CURRENT CONTROL OF A VSI-FED INDUCTION MACHINE BY PREDICTIVE TECHNIQUE

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KEYWORDS
Voltage inverter, predictive control, induction motor.

ABSTRACT

The paper deals with a technique which uses the predictive concepts in order to obtain the pulse width modulation strategy of a voltage fed inverter. After the technique is briefly described, it is applied for the case when the inverter supplies an induction motor, the reference values of the currents being obtained from a classical vector control scheme. The described technique is then simulated and the waveforms are compared with ones obtained with preset currents (bang-bang) pulse width modulation, as the behaviour of the two strategies are similar. Finally, the results are cross analysed and further actions are proposed for the work continuation.

INTRODUCTION

Nowadays, the hardware topologies of the inverters designed to supply the induction motors in variable speed drives are well crystallized. Besides quite special architectures, basically, there are the two level inverter (the classical three phase, six switches, bridge) and the multilevel inverters designed for very high power or voltage applications (Moller 2006, Steimel 2010).
Concerning control techniques of the inverters, the strategies are practically unlimited, the literature being quite rich and dynamic (Hartani 2010, Ursaru 2009, Milicevic 2013). Even the classification of these strategies can be performed by considering different criterions. We will take into account here only the source type which the inverter has: voltage or current. We speak all the time about the inverters which are supplied by a DC link having voltage source character, not real current source inverter, for which the DC link has a current source character.
We must see the modulation strategy only as a vector for obtaining the control of the whole driving system, which finally means the control of the developed torque.
Even from the basics of the vector stated by Leonhard, Blaschke and their followers in the 1970s, in rotating references, solidar with the rotor flux, stator flux or magnetizing flux respectively, there is an obvious decoupling between the two components of the stator *current*: while the direct component acts on the flux modulus only and produces the reactive component, the quadrature component generates the torque, being the active component. The two components of the stator current must be thus controlled independently and the flux and torque generation are thus decoupled, similarly to the DC motor.
This means that, as the torque is controlled by the current components, a current source inverter is more suited for the control of the torque developed by the drive. The previous work of the authors emphasized that the field oriented control (FOC) schemes based on current source inverters (preset currents, or bang-bang modulation) are more robust to the parameters variations and have very good dynamics. The main disadvantages of this simple modulation strategy are related to the very high necessary switching frequency (available only in the low range of power), variable switching frequency (difficult to estimate the losses) and interphases dependency. Different techniques were developed for improving the strategy (sinusoidal hysterezis, multilevel hysteresis comparators, Mohseni 2010), but the variable switching frequency rests always as an disadvantage.
The technique we propose has the behaviour of preset currents inverter, but it performs the pulse width modulation with fixed frequency, which is in fact the sampling frequency of the system.
From this point of view (fixed switching frequency given by the sampling one), the proposed technique has a similarity with another very simple method for the toque control, the Direct Torque Control (DTC), suited for electrical traction applications (Takahashi and Noguchi 1986, Baader et al. 1992, Ehsani et al. 1997, Faiz et al. 1999, Haddoun et al. 2007, Ivanov 2009, Ivanov 2010). As will be shown, as the direct controlled variables are the stator currents, the behaviour of the proposed technique is much better.
The proposed technique is feasible due to the increased computational capabilities of the existing DSP which allow the implementation of the predictive control at the

Proceedings 28th European Conference on Modelling and Simulation ©ECMS Flaminio Squazzoni, Fabio Baronio, Claudia Archetti, Marco Castellani (Editors)
ISBN: 978-0-9564944-8-1 / ISBN: 978-0-9564944-9-8 (CD)

level of the converters which induce the hybrid character of the overall control system of the drive. We infer that predictive control has established itself in the last 5-7 years as a very proficient form of controlling highly nonlinear and uncertain systems; moreover the most recent results show its applicability to fast processes among which drives and their converters have a central position (Seo et al. 2009, Prieur and Tarbouriech 2011, Geyer et al. 2008, Mariethoz et al. 2010, Geyer et al. 2009, Trabelsi et al. 2008, Shi et al. 2007, Rodriguez et al. 2007, Larrinaga et al. 2007, Richter et al. 2010, Almer et al. 2010).
The paper will briefly present in the first section the basics of the predictive control. The principle of the predictive control applied to a three phase inverter will be presented in Section 2. Section 3 will detail the predictive control of the inverter fed induction machine drive. Section 4 will present the results of the simulations performed based on a Simulink model. The results will be compared with the ones obtained with the bang-bang strategy and with DTC. Finally, conclusions will be issued and ideas for continuation will be pointed out.

