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Theoretical comparison between Field Oriented and Generalized Predictive Control for an Induction Motor

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Abstract:

In this work, we present an induction motor flux and speed control strategy. Initially, we used the field oriented control (FOC) principle to generate two structures of this machine control. The first structure is realized around two proportional integral regulators (PI), the second consists of the introduction of two Generalized predictive controllers (GPC). The proposed structure is based on the synthesis of the flux and speed regulators and on the conception of the rotor flux sliding mode observer. In order to improve the robustness of the two control strategies face to the parameters variations and the external perturbations, we have developed a rotor resistance adaptation algorithm based on the Lyapunov theory. A comparative study between the two suggested structures shows good performances of the generalized predictive control.

Keywords:

Induction motor, PI controller, generalized predictive control, Sliding mode observer, parameter adapter, Robustness.

List of symbols

$V_{s\alpha\beta}$: stator voltages in reference frame $\alpha\beta$,
$i_{s\alpha\beta}$: stator currents in reference frame $\alpha\beta$,
$\varphi_{r\alpha\beta}$: rotor fluxes in reference frame $\alpha\beta$,
R_s, R_r : stator and rotor resistances,
ℓ_s, ℓ_r : stator and rotor cyclic inductances,
M : stator-rotor cyclic mutual inductance,
T_{re} : the estimate of rotor time constant.
σ : Blondel coefficient,
ω_s : angular speed of the rotating field referred to the stator,
ω_r : angular speed of the rotating field referred to the rotor,
p : Laplace operator ,
J : moment of inertia,

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f : damping coefficient,
T_{em}, T_L : electromagnetic and load torque,
np : pairs pole number.

1. Introduction

The development of the control techniques at a variable speed makes necessary the modelling of the induction motor in a precise way to be able to simulate its behaviour in a dynamic mode [3]. The induction motor saw its application field of strongly increased thanks to the implementation of the FOC which makes it possible to decouple the basic variables from its operation to knowing the electromagnetic torque and rotor flux. Thus, the requested performances from the variators becoming increasingly raised, the sophistication of the controllers made only grow.

Sight that the regulation of the electrical and mechanical parameters machine by PI regulators has sensitivity with respect to the parameter variations; the generalized predictive control constitutes one of the new most interesting solutions to implant during this last decade [8,9,10]. In this work, we present two approaches of direct rotor field oriented control (DRFOC) by rotor flux and speed regulators. The first control is around two PI regulators whereas the second control uses two predictive controllers.

Knowing that rotor flux is a not measurable, its estimate or its observation appears of capital importance. Although the DRFOC gives good results for the speed control, it is practically penalized for its parameter variations sensitivity [1,3]. For that, we propose in this article to introduce a rotor flux observer based on the sliding mode principle, and including a rotor resistance adaptation algorithm.

This article is organized in six sections. The second section relates to the modelling of an induction motor. The two structures of the suggested control are detailed in the third section. Then, we develop a sliding mode flux observer in the fourth section. In the fifth section we set up a rotor resistance adaptation algorithm based on the theory of *Lyapunov* stability whereas the simulations results are illustrated in the last section. We close the paper by a conclusion.

2. Mathematical model of the induction motor

The induction motor can be described by five non-linear differential equations with four electrical variables, one mechanical variable and two control variables such as [2]:

$$\begin{cases} \dfrac{d}{dt} i_{s\alpha} = -\gamma\, i_{s\alpha} + \dfrac{k}{Tr}\varphi_{r\alpha} + n_p k \varphi_{r\beta}\omega_r + \dfrac{1}{\sigma \ell_s} V_{s\alpha} \\ \dfrac{d}{dt} i_{s\beta} = -\gamma\, i_{s\beta} + \dfrac{k}{Tr}\varphi_{r\beta} - n_p k \varphi_{r\alpha}\omega_r + \dfrac{1}{\sigma \ell_s} V_{s\beta} \\ \dfrac{d}{dt}\varphi_{r\alpha} = \dfrac{M}{Tr}\ i_{s\alpha} - \dfrac{1}{Tr}\varphi_{r\alpha} - n_p \omega_r \varphi_{r\beta} \\ \dfrac{d}{dt}\varphi_{r\beta} = \dfrac{M}{Tr}\ i_{s\beta} + n_p \omega_r \varphi_{r\alpha} - \dfrac{1}{Tr}\varphi_{r\beta} \end{cases} \tag{1}$$

with

$$k = \frac{M}{\sigma\, \ell_s \ell_r} \quad \text{and} \quad \gamma = \frac{M^2 R_r}{\sigma\, \ell_s \ell_r^{\,2}} + \frac{R_s}{\sigma\, \ell_s}$$

