$

Table. 1: Valid switching combinations for a matrix converter

N	S_{aA}	S_{bA}	S_{cA}	S_{aB}	S_{bB}	S_{cB}	S_{aC}	S_{bC}	S_{cC}
1	1	0	0	1	0	0	1	0	0
2	0	1	0	0	1	0	0	1	0
3	0	0	1	0	0	1	0	0	1
4	1	0	0	0	0	1	0	0	1
5	0	1	0	0	0	1	0	0	1
6	0	1	0	1	0	0	1	0	0
7	0	0	1	1	0	0	1	0	0
8	0	0	1	0	1	0	0	1	0
9	1	0	0	0	1	0	0	1	0
10	0	0	1	1	0	0	0	0	1
11	0	0	1	0	1	0	0	0	1
12	1	0	0	0	1	0	1	0	0
13	1	0	0	0	0	1	1	0	0
14	0	1	0	0	0	1	0	1	0
15	0	1	0	1	0	0	0	1	0
16	0	0	1	0	0	1	1	0	0
17	0	0	1	0	0	1	0	1	0
18	1	0	0	1	0	0	0	1	0
19	1	0	0	1	0	0	0	0	1
20	0	1	0	0	1	0	0	0	1
21	0	1	0	0	1	0	1	0	0
22	1	0	0	0	1	0	0	0	1
23	1	0	0	0	0	1	0	1	0
24	0	1	0	1	0	0	0	0	1
25	0	1	0	0	0	1	1	0	0
26	0	0	1	1	0	0	0	1	0
27	0	0	1	0	1	0	1	0	0

Based on the transfer matrix T_{LL}, the instantaneous output line-to-line voltages and the input phase currents can be determined, as given below:

$$V_{oll} = \begin{bmatrix} V_{ab} & V_{bc} & V_{ca} \end{bmatrix}^T = T_{LL}.\begin{bmatrix} V_A & V_B & V_C \end{bmatrix}^T = T_{LL}.V_i \tag{4}$$

$$i_{iph} = \begin{bmatrix} i_A & i_B & i_C \end{bmatrix}^T = T_{LL}^t.\begin{bmatrix} i_{ab} & i_{bc} & i_{ca} \end{bmatrix}^T = T_{LL}^t.i_{oll} \tag{5}$$

where T_{LL}^t is the transpose of T_{LL}; V_{ab}, V_{bc} and V_{ca} are the output line-to-line voltages; V_A, V_B and V_C are the input phase voltages; i_{ab}, i_{bc} and i_{ca} are the output line-to-line currents; i_A, i_B and i_C are the input phase currents. Alternatively, by using the transfer matrix T_{Ph}, the instantaneous output line-to-supply neutral voltages (Va, Vb and Vc) and the input phase currents can be found as:

$$V_{oph} = (V_a \ V_b \ V_c)^t = T_{ph}(V_A \ V_B \ V_C)^t = T_{ph}V_i$$
$$i_{iph} = (i_A \ i_B \ i_C)^t = T_{ph}{}^t (i_a \ i_b \ i_c)^t = T_{ph}{}^t i_{oph} \quad (6)$$

These 27 switching states are classified into three groups, according to the kind of output voltage and input current vector that each switching state generates [3]:

1) Zero space vector: all three output phases connected to the same input phase. Switching states from this group generate a space vector with amplitude zero.
2) Stationary space vectors: two output phases connected to a common input phase, and the third connected to a different input phase. This group generates stationary space vectors with varying amplitude and fixed direction.
3) Rotating space vectors: each output phase connected to a different input phase. Vectors have constant amplitude, but its angle varies at the supply angular frequency.

