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REVIEW ARTICLE

A NOVEL DTC-SVM ASSOCIATED WITH THE CALCULATION OF PI REGULATOR OF THE INDUCTION MACHINE

***Lakhdar Djaghdali, Farid Naceri and Sebti Belkacem**

Department of electrical Engineering, University of Batna, Street Chahid Med El Hadi boukhlouf, Batna 05000, Algeria

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ABSTRACT

We present, in this paper, a DTC-SVM control algorithm of an induction machine based on PI controllers. Both the torque and the flux are regulated by a PI controller, where the truth table and the hysteresis are eliminated. We present here a conception method of the PI controllers, associated with the flux and the torque regulation loops and gives analytical formulas for the proportional and integral parameters, depending on the parameters of the induction machine. The effectiveness of the proposed approach is shown by simulation results.

INTRODUCTION

We present, in this paper, a control algorithm SVM-DTC of induction machine based on PI controllers. The torque and flux are regulated by a PI controller, where the truth table and hysteresis are eliminated. Another cruise control is also used to obtain the reference signal of torque. The algorithm retains the basic idea of the technique DTC (Tang *et al.,* 2002; Takahashi and Ohmori, 1989). For this, the technique of orientation of the stator flux is used. Thus, the control voltages can be generated by PI and imposed by SVM technique (Lai and Bowes, 1996; Casadei *et al.,* 2000). In addition, the estimate of the torque and flux is based on the model of the machine voltage (Tang *et al.,* 2002; Habetler *et al.,* 1991). This control structure has the advantages of vector control and direct torque control and helps overcome the problems of conventional DTC (Lai and Chen, 2001; Habetler *et al.,* 1991). PI, so vector modulation technique is used to obtain a fixed switching frequency (Habelter and Profumo, 1992; Casadei *et al.,* 2003) and less torque pulsations and flux (Tang and Rahman, 2001; Kazmierkowski and Kasprowicz, 1995). In this work we present a method of analytical design of the PI controllers associated with the control loops of the flux and torque; thereby we give analytical formulas for the proportional and integral parameters of the regulator, according to the parameters of the machine induction. The effectiveness of the proposed approach is shown by the simulation results. The block diagram of the control structure is shown in (Fig. 1). Two PI controllers are used for controlling the flux and torque, and another for the speed regulation.

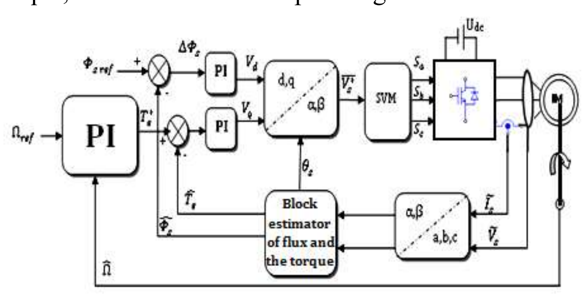


Fig 1. Diagram of DTC-SVM control of induction machine based on PI controllers

Model of the Machine for the Control

Among the various types of models used to represent the induction machine, there is one that uses each of the stator currents, stator flux, speed as state variables and voltages (V_{sd}, V_{sq}) as control variables. This model is presented in reference (d, q), related to the rotating field. This model is expressed by the following system of equations (Carloss, 2000 and Grellet and Clerc, 2000):

****Corresponding author: Lakhdar Djaghdali***
Department of electrical Engineering, University of Batna, Street Chahid Med El Hadi boukhlouf, Batna 05000, Algeria

$$\begin{cases} V_{ds} = R_s.I_{ds} + \frac{d\Phi_{ds}}{dt} - \omega_s.\Phi_{qs} \\ V_{qs} = R_s.I_{qs} + \frac{d\Phi_{qs}}{dt} + \omega_s.\Phi_{ds} \\ V_{dr} = 0 = R_r.I_{dr} + \frac{d\Phi_{dr}}{dt} - (\omega_s - p\Omega).\Phi_{qr} \\ V_{qr} = 0 = R_r.I_{qr} + \frac{d\Phi_{qr}}{dt} + (\omega_s - p\Omega).\Phi_{dr} \end{cases} \quad (1)$$

In addition, the components of the stator and rotor flux are expressed by:

$$\begin{cases} \Phi_{ds} = L_s.I_{ds} + L_m.I_{dr} \\ \Phi_{qs} = L_s.I_{qs} + L_m.I_{qr} \\ \Phi_{dr} = L_r.I_{dr} + L_m.I_{ds} \\ \Phi_{qr} = L_r.I_{qr} + L_m.I_{qs} \end{cases} \quad (2)$$

Moreover, the mechanical equation of the machine is given by:

$$j\frac{d\Omega}{dt} + f\Omega = T_e - T_r \quad (3)$$

The electromagnetic torque equation can be expressed in terms of stator currents and stator flux as follows:

$$T_e = P.(\Phi_{ds}.I_{qs} - \Phi_{qs}.I_{ds}) \quad (4)$$

Where :(I_{ds}, I_{qs}) ;(V_{ds}, V_{qs}) ;(Φ_{ds}, Φ_{qs}); (Φ_{dr}, Φ_{qr}) are currents, voltages, and stator and rotor flux axis d-q.