The mechanical equation is given by:

$$J\frac{d\,\Omega_m}{dt} + f\,\Omega_m = T_{em} - T_L \tag{2}$$

The expression of the electromagnetic torque is written as follows:

$$T_{em} = \frac{3 n_p M}{2\ell_r}(\varphi_{rd}\, i_{sq} - \varphi_{rq}\, i_{sd}) \tag{3}$$

3. Induction motor control

3.1. FOC Principle

The principle of the FOC is based on the adaptation of the d-axis Park's frame according to the rotor flux level [2,3], that allows to write:

$$\begin{cases} \varphi_{rd} = \varphi_r \\ \varphi_{rq} = 0 \end{cases} \tag{4}$$

In these conditions, one can write:

$$\begin{cases} 0 = R_r i_{rd} + \dfrac{d}{dt}\varphi_{rd} \\ 0 = R_r i_{rq} + \omega_r \varphi_{rd} \end{cases} \tag{5}$$

while the expression of the electromagnetic torque is given by:

$$T_{em} = \frac{3}{2} n_p \frac{M}{\ell_r} \varphi_{rd} i_{sq} \tag{6}$$

After some easy development, relations 5 and 6 show that the dynamics of the flux and the torque is directly connected to the control of d-axis and q-axis stator currents.

The concept of the proposed FOC structure is based on the determination of two transfer functions connected, respectively, the rotor flux in the d-axis stator voltage, and the mechanical speed in the q-axis stator voltage. To determine these transfer functions, we consider the following system:

$$\begin{cases} V_{sd} = R_s i_{sd} + \dfrac{d\varphi_{sd}}{dt} - \omega_s \varphi_{sq} \\ V_{sq} = R_s i_{sq} + \dfrac{d\varphi_{sq}}{dt} + \omega_s \varphi_{sd} \end{cases} \tag{7}$$

where d-axis and q-axis flux components are expressed by:

$$\begin{cases} \varphi_{sd} = \sigma \ell_S \, i_{sd} + \dfrac{M}{\ell_r} \varphi_r \\ \varphi_{sq} = \sigma \ell_S \, i_{sq} \end{cases} \tag{8}$$

During this development we used the speed and rotor flux expressed as follows:

$$\omega_s = \omega_r + \frac{M}{T_r} \frac{i_{sq}}{\varphi_r} \tag{9}$$

$$\frac{d}{dt}\varphi_r = -\frac{1}{T_r}\left(\varphi_r + M i_{sd}\right) \tag{10}$$

The load torque is considered proportional to the speed

$$T_L = K_c \omega_r \tag{11}$$

After some development using the Laplace transformer and relations (6) to (11), we rewrite the stator voltage system given by equation (12) :

$$\begin{cases} V_{sd} = \left(\left(R_s + p\sigma \ell_S + \dfrac{M^2}{\ell_r T_r} \right) \left(\dfrac{pT_r + 1}{M} \right) - \dfrac{M}{\ell_r T_r} \right) \varphi_r - \omega_s \sigma \ell_S i_{sq} \\ V_{sq} = \left(R_s + p\sigma \ell_S + \dfrac{M^2}{\ell_r T_r} \right) \left(\dfrac{Jp + K_c}{K\varphi_{ref}} \right) \omega_r + \dfrac{M}{\ell_r} \omega_r \varphi_r + \omega_s \sigma \ell_S i_{sd} \end{cases} \tag{12}$$

Finally, after some simplification, the two transfer functions can be expressed as follows:

$$\begin{cases} F_{\varphi_r}(p) = \dfrac{\varphi_r}{V_{sd1}} = \dfrac{M/R_s}{1+(T_s+T_r)p+\sigma T_s T_r p^2} \\ F_{\omega_r}(p) = \dfrac{\omega_r}{V_{sq1}} = \dfrac{K\varphi_{ref}}{K_c K_1 + (JK_1 + K_c \sigma \ell_s)p + J\sigma \ell_s p^2} \end{cases} \quad (13)$$

with

$$K = \frac{3}{2} n_p \frac{M}{\ell_r},\ K_1 = R_s + \frac{M^2}{\ell_r T_r},\ \begin{cases} V_{sd} = V_{sd1} + V_{sd2} \\ V_{sq} = V_{sq1} + V_{sq2} \end{cases} \text{ and } \begin{cases} V_{sd2} = -\omega_s \sigma \ell_s i_{sq} \\ V_{sq2} = \dfrac{M}{\ell_r} \omega_r \varphi_r + \omega_s \sigma \ell_s i_{sd} \end{cases}$$

The voltages V_{sd2} and V_{sq2} represent the voltage compensation terms.