III. INDUCTION MOTOR MODEL

A squirrel cage induction motor model fed by a DMC converter is used under simplified assumptions where iron saturation, skin effect, heating variations of stator and rotor resistances are neglected. The general model is expressed in the stator fixed (α-β) reference frame where outputs are stator currents and fluxes as:

$$\begin{cases} V_{s\alpha} = R_s i_{s\alpha} + \dfrac{d\phi_{s\alpha}}{dt} \\ V_{s\beta} = R_s i_{s\beta} + \dfrac{d\phi_{s\beta}}{dt} \\ 0 = R_r i_{r\alpha} + \dfrac{d\phi_{r\alpha}}{dt} + \omega\phi_{r\beta} \\ 0 = R_r i_{r\beta} + \dfrac{d\phi_{r\beta}}{dt} - \omega\phi_{r\alpha} \end{cases} \quad (7)$$

The state space model is as follows:

$$\dot{X} = AX + BU, \ X = \begin{bmatrix} i_{s\alpha} & i_{s\beta} & \phi_{r\alpha} & \phi_{r\beta} & \Omega \end{bmatrix}^T, \ U = [V_{s\alpha} \ V_{s\beta}]^T$$

$$\begin{cases} \dfrac{di_{s\alpha}}{dt} = -\gamma i_{s\alpha} + \dfrac{\kappa}{T_r}\phi_{r\alpha} + \omega\,\kappa\,\phi_{r\beta} + \dfrac{V_{s\alpha}}{\sigma L_s} \\ \dfrac{di_{s\beta}}{dt} = -\gamma i_{s\beta} - \omega\,\kappa\,\phi_{r\alpha} + \dfrac{\kappa}{T_r}\phi_{r\beta} + \dfrac{V_{s\beta}}{\sigma L_s} \\ \dfrac{d\phi_{r\alpha}}{dt} = \dfrac{L_m}{T_r} i_{s\alpha} - \dfrac{1}{T_r}\phi_{r\alpha} - \omega\,\phi_{r\beta} \\ \dfrac{d\phi_{r\beta}}{dt} = \dfrac{L_m}{T_r} i_{s\beta} + \omega\,\phi_{r\alpha} + \dfrac{1}{T_r}\phi_{r\beta} \end{cases} \quad (8)$$

The electromagnetic torque can be derived as:

$$T = p\left(\phi_{s\alpha} i_{s\beta} - \phi_{s\beta} i_{s\alpha}\right) \quad (9)$$

The mechanical equation is given by:

$$J\frac{d\Omega}{dt} = T - T_l - f\Omega \quad (10)$$

$$A = \begin{bmatrix} -\gamma & 0 & \dfrac{\kappa}{T_r} & \omega\kappa \\ 0 & -\gamma & -\omega\kappa & \dfrac{\kappa}{T_r} \\ \dfrac{L_m}{T_r} & 0 & -\dfrac{1}{T_r} & -\omega \\ 0 & \dfrac{L_m}{T_r} & \omega & \dfrac{1}{T_r} \end{bmatrix}, \ B = \begin{pmatrix} \dfrac{1}{\sigma L_s} & 0 & 0 & 0 & 0 \\ 0 & \dfrac{1}{\sigma L_s} & 0 & 0 & 0 \end{pmatrix}^T$$

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

Where (R_s, R_r) are respectively stator and rotor resistance per phase, $\left(i_{s\alpha}, i_{s\beta}, i_{r\alpha}, i_{r\beta}\right)$ are stator and rotor current vectors components, $\left(\phi_{s\alpha}, \phi_{s\beta}, \phi_{r\alpha}, \phi_{r\beta}\right)$ are stator and rotor flux vectors components respectively, (L_s, L_r, L_m) are stator, rotor and mutual inductances, $\Omega = \dfrac{\omega}{p}$ is the mechanical rotor speed, T is the electromagnetic torque of the machine and p is the number of pair poles.