BASICS OF PREDICTIVE CONTROL

The model predictive control is a control technique which has been successfully implemented in industry. The predictive control techniques were used to control both continuous as well as discrete systems (Camacho and Bordons 2004, Bemporad 2007, Lazăr 2006, Maciejowski 2000, Stinga 2012).
The predictive control is derived from optimal control, yet, in this case the optimal control problem involves additional constraints.
The predictive control techniques require solving an open loop optimal control problem, taking into account constraints on input, state and/or output variables. At every moment k, the measured variables and the model of the process are used to compute (to predict) the future behaviour of the system over a prediction horizon N (Fig. 1).

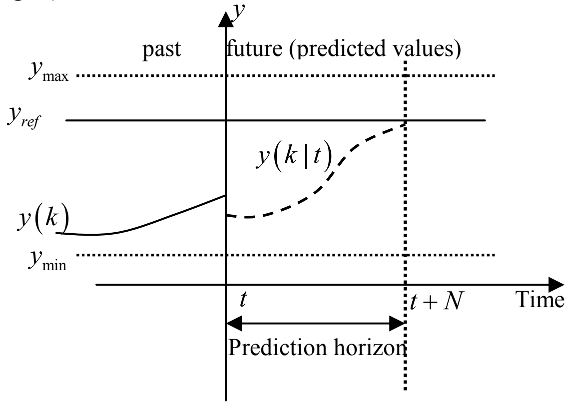

Figure 1: Evolution of System Output Using Predictive Control Strategy (Stinga 2012)

This task is accomplished by determining a set of future control inputs such that the objectives and the system constraints are satisfied. The control input is determined by minimization of a cost function over a time horizon N_c.
Generally, the cost function used in predictive control is defined as follows:

$$J(k)=\sum_{t=1}^{N}\left\|y(k|t)-y_{ref}(k)\right\|_{Q(t)}^{2}+\sum_{t=1}^{N_c}\left\|u(k|t)\right\|_{R(t)}^{2}, \quad (1)$$

subject to constraints specified on the inputs, outputs and input increments (Fig. 2):

$$u_{\min}\le u(k)\le u_{\max},$$
$$y_{\min}\le y(k)\le y_{\max},$$

where:
$Q(t)$ - positive definite error weighting matrix;
$R(t)$ - positive semi-definite control weighting matrix;
$y(k|t)$ - vector of predicted output signals;
$y_{ref}(k)$ - vector of future set points;
$u(k|t)$ - vector of future control inputs;
N - prediction horizon;
N_c - control horizon.

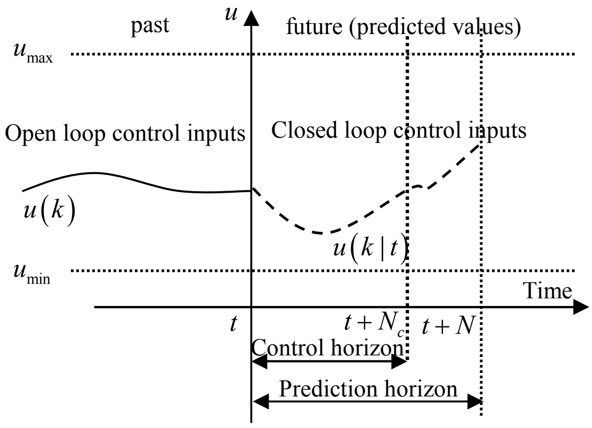

Figure 2: The Control Inputs Applied to the System Using the Predictive Control Strategy (Stinga 2012)

PREDICTIVE CONTROL OF THE THREE PHASE INVERTER

The basic ideas of the predictive control of the three phase bridge inverter are presented in Rodriguez 2012, for a simple R-L load.
The predictive command of the inverter is facilitated by the limited number of possible future states. In fact, the inverter can have only eight different topologies (Fig. 3). These eight different topologies determine seven positions of the voltage phasor (Fig. 4). It is to note that two topologies (7 and 8) are equivalent and determine the same position of the voltage phasor. In practice, one of the two is chosen depending on the actual state of the inverter in order to minimize the number of switches. If the actual state is one of 2, 4 or 6 and the zero phasor must be obtained, the topology 7 will be chosen.