The discrete linearized transfer function for the rotor flux magnitude including a zero-order hold is then given by [8]

$$F_{\varphi_r}(q^{-1}) = \frac{B_\varphi(q^{-1})}{A_\varphi(q^{-1})} = Z[B_0(p) F_{\varphi_r}(p)] \quad (14)$$

It results in the following transfer function

$$A_\varphi(q^{-1}) = 1 + a_{\varphi 1} q^{-1} + a_{\varphi 2} q^{-2}$$
$$B_\varphi(q^{-1}) = b_{\varphi 0} + b_{\varphi 1} q^{-1}$$

The same approach is considered to obtain the discrete linearized transfer function of the rotor speed

$$F_{\omega_r}(q^{-1}) = \frac{B_\omega(q^{-1})}{A_\omega(q^{-1})} = Z[B_0(p) F_{\omega_r}(p)] \quad (15)$$

where

$$A_\omega(q^{-1}) = 1 + a_{\omega 1} q^{-1} + a_{\omega 2} q^{-2}$$
$$B_\omega(q^{-1}) = b_{\omega 0} + b_{\omega 1} q^{-1}$$

Where Z[.] is the Z transform, $B_0(p) = \dfrac{\left(1 - e^{-Te\,p}\right)}{p}$ is the zero-order hold transfer function, and Te is the sampling period.

The coefficients of A_φ, B_φ, A_ω and B_ω polynomials are given in the appendix.

3.2. Proportional Integral Regulator

The method that we propose consists in introducing two PI regulators. The first regulator concerns the speed and the second is for the rotor flux regulation:

$$u(k) = u(k-1) + k_p\left[y_c(k) - y(k)\right] + k_i\left[y_c(k-1) - y(k-1)\right] \tag{16}$$

The determination of the regulators parameters depends on the considered loop and the induction motor load type. In our case we calculated these parameters by poles compensation technique [2]. The proportional action is used on the measurement and the integral action is used on the tracking error.

3. 3. Predictive controller

The prediction digital model is defined by an input/output transfer function and it is represented as a controlled auto regressive integrated moving average form (CARIMA) [8,9,10].

$$A(q^{-1})y(k) = B(q^{-1})u(k-1) + \frac{\xi(k)}{D(q^{-1})} \tag{17}$$

where $u(k)$ and $y(k)$ are the process input and output. *A* and *B* are two polynomials in the backward shift operator q^{-1}.

$$A(q^{-1}) = 1 + a_1 q^{-1} + \cdots + a_{n_a} q^{-n_a}$$

$$B(q^{-1}) = b_0 + b_1 q^{-1} + \cdots + b_{n_b} q^{-n_b}$$

$\xi(k)$ is an uncorrelated random sequence with zero mean and unit variance. The operator $D(q^{-1}) = 1 - q^{-1}$ ensures an integral control law.

By assuming that the system output is affected by noise, the two transfer functions given in equations (14) and (15) , can be easier written as a CARIMA model.

The control law is obtained by minimization of a quadratic cost function on the future tracking errors with a weighting control term. The quadratic criterion is expressed by:

$$J = \sum_{j=N_1}^{N_2} \left(\hat{y}(k+j) - y_c(k+j)\right)^2 + \lambda \sum_{j=0}^{N_u-1} \left(Du(k+j)\right)^2 \tag{18}$$

with,

$\hat{y}(k+j)$, $y_c(k+j)$ and $Du(k+j)$ represent respectively, the predicted output, the reference output and the control increment.

N_1, N_2, N_u and λ represent respectively, the initialization horizon, the prediction horizon, the control horizon and the control weighting coefficient.