IV. PREDICTIVE TORQUE AND FLUX CONTROL

Based on a given stator component voltage vector $V_{si}(k)$, measured current $i_s(k)$ and estimated stator flux $\phi_s(k)$ at current sampling instant, it is possible to obtain one step ahead prediction of stator current $i_s(k+1)$ and stator flux $\phi_s(k+1)$. Also, using (9) it is possible to predict the machine torque $T(k+1)$ for this voltage vector $V_{si}(k)$ where $V_s = \left[V_1, ...V_{27}\right]$. The predicted values of torque and stator flux are used to evaluate a cost function F that minimizes the quadratic error between predicted values and their references and the switching state that produces the minimum value of this cost function is selected to be applied on machine terminals in the next sampling time according to receding horizon control. Assuming that it is possible to define a first order approximation for the derivatives due to the first order nature of the state equations of induction motor model, we can write that:

$$\dot{x} = \frac{x(k+1) - x(k)}{T_s} \quad (11)$$

where T_s is the sampling period. For one step ahead prediction of stator flux, stator current and torque can be made based on previous standard estimation as:

$$\phi_{s\alpha}(k+1) = \phi_{s\alpha}(k) + T_s\ V_{s\alpha}(k+1) - R_s\ T_s\ i_{s\alpha}(k)$$
$$\phi_{s\beta}(k+1) = \phi_{s\beta}(k) + T_s\ V_{s\beta}(k+1) - R_s\ T_s\ i_{s\beta}(k)$$
$$i_{s\alpha}(k+1) = \left(1 - \gamma T_s\right) i_{s\alpha}(k) + T_s\left(\frac{\kappa}{Tr}\phi_{r\alpha}(k) + \omega\kappa\phi_{r\beta}(k) + \frac{V_{s\alpha}(k+1)}{\sigma L_s}\right)$$
$$i_{s\beta}(k+1) = \left(1 - \gamma T_s\right) i_{s\beta}(k) + T_s\left(-\omega\kappa\phi_{r\alpha}(k) + \frac{\kappa}{Tr}\phi_{r\beta}(k) + \frac{V_{s\beta}(k+1)}{\sigma L_s}\right)$$
$$T(k+1) = p\left(\phi_{s\alpha}(k+1) i_{s\beta}(k+1) - \phi_{s\beta}(k+1) i_{s\alpha}(k+1)\right) \quad (12)$$

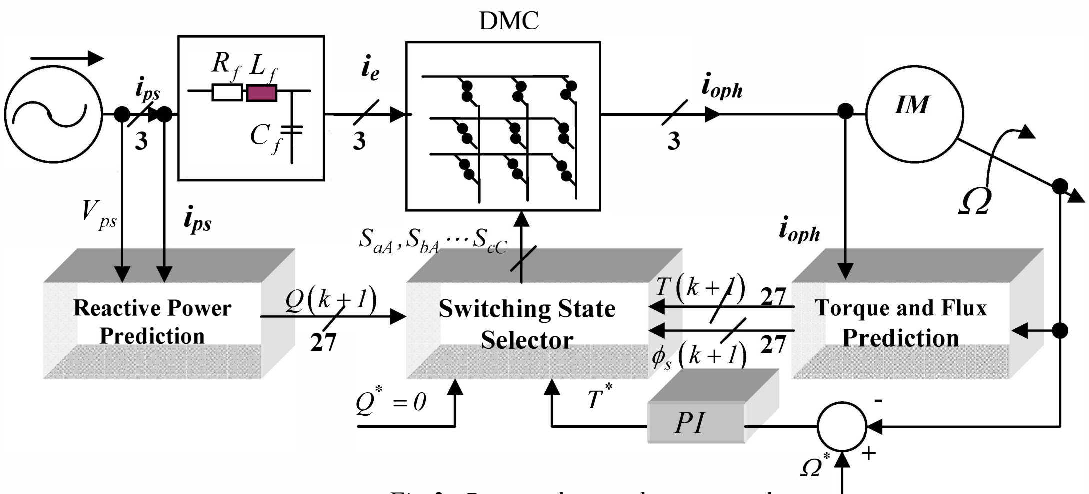

Fig.3 : Proposed control strategy scheme

A. Discrete System Model

In order to predict the behavior of the input current, an adequate model of the input filter, the load and the converter is indispensable. According to the Fig.3., the input filter can be represented by the following equation system [4][6]:

$$v_{ps}(t) = R_f i_{ps}(t) + L_f \frac{di_{ps}(t)}{dt} + v_e(t) \tag{13}$$

$$i_{ps}(t) = C_f \frac{dv_e(t)}{dt} + i_e(t) \tag{14}$$

Where v_e is the capacitance voltage, then a discrete-time form of the input side for a sampling time T_s can be employed to calculate the future value of the input currents considering the voltages and currents measurements at the k^{th} sampling time. The input side can be represented by a state space model with the variables i_{ps} and v_e:

$$\begin{pmatrix} \dot{v}_e \\ \dot{i}_{ps} \end{pmatrix} = A \begin{pmatrix} v_e \\ i_{ps} \end{pmatrix} + B \begin{pmatrix} v_{ps} \\ i_e \end{pmatrix} \tag{15}$$

where

$$A = \begin{pmatrix} 0 & 1/C_f \\ -1/L_f & \dfrac{-R_f}{L_f} \end{pmatrix}, \quad B = \begin{pmatrix} 0 & -1/C_f \\ 1/L_f & 0 \end{pmatrix} \tag{16}$$

Then, the discrete time state-space model is obtained as follows:

$$\begin{pmatrix} v_e(k+1) \\ i_{ps}(k+1) \end{pmatrix} = D \begin{pmatrix} ve(k) \\ i_{ps}(k) \end{pmatrix} + \Psi \begin{pmatrix} v_{ps}(k) \\ i_e(k) \end{pmatrix} \tag{17}$$

Where [6]:

$$D = \begin{pmatrix} D_{11} & D_{12} \\ D_{21} & D_{22} \end{pmatrix} = e^{AT_s}$$
$$\Psi = \begin{pmatrix} \Psi_{11} & \Psi_{12} \\ \Psi_{21} & \Psi_{22} \end{pmatrix} = A^{-1}(D - I_{2*2})B \tag{18}$$

In this way, the input current and capacitor voltage can be easily derived by:

$$i_{ps}(k+1) = D_{21} v_e(k) + D_{22} i_{ps}(k) + \Psi_{21} v_{ps}(k) + \Psi_{22} i_e(k) \tag{19}$$

$$v_e(k+1) = D_{11} v_e(k) + D_{12} i_{ps}(k) + \Psi_{11} v_{ps}(k) + \Psi_{12} i_e(k) \tag{20}$$

The instantaneous reactive input power can be predicted, based on predictions of the input voltage and current as:

$$Q(k+1) = v_{ps\beta}(k+1) i_{ps\alpha}(k+1) - v_{ps\alpha}(k+1) i_{ps\beta}(k+1) \tag{21}$$

Finally, the equations (19)-(20)-(21) are calculated for each valid switching state and used to evaluate the cost function *F*.

B. Cost function

The predictive torque and flux control scheme is given by Figure 3. The objective control is to get high performances in term of rapid and precise dynamic torque and flux responses as in DTC control by using a quadratic cost function that minimizes the error between reference torque and flux to their computed values.
The predictions on flux and torque are used to evaluate the impact of every voltage vector on motor torque and stator flux. The reference torque is generated from the external speed control loop via a simple PI controller while the flux reference is kept constant to its nominal value for normal speed operation as it is given by fig.3. The cost function is formulated as in [4] as:

$$F = \lambda \frac{\left(T^*(k+1) - T(k+1)\right)^2}{T_n^2} + \mu \frac{\left(\phi^*_s(k+1) - \phi_s(k+1)\right)^2}{\phi_{sn}^2} \tag{22}$$

where T_n and ϕ_{sn} are the nominal torque and nominal stator flux values. λ, μ are weighted factors as in GPC control.

One of the most benefits of DMC converter is the possibility to control the displacement factor in the supply voltage side by minimizing the input reactive power. Multiple objectives can be achieved at the same time by adding more functions in the global cost function *F* as [7]:

$$F = \lambda \frac{\left(T^*(k+1) - T(k+1)\right)^2}{T_n^2} + \mu \frac{\left(\phi^*_s(k+1) - \phi_s(k+1)\right)^2}{\phi_{sn}^2} + \gamma \left| Q^*(k+1) - Q(k+1) \right| \tag{23}$$

where λ, μ, γ are the weight coefficients which denotes the priority in the control.