(R_s, R_r) : stator and rotor Resistance.
(L_s, L_r) : stator and rotor Self Inductance.
(L_m, p): mutual Inductance and Number of pole pairs.
(ω_s , Ω): Stator pulsation and mechanical rotor speed.

Control of The Stator Flux

In the case of the stator flux orientation in the repository (d, q) (Fig. 2), the axis 'd' coincides with the direction of the stator flux vector. $\overline{\Phi_s}$ The 'd' axis component of the stator current I_{sd} is then directly proportional to the magnitude of the stator flux. By controlling and maintaining constant the amplitude of the component of the stator current I_{sd}, we obtain the decoupling of the torque control and the flux of the machine.

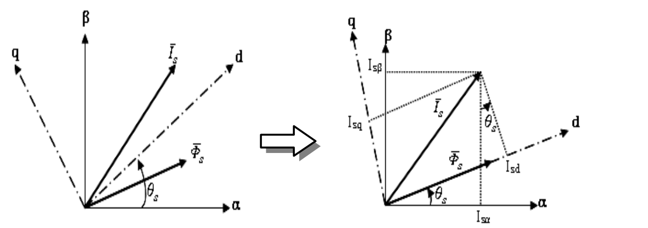


Fig 2. Vectorial representation of the strategy orientation of stator flux

From the machine model developed previously, we can deduce an expression of stator flux vector [10]. So if the stator flux is oriented on the axis we find:

$$\Phi_{sd} = \Phi_s \text{ and } \Phi_{sq} = 0 \quad (5)$$

Then (1) guess:

$$\begin{cases} V_{ds} = R_s.I_{ds} + \frac{d\Phi_s}{dt} \\ V_{qs} = R_s.I_{qs} + \omega_s.\Phi_s \\ V_{dr} = 0 = R_r.I_{dr} + \frac{d\Phi_{dr}}{dt} - (\omega_s - p\Omega).\Phi_{qr} \\ V_{qr} = 0 = R_r.I_{qr} + \frac{d\Phi_{qr}}{dt} + (\omega_s - p\Omega).\Phi_{dr} \end{cases} \quad (6)$$

The torque becomes:

$$T_e = P.\Phi_s.I_{qs} \quad (7)$$

Autopilot with the law:

$$\omega_s = \omega_r + p\Omega \quad (8)$$

Currents and rotor flux can be expressed as a function of the stator currents by:

$$\begin{cases} I_{dr} = \frac{1}{L_m}(\Phi_s - L_s I_{ds}) \\ I_{qr} = -\frac{L_s}{L_m} I_{qs} \end{cases} \quad (9)$$

$$\begin{cases} \Phi_{dr} = \frac{L_r}{L_m}(\Phi_s - \sigma L_s I_{ds}) \\ \Phi_{qr} = -\frac{\sigma L_s L_r}{L_m} I_{qs} \end{cases} \quad (10)$$

Substituting (9), (10) in (6), and taking, into account the Laplace transform, we have:

$$\Phi_s = \left((1 + \sigma t_r s) I_{ds} + \sigma t_r I_{qs} w_r\right) \frac{L_s}{1 + \sigma t_r s} \quad (11)$$

$$I_{qs} = \left(\frac{1}{L_s}\Phi_s - \sigma I_{ds}\right) \frac{t_r W_r}{1 + \sigma t_r s} \quad (12)$$

With: $t_r = \frac{L_r}{R_r}$, $t_s = \frac{L_s}{R_s}$: the constants time respectively rotor, stator, σ : its leakage coefficient.