The optimal sequence of future control is given by:

$$\tilde{u} = M[y_C - \mathrm{if}(q^{-1})y(t) - \mathrm{ih}(q^{-1})Du(t-1)] \tag{19}$$

with,

$$M = \left[G^T G + \lambda I_{N_u}\right]^{-1} G' = \begin{bmatrix} m^T{}_1 \\ . \\ . \\ m^T{}_{N_u} \end{bmatrix} \quad \text{and} \quad \begin{array}{l} \mathrm{if}(q^{-1}) = [F_{N1}(q^{-1}) \;\; \ldots \;\; F_{N2}(q^{-1})]' \\ \mathrm{ih}(q^{-1}) = [H_{N1}(q^{-1}) \;\; \ldots \;\; H_{N2}(q^{-1})]' \\ y_C = [y_C(t+N_1) \;\; \ldots \;\; y_C(t+N_2)]' \\ \tilde{u} = [Du(t) \;\; \ldots \;\; Du(t+N_u-1)]' \end{array}$$

In predictive control, only the first value of the sequence of future control is actually applied to the system.

The polynomial representation of the equivalent regulator in RST form is given by the following equation:

$$S(q^{-1})Du(t) = -R(q^{-1})y(t) + T(q^{-1})y_C(t) \tag{20}$$

Whereas, the three polynomials are given by:

$$S(q^{-1}) = 1 + m_1{}^T \mathrm{ih}(q^{-1})q^{-1} \tag{21}$$

$$R(q^{-1}) = m_1{}^T \mathrm{if}(q^{-1}) \tag{22}$$

$$T(q) = m_1{}^T [q^{N1} \;\; \ldots \;\; q^{N2}]^T \tag{23}$$

With the polynomial controller, the stability of the controlled loop may be examined for specific parameter values. The selection process for the RST parameters is accomplished by generating a large parameter set (N_1, N_2, N_u, λ) assuring closed-loop stability [11].

- The $T_e N_1$ product is chosen equal to the system dead time. Generally N_1 can be selected equal to one for a traditional configuration without particular dead time.

- The $T_e N_2$ product is limited by the response time value. When N_2 is large, the corrected system is stable and slow. The improvements brought on stability are minimal compared to the increase in the response time. .

- For an open loop stable process, N_u is chosen equal to one due to its fair influence on the stability margins and for simplicity of calculations.

- The control weighting coefficient λ is attached to the system static gain. The optimal value of λ assuring a better stability margins, can be given by :

$$\lambda = Tr(G^T G) \tag{24}$$

where $Tr(.)$ is the trace of the matrix.

Table 1. Choice of the GPC parameters

	N_1	N_2	N_u	λ
Flux	1	12	1	4
Speed	1	12	1	0, 05

4. Adaptive Sliding mode observer

The major advantage of a sliding mode observer is its insensitivity to parameter variations and external load disturbances once on the switching surface. The sliding mode observer design procedure consists of performing the following two steps:

- Design the manifold (the intersection of the sliding mode surface) S.
- The observer gain is determined to drive the estimation error trajectories to S and maintain it on the set [2,5,6,7].

The proposed adaptive observer has the following form:

$$\begin{cases} \frac{d}{dt}\hat{i}_{s\alpha} = -\gamma_e \hat{i}_{s\alpha} + \frac{k}{T_{re}}\hat{\varphi}_{r\alpha} + n_p k \hat{\varphi}_{r\beta}\,\omega_r + \frac{1}{\sigma \ell_s} V_{s\alpha} + A_{11} Is_1 + A_{12} I_{s2} \\ \frac{d}{dt}\hat{i}_{s\beta} = -\gamma_e \hat{i}_{s\beta} + \frac{k}{T_{re}}\hat{\varphi}_{r\beta} - n_p k \hat{\varphi}_{r\alpha}\,\omega_r + \frac{1}{\sigma \ell_s} V_{s\beta} + A_{21} Is_1 + A_{22} I_{s2} \\ \frac{d}{dt}\hat{\varphi}_{r\alpha} = \frac{M}{T_{re}}\hat{i}_{s\alpha} - \frac{1}{T_{re}}\hat{\varphi}_{r\alpha} - n_p \omega_r \hat{\varphi}_{r\beta} + A_{31} Is_1 + A_{32} I_{s2} \\ \frac{d}{dt}\hat{\varphi}_{r\beta} = \frac{M}{T_{re}}\hat{i}_{s\beta} + n_p \omega_r \hat{\varphi}_{r\alpha} - \frac{1}{T_{re}}\hat{\varphi}_{r\beta} + A_{41} Is_1 + A_{42} I_{s2} \end{cases} \tag{25}$$