For each stator voltage vector available, the cost function F is evaluated, and the stator voltage producing the minimum value of F is selected to be applied on motor terminals.

The predictive controller is summarized in the next steps:

1)- Measure of mechanical speed and stator currents $i_s(k),\Omega(k)$

2)- These measurements are used to predict torque and stator flux values $T(k+1)$, $\phi_s(k+1)$ for all twenty seven different voltage vectors by using relation (12)

- The input voltages $v_{ps}(k)$, $v_e(k)$, input currents $i_{ps}(k)$ are measured for predict reactive power $Q(k+1)$ for all twenty seven different voltage vectors by using relation (21)

3)- The twenty seven prediction are evaluated using the cost function F by using relation (22) or (23).

4)- The optimal switching state that corresponds to the optimal voltage vector that minimizes the cost function is selected to be applied on terminal machine in the next sampling time.

All the steps cited above are repeated each sampling time accounting for the new references and measurements in accordance with the receding horizon control.

IV. SIMULATION RESULTS

The proposed predictive control is tested via simulation on the bench of figure 3 with a sampling time of 30 μs .The drive is tested in a four quadrant (4Q) operational capability where the machine is running in the steady state at 50 rd/s with 20 N.m, then a speed reversal setpoint of –50 rad/s is applied at 0.3s. The profile of the load torque, mechanical speed, electromagnetic torque, stator flux magnitude, load current and output voltage are visualized throw fig. 4. and fig.5.

The first test is used to control only the torque and flux without minimization of the reactive power ($\lambda = 1000,\ \mu=15000,\ \gamma=0$).

Fig. 4 shows that the torque presents a very good dynamic and precise response as in the case of DTC control and it is completely decoupled from the stator flux that is kept constant at all times as can be seen in fig.5 during dynamic transitions. The load currents appear highly sinusoidal, although no current controllers are used in the control algorithm.

Fig. 6 illustrates the waveforms of the reactive power, input current and supply voltage without reactive power minimization by using relation (22) to control only torque and flux. The input current presents high distortion and different phase with the related input voltage. The ripple and distortion observed in the input current change during the speed reversal performed by the drive. The next test performed when a reactive power minimization is included in the cost function of (23) ($\alpha = 1000,\ \beta=15000,\ \lambda=0.0365$).the reactive power reference is set to 0 in order to make input current in phase with the supply voltage. The tracking performances of speed, torque, flux, load current and output voltage are similar to those of fig.4 and fig.5.

On fig.7. one can see that sinusoidal input current in phase with supply voltage is achieved during motoring operation of the drive. When the drive operates to regenerate energy (regenerating mode), it is also observed that sinusoidal input current is also achieved with π rad phase shift with the phase voltage, making the energy flow from the motor to the mains (regenerating) possible. The main parameters used in simulation are listed in tab.2.

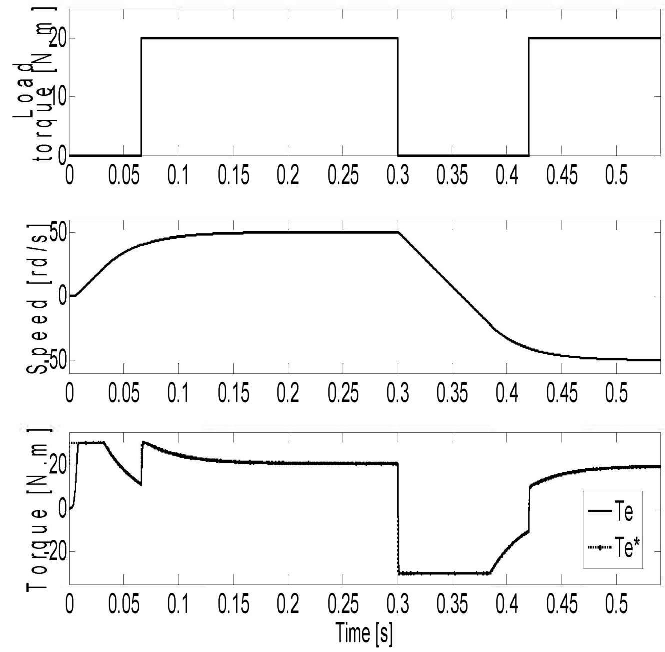

Fig. 4: Performances of rotor speed and torque with/without reactive power minimization