Expressing the 'd' component of the stator current as a function of the 'q' component of the stator flux and the stator voltages are expressed as follows:

$$\begin{cases} V_{ds} = \frac{\Phi_s}{G_{\Phi_s}} + E_d \\ V_{qs} \approx \omega_s.\Phi_s \end{cases} \quad (13)$$

With:

$$G_{\Phi_s} = \frac{t_s(1 + \sigma t_r s)}{1 + (t_r + t_s)s + \sigma t_r t_s s^2} \quad (14)$$

$$E_d = -\frac{\sigma R_s t_r}{1 + \sigma t_r s} I_{qs} w_r \quad (15)$$

Thus the stator flux can be controlled by the d-component of the stator voltage. (Fig. 3) shows the relationship between Φ_s and V_{sd}, an equivalent system with a second-order perturbation E_d. A PI controller can be used to obtain the desired performance and to maintain the stator flux reference value $\Phi_{s\,ref}$. (Bounadja *et al.,* 2010).

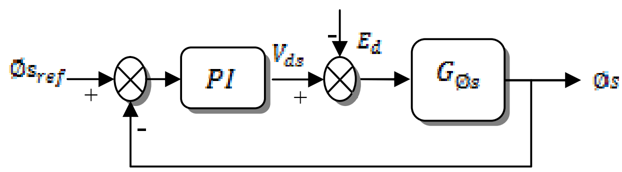


Fig 3. Block diagram of the flux control

Calculating Parameters of the PI Regulator of Flux

The transfer function of the PI is given by:

$$C(s) = K_{pf} + \frac{K_{if}}{s} = K_{pf}\left(1 + \frac{1}{\tau_f s}\right) = K_{pf}\frac{(1+\tau_f s)}{\tau_f s} \quad (16)$$

Where K_{pf} ,K_{if} denote proportional and integral gains of the corrector, and $\tau_f = \frac{K_{pf}}{K_{if}}$: Its constant of time.

The transfer function of the open loop (TFOL) is given by:

$$\text{TFOL} = C(s).G_{\Phi_s} = K_{pf}\frac{(1+\tau_f s)}{\tau_f s}.\frac{t_s(1+\sigma t_r s)}{1+(t_r+t_s)s+\sigma t_r t_s s^2} \quad (17)$$

From the command "roots" in Matlab, one can find the roots of the polynomial:

$$1 + (t_r + t_s)s + \sigma t_r t_s s^2$$

There are two roots:

$P_1 = f\,(t_r, t_s, \sigma)$
$P_2 = f\,(t_r, t_s, \sigma)$

By eliminating the pole dominating, which is the pole nearest to the secondary axis (either the pole P_2). We can write the relation (17) in form 'pole-zero' as follows:

$$\text{TFOL} = K_{pf}\frac{\left(\frac{1}{\tau_f}+s\right)}{s}.\frac{\sigma t_r t_s\left(\frac{1}{\sigma t_r}+s\right)}{(s-P_1)(s-P_2)} \quad (18)$$

To eliminate the dominant pole is placed:

$$(s - P_2) = \left(\frac{1}{\tau_f} + s\right) \quad (19)$$

We have:

$$\text{TFOL} = \frac{K_{pf}}{s}.\frac{t_s(1+\sigma t_r s)}{(s-P_1)} \quad (20)$$

The transfer function in closed loop (TFCL) is written as follows:

$$\text{TFCL} = \frac{\text{TFOL}}{1+\text{TFOL}} \quad (21)$$

Substituting (20) into (21) and after simplification
We have:

$$\text{TFCL} = \frac{K_{pf}t_s(1+\sigma t_r s)}{s^2+(K_{pf}t_s\sigma t_r-P_1)s+K_{pf}t_s} \quad (22)$$

To control the closed loop system, it is necessary to choose the coefficients K_{pf} and K_{if}. For this, we use the method of the imposition of the poles. The transfer function of a standard second order system is characterized by:

$$F(s) = \frac{k\omega_n^2}{s^2+2\xi\omega_n s+\omega_n^2} \quad (23)$$

Where ξ and ω_n are the damping coefficient and the natural angular frequency of the system. By analogy between the expressions (22) and (23) taking into account the expression (19), we find:

$$\begin{cases} (s - P_2) = \left(s + \frac{1}{\tau_f}\right) \\ 2\xi\omega_n = (K_{pf}t_s\sigma t_r - P_1) \end{cases}$$

$$\Rightarrow \begin{cases} \frac{1}{\tau_f} = -P_2 \quad \Rightarrow \quad K_{if} = -K_{pf}P_2 \\ \qquad K_{pf} = \frac{2\xi\omega_n+P_1}{t_s\sigma t_r} \end{cases} \quad (24)$$

Gains are obtained for the corrector that has a minimum response time while ensuring the absence of overshoot. This technique involves the imposition of values damping and the pulsation to determine the coefficients K_{pf} *and* K_{if}.