Contrary, if the actual state is one of 1, 3 or 5 and the zero phasor must be obtained, the topology 8 will be chosen.

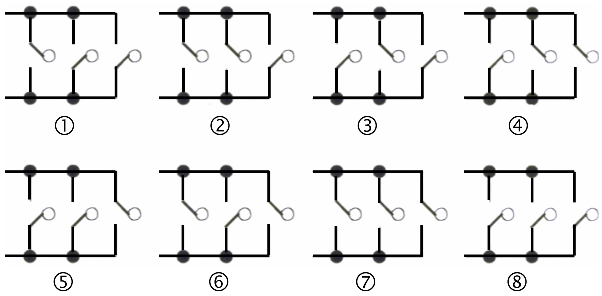

Figure 3: Possible Topologies of the Three Phase Bridge

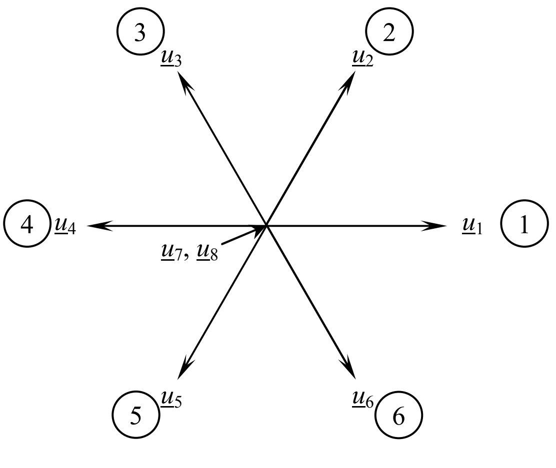

Figure 4: Positions of the Voltage Phasor

The operating principle is to compute at each sampling step the estimations of the (α, β) components of the currents, i_α^e, i_β^e, for all the seven different values of the voltages corresponding to different topologies. Then, the topology that will be applied for the next sampling period will be chosen the one which minimises the cost function

$$J(k) = \left| i_\alpha^*(k+1) - i_\alpha^e(k+1) \right| + \left| i_\beta^*(k+1) - i_\beta^e(k+1) \right|, \quad (2)$$

where i_α^*, i_β^* are the preset values of the (α, β) components of the currents.

PREDICTIVE CONTROL OF THE VSI FED INDUCTION MOTOR DRIVE

The principle described above must be applied considering as load of the inverter, an induction machine.

At each sampling period, having as initial conditions the actual values of the stator and rotor currents components, $i_{s\alpha}$, $i_{s\beta}$, $i_{r\alpha}$, $i_{r\beta}$, the state equation model of the motor (3) is integrated for all the seven different values of the input vector, which consists in the voltage components $[uu] = \left[u_{s\alpha},\ u_{s\beta},\ u_{r\alpha},\ u_{r\beta}\right]^T$,

$$\frac{d}{dt}[i] = [ML]^{-1}\left([uu] - \left([MR] + [MXr]\right)\cdot[i]\right), \quad (3)$$

where:

$[i] = \left[i_{s\alpha},\ i_{s\beta},\ i_{r\alpha},\ i_{r\beta}\right]^T$ - the stator and rotor (α, β) current components;

$[ML] = \begin{bmatrix} L_s & 0 & L_m & 0 \\ 0 & L_s & 0 & L_m \\ L_m & 0 & L_r & 0 \\ 0 & L_m & 0 & L_r \end{bmatrix}$ - the inductances matrix, whose components are:

L_s, L_r – total stator and rotor inductances;

L_m – mutual inductance;

$[MR] = \begin{bmatrix} R_s & 0 & 0 & 0 \\ 0 & R_s & 0 & 0 \\ 0 & 0 & R_r & 0 \\ 0 & 0 & 0 & R_r \end{bmatrix}$ - the resistances matrix;

$[MXr] = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & P\omega_r L_m & 0 & P\omega_r L_r \\ -P\omega_r L_m & 0 & -P\omega_r L_r & 0 \end{bmatrix}$ - the reactance matrix, whose terms depend on the

P – number of pairs of poles and

Ω_r – mechanical speed of the rotor.

The model (3) is completed with the movement equation which must be also be integrated at each sampling period

$$\frac{d\omega_r}{dt} = \frac{1}{J}\left[\frac{3}{2}PL_m\left(i_{s\beta}i_{r\alpha} - i_{s\alpha}i_{r\beta}\right) - T_s\right], \quad (4)$$

where:

J – total inertia at the motor shaft;

T_s – static torque applied to the rotor shaft.