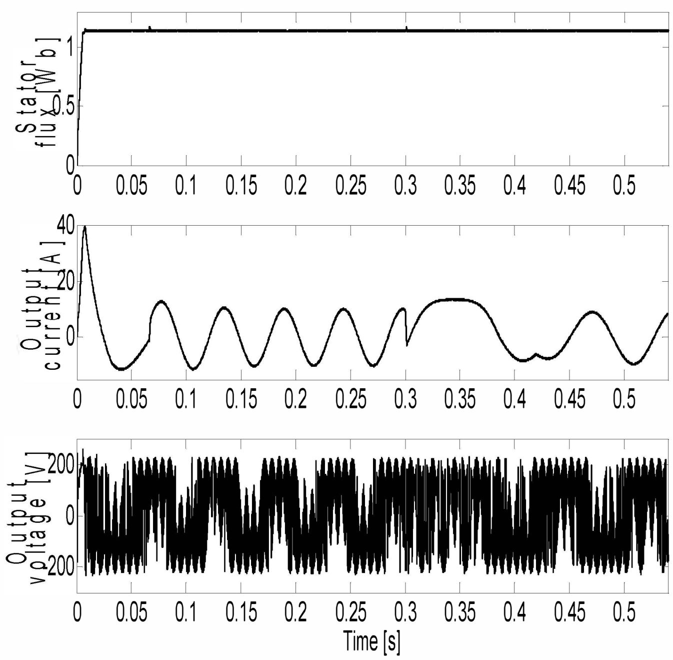

Fig. 5: Stator flux magnitude, load current and load voltage profile with/without reactive power minimization

Table 2: Main circuit parameters

C_f , L_f, R_f	$90\ \mu F$, $400\ \mu H$, $0.5\ \Omega$
Ts(sampling time)	$30\ \mu s$
Weighting factors	$\alpha = 1000,\ \beta=15000,\ \lambda=0$ or 0.0365
Machine parameters	T_n=30N.m; ϕ_{sn}=1.14Wb; J=0.035(USI); R_r=1.83 Ω ; R_s=0.97 Ω ; lr=0.165H; ls=0.161H; p=2; ff=0.01(USI); Tr=lr/Rr; M=0.154H