Control of the Electromagnetic Torque

From relations (11), (12), the current along the axis 'q' can be expressed as:

$$I_{qs} = \frac{t_r(1-\sigma)}{L_s}\frac{\Phi_s W_r}{(1+\sigma t_r s)^2+(\sigma t_r W_r s)^2} \quad (25)$$

Substituting (25) into (7) gives:

$$T_e = p\frac{t_r(1-\sigma)}{L_s}\frac{\Phi_s^2 W_r}{(1+\sigma t_r s)^2+(\sigma t_r W_r)^2} \quad (26)$$

As the modulus of the vector $\overline{\Phi_s}$ remains constant and equal to its reference value Φ_s^*, and $\sigma t_r \ll 1$, the relationship (26) can be simplified as follows (Bounadja *et al.*, 2010):

$$T_e = p\frac{t_r(1-\sigma)}{L_s}\frac{\Phi_s^{*2}}{(1+2\sigma t_r s)}(\omega_s - p\Omega) \quad (27)$$

The electromagnetic torque is proportional to the slip pulse, so equation (27) can be written as follows:

$$T_e = G_{T_e}(\omega_s - p\Omega) \quad (28)$$

Where

$$G_{T_e} = p\frac{t_r(1-\sigma)}{L_s}\frac{\Phi_s^{*2}}{(1+2\sigma t_r s)} \quad (29)$$

As well the torque can be controlled by the stator pulsation. (Fig. 4) shows the relationship between T_e and ω_s. A PI controller is used to achieve the desired performance and maintain the torque reference value
$T_{e\,ref}$.

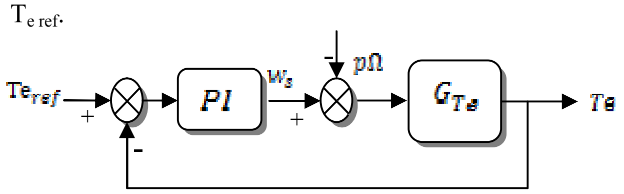


Fig 4. Block diagram of torque control

Calculating Parameters of the PI Regulator of the torque

The transfer function of the PI is given by:

$$C(s) = K_{pc} + \frac{K_{ic}}{s} = K_{pc}\left(1 + \frac{1}{\tau_c s}\right) = K_{pc}\frac{(1+\tau_c s)}{\tau_c s} \quad (30)$$

Where K_{pc} , K_{ic} denote proportional and integral gains of the corrector, and $\tau_c = \frac{K_{pc}}{K_{ic}}$: Its time-constant.

The transfer function in a closed loop (TFCL) is given as follows:

$$\text{TFCL} = \frac{C(s).G_{T_e}}{1+C(s).G_{T_e}} \quad (31)$$

Substituting (29) and (30) into (31) and, after simplification one finds:

$$\text{TFCL} = \frac{K_{ic}Pt_r(1-\sigma)\Phi_s^2(1+\tau_c s)}{2\sigma t_r L_s s^2 + (K_{ic}\tau_c P t_r(1-\sigma)\Phi_s^2 + L_s)s + K_{ic}Pt_r(1-\sigma)\Phi_s^2}$$

$$= \frac{(1+\tau_c s)}{\frac{2\sigma t_r L_s}{K_{ic}Pt_r(1-\sigma)\Phi_s^2}s^2 + \frac{(K_{ic}\tau_c P t_r(1-\sigma)\Phi_s^2 + L_s)}{K_{ic}Pt_r(1-\sigma)\Phi_s^2}s + 1} \quad (32)$$

TFCL the form of a second order system is characterized by:

$$F(s) = \frac{k}{\frac{1}{\omega_n^2}s^2 + \frac{2\xi}{\omega_n^2}s + 1} \quad (33)$$

Identification by the equations (32) and (33) we will have:

$$\begin{cases} \frac{1}{\omega_n^2} = \frac{2\sigma t_r L_s}{K_{ic}Pt_r(1-\sigma)\Phi_s^2} \\ \frac{2\xi}{\omega_n^2} = \frac{(K_{ic}\tau_c P t_r(1-\sigma)\Phi_s^2 + L_s)}{K_{ic}Pt_r(1-\sigma)\Phi_s^2} \end{cases}$$

$$\Rightarrow \begin{cases} K_{ic} = \frac{2\sigma t_r L_s \omega_n^2}{Pt_r(1-\sigma)\Phi_s^2} \\ \tau_c = \frac{K_{pc}}{K_{ic}} = \frac{2\xi}{\omega_n^2} - \frac{L_s}{K_{ic}Pt_r(1-\sigma)\Phi_s^2} \end{cases} \quad (34)$$

The technique is always about the imposition of values and pulsation ω_n and damping ξ to determine the coefficients. K_{pc}, K_{ic} .