It result seven sets of state variables estimations

$$\left[i_{s\alpha}^e,\ i_{s\beta}^e,\ i_{r\alpha}^e,\ i_{r\beta}^e,\ \omega_r^e\right]_{1\ldots7}$$

and the cost function (2) is computed for the stator currents components

$$J(k)\big|_{1\ldots7} = \left|i_{s\alpha}^* - i_{s\alpha}^e\right| + \left|i_{s\beta}^* - i_{s\beta}^e\right|. \quad (5)$$

The next topology of the inverter is chosen the one which corresponds to the minimum of the seven values given by (5).

At this stage, there are not considered the both topologies which determine the zero voltage phasor (7 and 8), because the aim is only to determine the next stator voltages which minimises the cost function, not the real topology of the inverter. This presents only practical importance at the implementation stage.

The preset values of the stator currents components $i^*_{s\alpha}$, $i^*_{s\beta}$ are the results of a classical FOC of the induction machine supplied by a preset currents inverter, Fig.5.

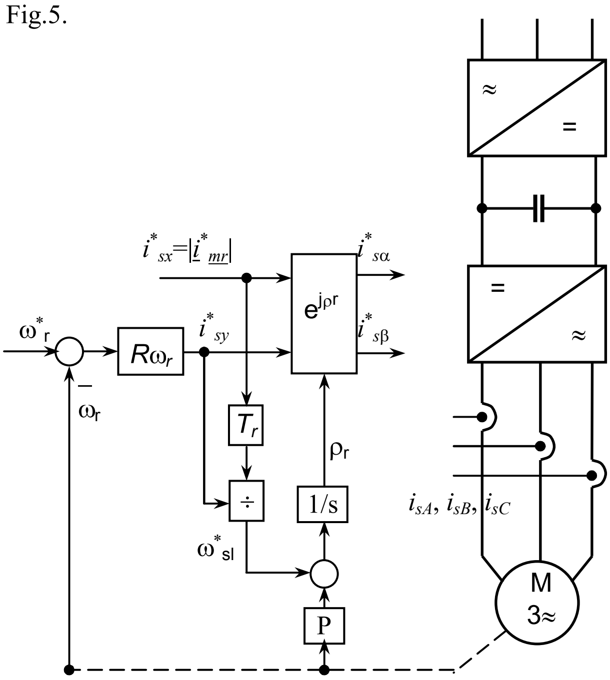

Figure 5: The Classical FOC of the Induction Machine Supplied by Preset Currents Inverter

The block which estimates the currents and computes the cost function (5) must be placed between the rotation transformation block $e^{j\rho r}$ and the inverter.

SIMULINK MODEL OF THE DRIVE

The complete Simulink model of the drive is depicted in Fig. 6.

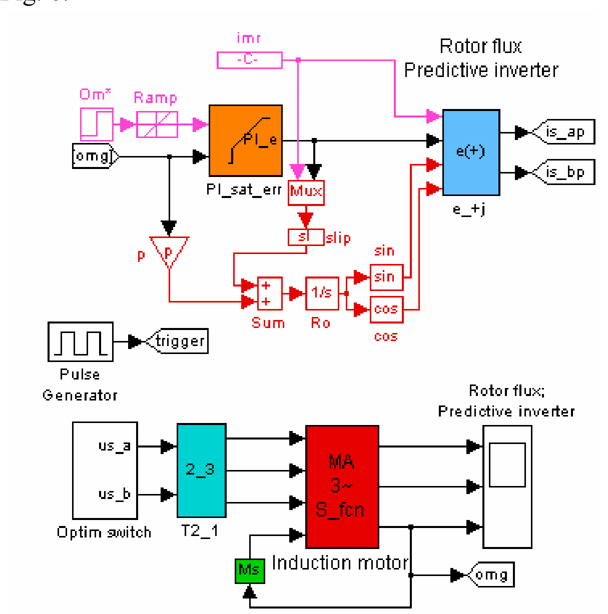

Figure 6: Simulink Model of the Drive

As can be seen, the classical FOC (upper part of the figure) outputs the preset values of the two stator currents components. They a applied (by the way of GoTo tags) to the *Optim switch* block. This block (Fig. 7) computes all the currents in the model (3) and the speed (4), for all the seven possible values of the stator voltages (blocks I_1 to I_7).