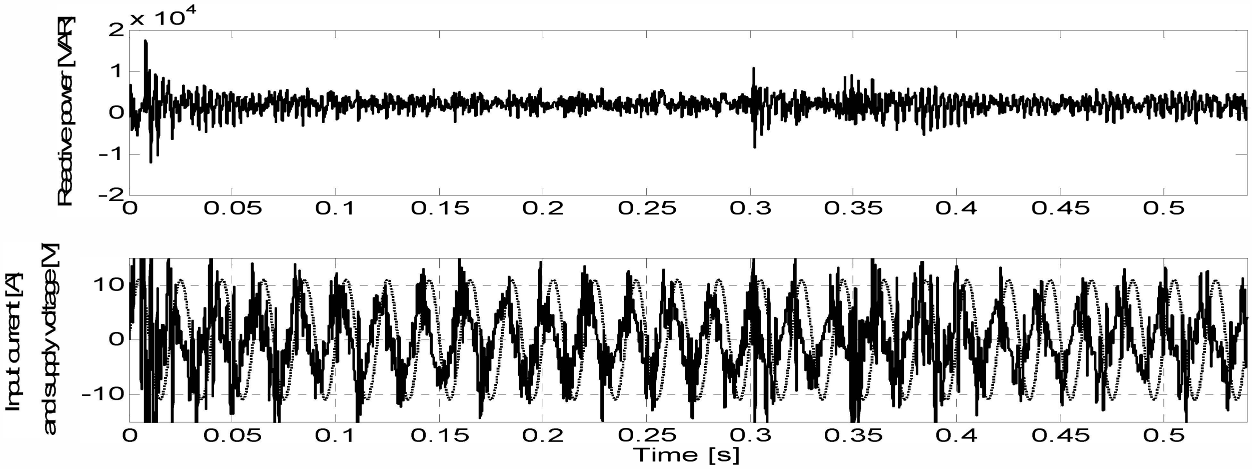

Fig.6: Reactive power, input current and supply voltage without reactive power minimization.

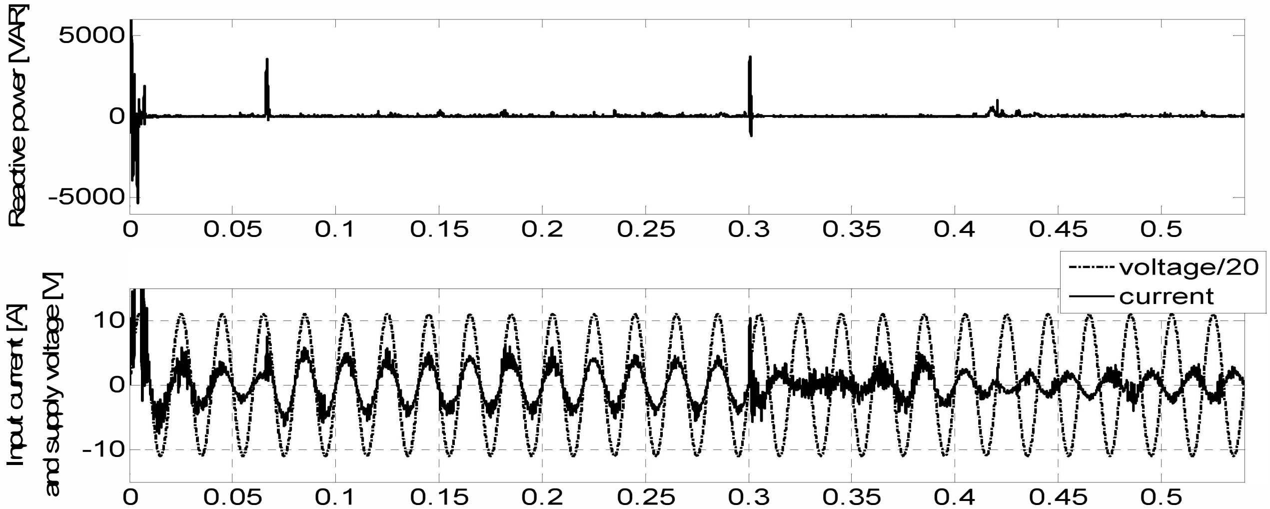

Fig.7: Reactive power, input current and supply voltage with reactive power minimization.

VI. CONCLUSION

A finite states predictive torque and flux control strategy with reactive power minimization is applied to a direct matrix converter. The most advantage of the proposed control is its simplicity in implementation, since the method avoids the use of any linear or nonlinear controllers except for the external speed loop and there is no need for any type of modulator such as in PWM or SVM modulation. This can reduce the overall cost of the drive system.

The control scheme is very simple and uses discrete model of the converter to predict the behavior of torque and stator flux of the drive system and to obtain the best suited converter switching state considering the torque and flux errors by evaluating twenty seven (27) possible combinations of the topology. Two cost functions are used to control simultaneously torque and flux. Simulation results show accurate tracking performances of torque and flux, also a unity power factor in the supply side can be achieved by minimizing the reactive power for both motoring and regeneration modes capability of the drive system. The FSMPC method show that multiple objectives can be obtained simultaneously just by adding terms in the global cost function but the main drawback of the method is the choose of the weighting parameters which is an open topic for research.

VII. REFERENCES

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