The variable $\hat{x}$ is the estimates of parameter x .
with,

$$T_{re} = \frac{\ell_r}{R_{re}} \;,\; \gamma_e = \frac{M^2 R_{re}}{\sigma \ell_s \ell_r^2} + \frac{R_s}{\sigma \ell_s} \;,\; Is = \begin{pmatrix} I_{s1} \\ I_{s2} \end{pmatrix}$$

T_{re} is the estimate of rotor time constant and $A_{ij}\ (i = 1,2,3,4\ et\ j = 1,2)$ represent the observer gains, I_s is the vector signs of the chosen sliding surface.
The errors of stator currents are directly connected to the sliding surface by:

$$S = \begin{pmatrix} S1 \\ S2 \end{pmatrix} = D^{-1} \begin{pmatrix} i_{s\alpha} - \hat{i}_{s\alpha} \\ i_{s\beta} - \hat{i}_{s\beta} \end{pmatrix} \quad (26)$$

with

$$D = \begin{pmatrix} \frac{k}{Tr} & n_p k \omega_r \\ -n_p k \omega_r & \frac{k}{Tr} \end{pmatrix}^{-1}$$

The matrix *D* depends on induction motor electrical and mechanical parameters and it allows fixing the observer dynamics. If we consider the numerical treatment of the stator currents is perfect, one can consider that the observation of these sizes gives a zero error. Under these conditions, the currents matrix gains is given by:

$$A_I = \begin{pmatrix} A11 & A12 \\ A21 & A22 \end{pmatrix} = D \begin{pmatrix} \delta 1 & 0 \\ 0 & \delta 2 \end{pmatrix} \quad (27)$$

with

δ_1 and δ_2 are positive constants satisfied the *Lyapunov* approach stability condition. [5,6].

$$A\Phi = \begin{pmatrix} A31 & A32 \\ A41 & A42 \end{pmatrix} = \begin{pmatrix} (q_1 - \frac{1}{Tr})\delta 1 & n_p \omega_r\ \delta 2 \\ n_p \omega_r\ \delta 1 & (q_2 - \frac{1}{Tr})\delta 2 \end{pmatrix} \quad (28)$$

with, q_1, q_2, δ_1 and δ_2 are positive constants.

The sliding mode observer gains are chosen in such away that *Lyapunov* stability conditions are satisfied [5] .This yields:

$$\begin{cases} \delta_1 \geq \left| \left(\left(\frac{kM}{\ell_r}(Rr - Rre)\left(\frac{\varphi_{r\alpha}}{M} - i_{s\alpha}\right) + k\omega_r e_4 + \frac{Rre}{\ell_r} ke_3 \right) \right| \\ \delta_2 \geq \left| \left(\left(\frac{kM}{\ell_r}(Rr - Rre)\left(\frac{\varphi_{r\beta}}{M} - i_{s\beta}\right) + k\omega_r e_3 + \frac{Rre}{\ell_r} ke_4 \right) \right| \end{cases} \tag{29}$$

where e_3, e_4 are flux errors.

It is well known that sliding mode techniques generate undesirable chattering. This problem can be discarded by replacing the switching function by a smooth continuous function.

To this end, we propose to replace the Sign function by the following function:

$$Sat(S_i) = \begin{cases} 1 & if \quad S_i > \gamma \\ -1 & if \quad S_i < -\gamma \\ \frac{S_i}{\gamma} & if \quad |S_i| < \gamma \end{cases} \tag{30}$$

where γ represents the thickness of the boundary layer .

The estimate of the rotor resistance is given as follows:

$$\frac{dR_{re}}{dt} = q_3 (A_{1,2} I_s)^T K \tag{31}$$

with, $q_3 > 0$ and $K = \frac{\varphi_{re}}{\ell r} - \frac{M}{\ell r} i_s$

It should be noted that the sliding mode observer is asymptotically stable while being based on the conditions given by [5], and that the predictive control law, presented in this work, is practically stabilizing and limited by a good choice of the synthesis parameters. These conditions enable us to preserve the separation principle, and to lead thereafter to a global stabilization via output feedback control.