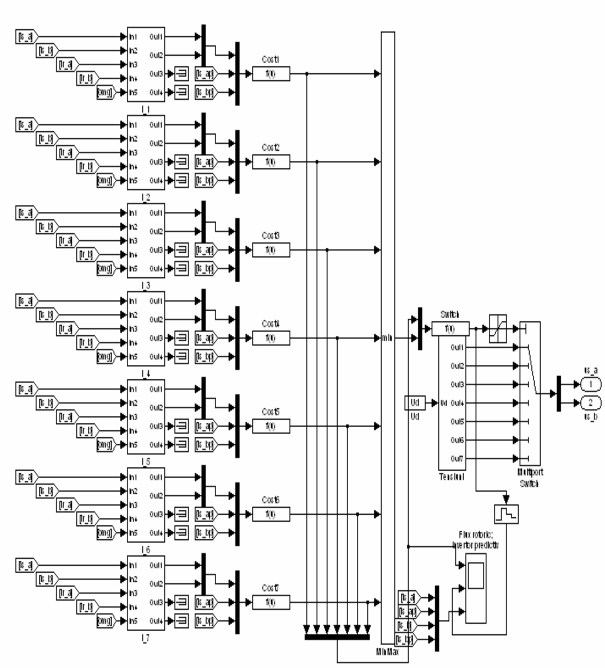

Figure 7: The Optim Switch Block

Each of the seven blocks computes the estimated values of the stator and rotor currents components based on (3), the integrators being reset with the actual values of the four currents components (Fig. 8).

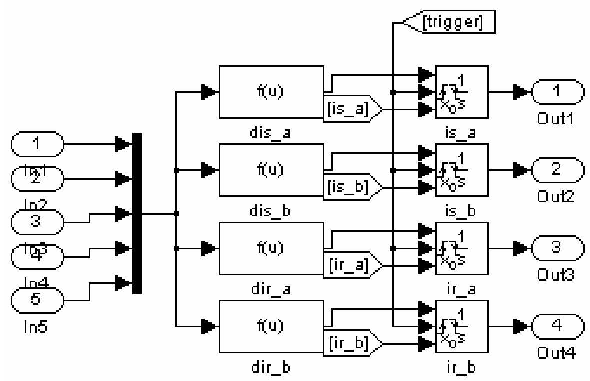

Figure 8: The Computing of the Stator and Rotor Currents Components

The cost function (5) is then computed for all the seven possible values of the stator voltages, the minimum of the seven is determined and it is identified the topology which determines that this minimum is achieved. The output vector consists of the two stator voltage components $u_{s\alpha}$, $u_{s\beta}$, the rotor being considered squirrel cage and consequently, $u_{r\alpha} = u_{r\beta} = 0$.

SIMULATION RESULTS AND COMPARISONS

The simulation results of three types of command are presented in Fig. 9, 10 and 11, in all cases the simulation step being constant and equal to 100 µs. Only the phase currents are plotted as results of the simulations, the comparison being performed from this point of view. Of course, a better (smaller) ripple of the phase currents determines better overall behaviour of the drive (smaller torque ripple, greater average torque and better dynamics).
Fig. 9 plots a detail of the currents obtained by using the presented technique.

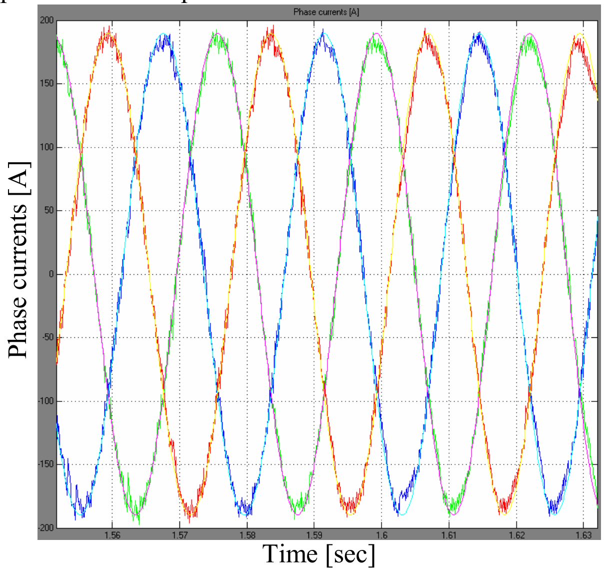

Figure 9: Stator Currents for Predictive Command of the Inverter

These waveforms must be compared with the ones obtained with a classical bang-bang modulator (preset currents), but with fixed switching frequency (the same as for predictive control). In this case (Fig. 10), the ripple of the currents is significantly higher.