Synthesis of PI Speed

The speed control is an essential need in the industry against undesirable variations in the load. For this closed loop control we use a type corrector (PI), which combines the proportional and integral action to improve the steady state and transient response speed. The (fig. 5) shows the block diagram of the speed control.

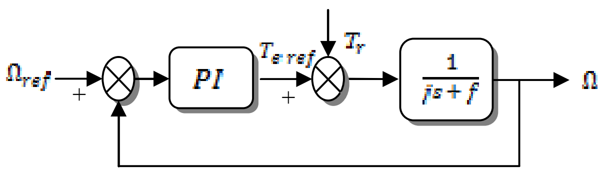


Fig 5. Conventional Control Speed

The equation in the temporal mode of this corrector is given below:

$$u(t) = K_p e(t) + K_i \int_0^t e(\tau)d\tau \quad (35)$$

Where, $e(t)$ $u(t)$ K_p and K_i, denote respectively the error at time t, the command generated and gains of the corrector. The corresponding transfer function is given by

$$PI(s) = K_p + \frac{K_i}{s} = K_p\left(1 + \frac{1}{\tau s}\right) \quad (36)$$

Where s is the Laplace operator derived,
$\tau = \frac{K_p}{K_i}$: Time-constant.

The transfer function in a closed loop is given by:

$$\text{TFCL} = \frac{PI(s).\frac{1}{js+f}}{1+PI(s).\frac{1}{js+f}} \quad (37)$$

Substituting equation (36) in (37), with $T_r = 0$, after simplification is obtained as follows :

$$\text{TFCL} = \frac{(1+\tau s)}{\frac{j}{K_i}s^2 + \left(\frac{f+K_p}{K_i}\right)s + 1} \quad (38)$$

To control the closed loop system, it is necessary to choose the coefficients K_p and K_i, in this case we use the method of the imposition of the poles.
The transfer function of a second order system closed loop is characterized by:

$$F(s) = \frac{K}{1 + \frac{2\xi}{\omega_n}s + \frac{1}{\omega_n^2}s^2} \quad (39)$$

The characteristic equation is: $1 + \frac{2\xi}{\omega_n}s + \frac{1}{\omega_n^2}s^2$ where ξ: the damping coefficient and ω_n: the natural angular frequency of the system. By identifying the relationship (38) we will have the following system:

$$\begin{cases} \frac{1}{\omega_n^2} = \frac{j}{K_i} \Rightarrow K_i = j\omega_n^2 \\ \frac{2\xi}{\omega_n} = \frac{K_p+f}{K_i} \Rightarrow K_p = \frac{2\xi K_i}{\omega_n} - f \end{cases} \quad (40)$$

The Gains of the corrector are obtained to have a minimal response time while ensuring the absence of overshoot. This technique involves the imposition of values of damping and the pulsation ξ and ω_n to determine the K_p and K_i coefficients.

Space Vector Pulse Width Modulation

The voltage vectors, produced by a 3-phase PWM inverter, divide the space vector plane into six Sectors as shown in (Fig. 6) (Habelter and Profumo, 1992).

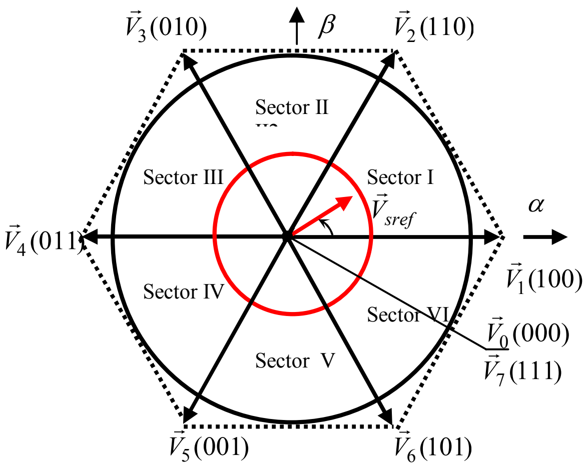


Fig 6. Switching States of Space Vector Modulation

In every sector, the voltage vector is arbitrary synthesized by basic space voltage vector of the two sides of one sector and zero vectors. For example, in the first sector (Fig. 7), $\bar{V}_{s\,ref}$ is a synthesized voltage, space vector and its equation is given by:

$$\bar{V}_{s\,ref} T_s = \bar{V}_0 T_0 + \bar{V}_1 T_1 + \bar{V}_2 T_2 \quad (41)$$

$$T_s = T_0 + T_1 + T_2 \quad (42)$$

Where, T_0 , T_1 and T_2 is the work time of basic space voltage vectors $\bar{V}_0$, $\bar{V}_1$ and $\bar{V}_2$ respectively.