5. Control Structure

The FOC structure suggested in this work introducing a rotor flux observer based on the sliding mode techniques is given by figure 1. The control through PI regulators is illustrated by figure 1.a whereas that constructed around the RST form predictive regulators is given by figure 1.b.

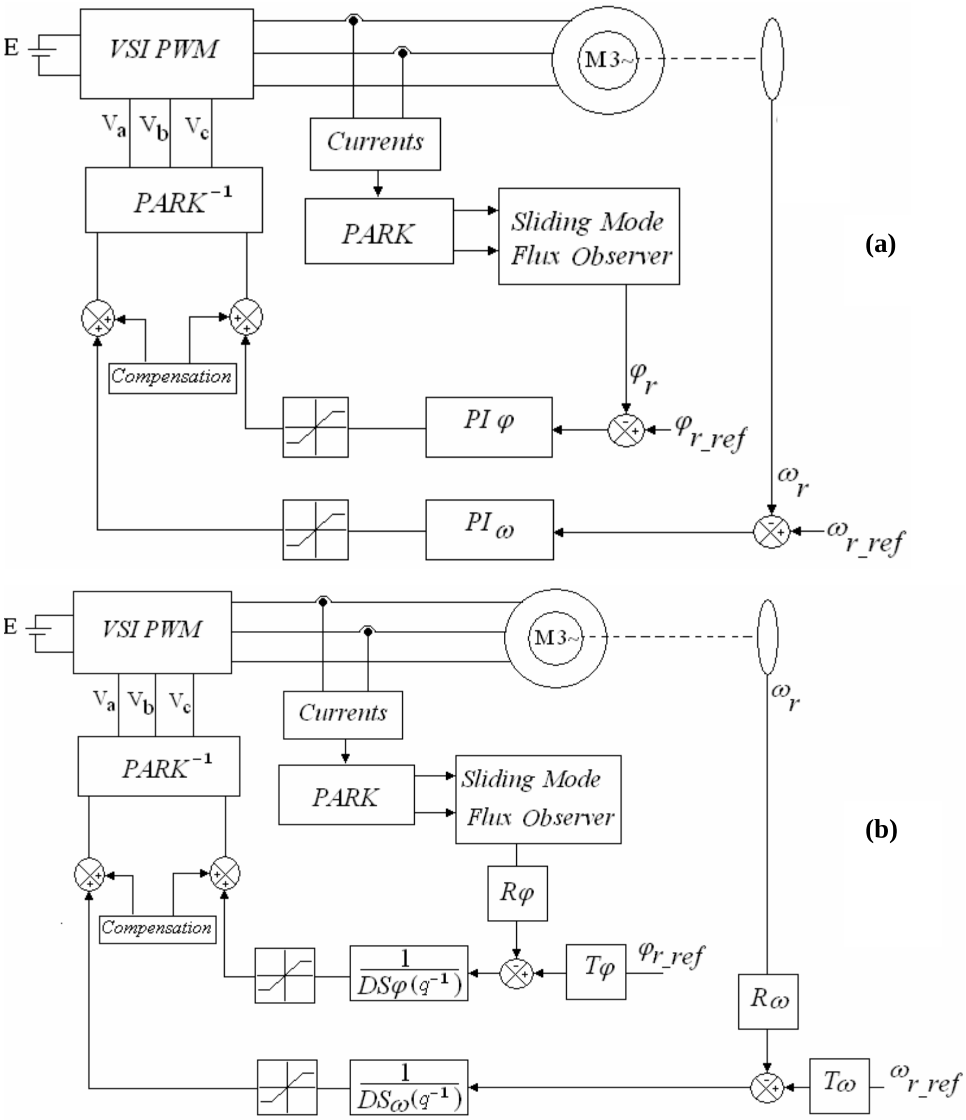

Figure 1 : FOC Structure for an induction motor including a sliding mode flux observer : (a) Structure with PI regulators, (b) Structure with predictive controller.

6. Simulation results

In our simulations, we considerate a two horse power tetra-polar induction motor. The motor parameters are given in the appendix.

The simulated mode relates to a case of a load starting. The load torque is considered proportional to speed (the proportionality constant is equal to 0.067).

The reference speed is taken as being a slope equation $\omega_{r\,ref} = 100\,t$ beginning with t = 2s and limited to 100rd/s. In the same way, the reference flux is a slope equation $\varphi_{ref} = 1{,}48\,t$ and limited to 0.89Wb (nominal flux).