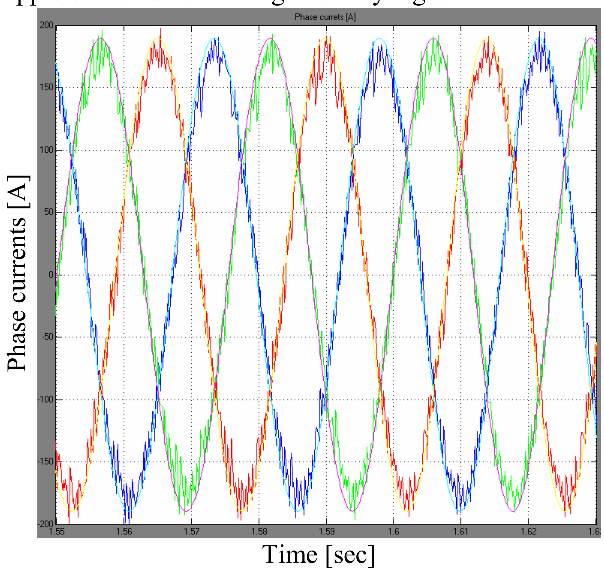

Figure 10: Stator Currents for Preset Currents Modulation

We notice that, for the same constant switching frequency, the current ripple can arise to be four times larger (40 A, compared with 10A). This is because, for preset currents (bang-bang) modulation, the switches are obtained independently on the three phases. For the predictive modulation, the topology of the inverter is chosen globally, as the one which minimizes the currents errors.
Finally, the waveforms are compared with the ones resulted when a classical DTC controls the induction motor. Once again, the sampling period is the same 100 µs. We make this comparison due to the similarity of the commands: the both determine the next stator voltage phasor, but considering different criteria.

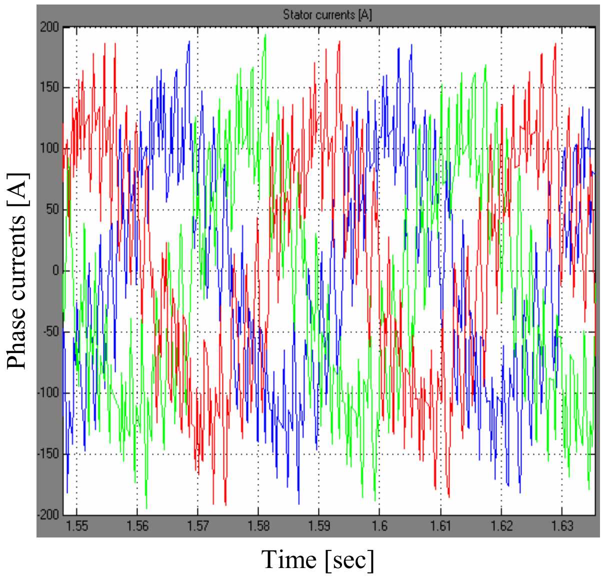

Figure 11: Stator Currents for Classical DTC

We notice that the results are the worst, from the point of view of currents ripple.

All the simulations were performed for a motor with the rated values:

- power: 55 kW;
- speed: 981 r/min;
- L-L voltage: 380 V;
- stator phase resistance: 0.068 Ω;
- rotor phase resistance: 0.044 Ω;
- stator leakage inductance: 0.5 mH;
- rotor leakage inductance: 0.5 mH;
- magnetising inductance: 21.5 mH;
- pairs of poles: 3.

CONCLUSIONS

The paper describes a strategy for pulse width modulation of the inverters which supply induction machines, based on the predictive technique.
A Simulink model of the drive is presented and the results of the simulation are compared, for similar conditions (same fixed step), with other two modulation techniques: preset currents and DTC. The currents ripple

is the smallest when the proposed modulation technique is used. The consequences are favourable in what concern the torque ripple and general dynamic behaviour. The goal of the research is to implement the predictive control for an industrial drive which will be offered on the market.

Further research will be focused on optimizing the proposed technique, in order to reduce the computing time.

ACKNOWLEDGMENTS

Authors whish to thank the UEFISCDI and their partners in the HYDICO project (PN-II-PT-PCCA-2011-3.2-1082), in the frame of which their study has been performed.

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