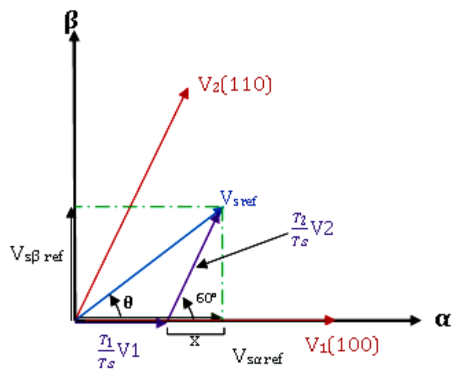


Fig 7. Projection of the Reference Voltage Vector

The determination of the Amount of times T_1 and T_2 given by mere projections is:

$$\begin{cases} V_{s\alpha\,ref} = \frac{T_1}{T_s}|\vec{V}_1| + x \\ V_{s\beta\,ref} = \frac{T_2}{T_s}|\vec{V}_2|\sin(\frac{\pi}{3}) \\ x = \frac{V_{s\beta\,ref}}{tg(\frac{\pi}{3})} \end{cases} \quad (43)$$

$$\Rightarrow \begin{cases} T_1 = \frac{T_s}{2U_{dc}}\left(\sqrt{6}V_{s\alpha\,ref} - \sqrt{2}V_{s\beta\,ref}\right) \\ T_2 = \frac{T_s\sqrt{2}}{U_{dc}} V_{s\beta\,ref} \end{cases} \quad (44)$$

The rest of the period is applying the null-vector. For every sector, switching duration is calculated. The amount of times of the vector implementation can all be related to the following variables (Habelter and Profumo, 1992):

$$\begin{cases} X = \frac{T_s}{U_{dc}}\left(\sqrt{2}V_{s\beta\,ref}\right) \\ Y = \frac{T_s}{2U_{dc}}\left(\sqrt{6}V_{s\alpha\,ref} + \sqrt{2}V_{s\beta\,ref}\right) \\ Z = \frac{T_s}{2U_{dc}}\left(-\sqrt{6}V_{s\alpha\,ref} + \sqrt{2}V_{s\beta\,ref}\right) \end{cases} \quad (45)$$

The implementation of the durations sector boundary vectors is tabulated as follows:

Table 1. Duration of the Sector Boundary Vectors

SECTOR	1	2	3	4	5	6
T_I	-Z	Y	X	Z	-Y	-X
T_{I+1}	X	Z	-Y	-X	-Z	Y

The third step is to compute the duty cycles that have three necessary times:

$$\begin{cases} T_{aon} = \frac{T_s - T_i - T_{i+1}}{2} \\ T_{bon} = T_{aon} + T_i \\ T_{con} = T_{bon} + T_{i+1} \end{cases} \quad (46)$$

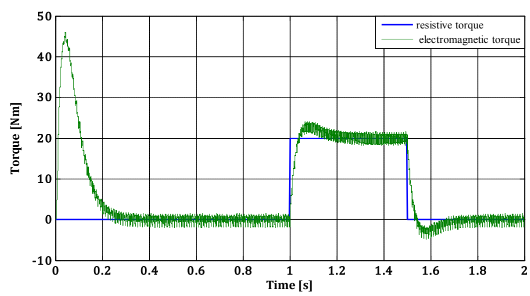
The last step is to assign the right duty cycle (T_{xon}) to the right motor phase according to the sector.

Table 2. Assigned Duty Cycles to the PWM Outputs

Sector	1	2	3	4	5	6
S_a	T_{aon}	T_{bon}	T_{con}	T_{con}	T_{bon}	T_{aon}
S_b	T_{bon}	T_{aon}	T_{aon}	T_{bon}	T_{con}	T_{con}
S_c	T_{con}	T_{con}	T_{bon}	T_{aon}	T_{aon}	T_{bon}

RESULTS OF DISCUSSION

To approve the proposed approach, we simulated the behavior of the drive system represented by the block diagram of (Fig. 1) for Ω_{ref} = 100rd/s. (Fig. 8,9,10,11,12) show the simulation results in the introduction of a load torque nominal of 20 Nm after booting empty at time t = 1s, then canceled it at time t = 1.5s. The modulus of the stator flux is quickly established in its reference value. It was found that the speed reaches its reference Ωref = 100rd / s without overshoot and the disturbance caused by discharges of instructions loads applied to different times above are eliminated.



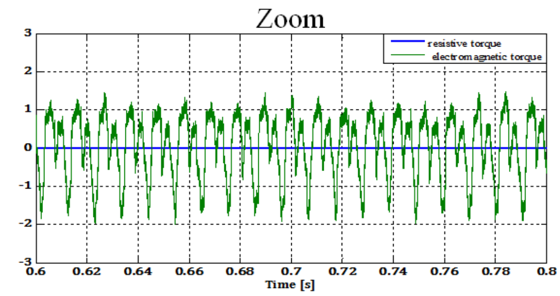


Fig 8. The Electromagnetic Torque

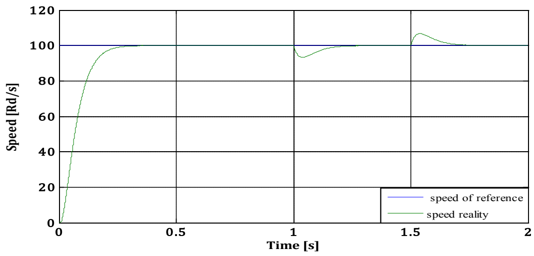


Fig 9. The Mechanical Speed

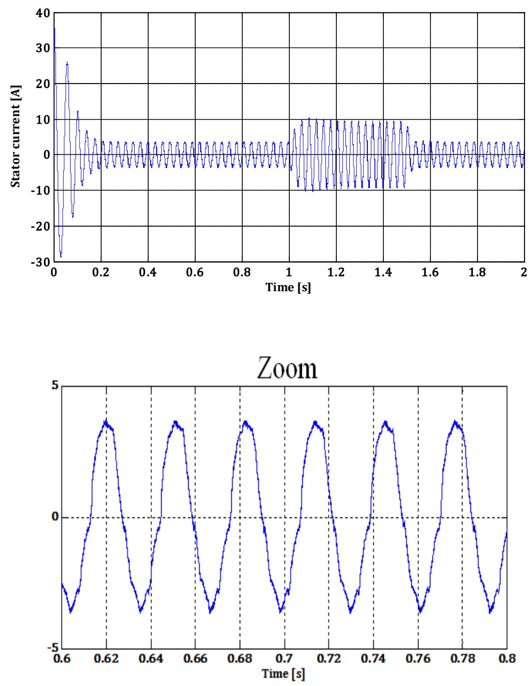
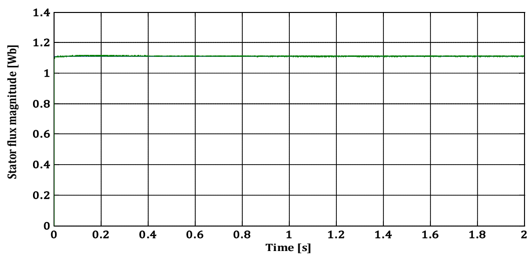


Fig 10. The Stator Current



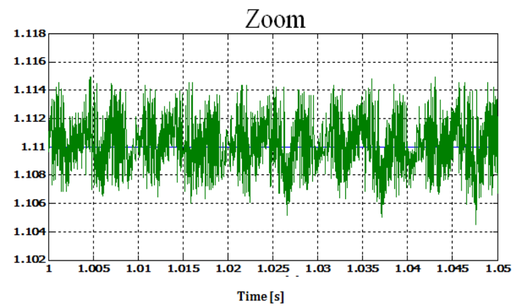


Fig 11. The Magnitude of the Stator Flux

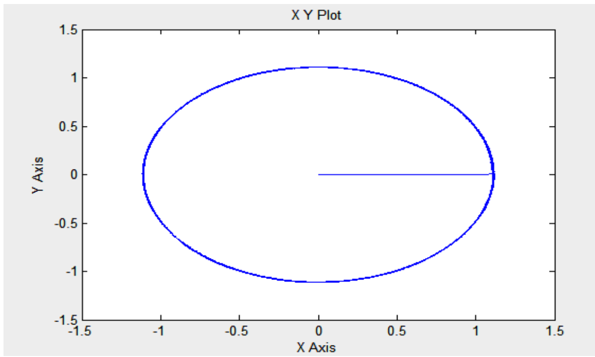


Fig 12. The Stator Flux Trajectory

Indeed, the electromagnetic torque is quickly about to follow the instructions of charges introduced. The dynamic component of the stator flux is not affected by the application of these instructions loads. The simulation results show that the proposed analytical approach to the design of PI controllers of flux and torque is quite rigorous.

Conclusion

In this article we synthesized and provided analytical formulas for the proportional and integral parameters of PI controllers associated with loops of flux and torque. The effectiveness of the approach is approved by simulation tests. The simulation results show that the proposed approach for the design of PI flux and torque is sufficiently rigorous. Also as showed by the simulation that the algorithm presents high performance in terms of minimizing the torque ripple and flux as well as the reduction of the switching frequency of the inverter. Additionally, the application of SVM guarantee:

- Inverter switching frequency is constant;
- Distortion Caused by exchange sector is delimited;
- Low frequency sampling is required;
- High robustness;
- Good dynamic response;
- Low complexity.

Characteristics of the Machine used for Simulation

parameter	symbol	Value
Number of pole pairs	p	2
Power	P_u	3 KW
Line voltage	U_n	380V
Line current	I_n	6.3A
Nominal frequency	f	50Hz
Mechanical rotor speed	N_n	1430 tr/mn
Electromagnetic torque	T_e	20Nm
Stator Resistance	R_s	3.36 Ω
Rotor Resistance	R_r	1.09 Ω
Stator cyclic inductance	L_s	0.256H
Mutual cyclic Inductance	L_m	0.236H
Rotor cyclic inductance	L_r	0.256H
Rotor inertia	j	$4,5.10^{-2}$ Kg.m
Viscosity coefficient	f	$6,32.10^{-4}$ N.m.sec.

REFERENCES

Bounadja, M., Belarbi, A.W. and Belmadani, B. 2010. "*A high performance svm-dtc scheme for induction machines have integrated starter generator in hybrid electric vehicles*". Journal "*Nature and Technology*". No. 02 . pages 41 to 47.

Carloss of wit, C. 2000. "*Modeling and Vector Control DTC* ". Science Edition Hermes Europe.

Casadei, D., Serra, G., Tani, A., Zarri, L. and Profumo, F. 2003. "*Performance Analysis of a Speed-Sensorless Induction Motor Drive Based on a Constant-Switching-Frequency DTC Scheme*". IEEE Trans. on Industry Applications, 39(2):476–484.

Casadei, D., Grandi, G. and Serra, G. 1993. "*Rotor Flux Oriented Torque-Control of Induction Machines Based on Stator Flux Vector Control*". in Proc. EPE'93, Brighton, UK, 5:67–72,.

Casadei, D., Serra, G. and Tani, A. 2000. "*Implementation of a Direct Torque Control Algorithm for Induction Motors Based on Discrete Space Vector Modulation*". IEEE Trans. on Power Electronics, 15(4):769–777.

Grellet, G. and Clerc, G. 2000. "*Electric Actuators, Principle, Models, Order* ". Electro Collection. Eyrolles Edition.

Habelter, T. and Profumo, F. 1992. "*Direct Torque Control of Induction Machines using Space Vector Modulation*". IEEE Transactions on Industrial Applications, pp-1045-1053.

Habetler, T. G., Profumo, F., Pastorelli, M. and Tolbert, L. M. 1991. "*Direct torque control of induction machines using space vector modulation*". in Conference Record IEEE IAS Annual Meeting, vol. 1, 1991, pp. 428–435.

Kazmierkowski, M. P. and Kasprowicz, A. 1995. "*Improved direct torque and flux vector control of PWM inverter-fed induction motor drives*". IEEE Transactions on Industrial Electronics, Vol. 42, No.4, pp. 344–350.

Lai, Y. S. and Bowes, S. R. 1996. "*A universal space vector modulation based on regular-sampled pulse-width modulation*". in Proc. of the IEEE IECON, 1996, pp. 120–126.

Lai, Y.S. and Chen, J.H. 2001. "*A New Approach to Direct Torque Control of Induction Motor Drives for Constant Inverter Switching Frequency and Torque Ripple Reduction*". IEEE Trans. on Energy Conversion, 16(3):220–227.

Takahashi, I. and Ohmori. Y. 1989. "*High-Performance Direct Torque Control of an Motor*". IEEE Trans. on Industrial Applications, 25(2):257–264.

Tang, L. and Rahman, M.F. 2001. "*A New Direct Torque Control Strategy for Flux and Torque Ripple Reduction for Induction Motors Drive by Using Space Vector Modulation*". IEEE PESC 32nd International Conference, pages 1440–1445.

Tang, L., Zhong, L., Rahman, A. F. and Hu, Y. 2002. "*An investigation of a modified direct torque control strategy for flux and torque ripple reduction for induction machine drive system with fixed switching frequency*", IEEE-IAS Industry Applications Conf

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