From the permanant established state, we introduced an increase of 50% of the initial rotor resistance value, that is to say a change of resistance of $4.282\,\Omega$ to $6.42\,\Omega$

With an aim of seeking of motor robustness advantage, we simulated the machine response by introducing one rotor resistance adapter. The response of the variation of rotor resistance is illustrated by figure 2. One illustrates on figures 3, 4 and 5 the motor responses operating under FOC without parametric variations, with parametric variations and with rotor resistance adapter respectively.

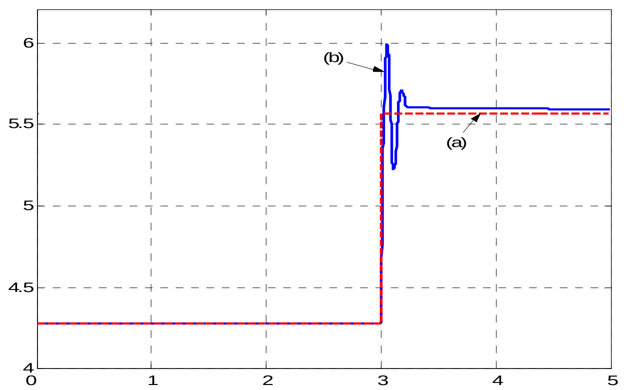

Figure 2 : Response of the variation of rotor resistance following an increase of 50% of its reference value: (a) Actual Resistance (b) Estimated Resistance.

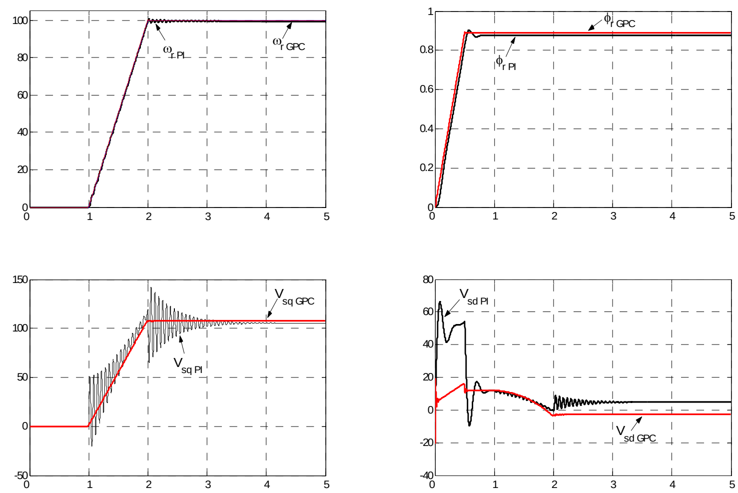

Figure 3: Simulation results for an induction motor operating under DFOC with PI/GPC regulators including a sliding mode observer without rotor resistance variations.
Legend: (3.a): Angular speed vs time, (3.b): rotor flux vs time, (3.c): q-axis control stator voltage vs time and (3.d): d-axis control stator voltage vs time.

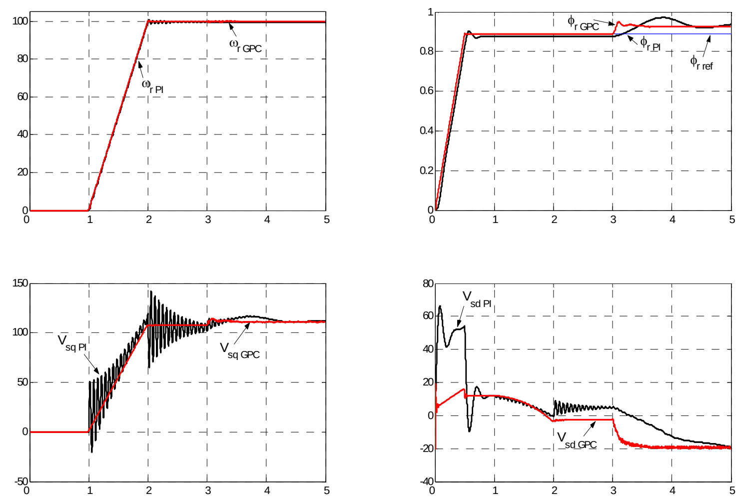

Figure 4: Simulation results for an induction motor operating under DFOC with PI/GPC regulators including a sliding mode observer without rotor resistance adapter. Legend: same as in figure 3.

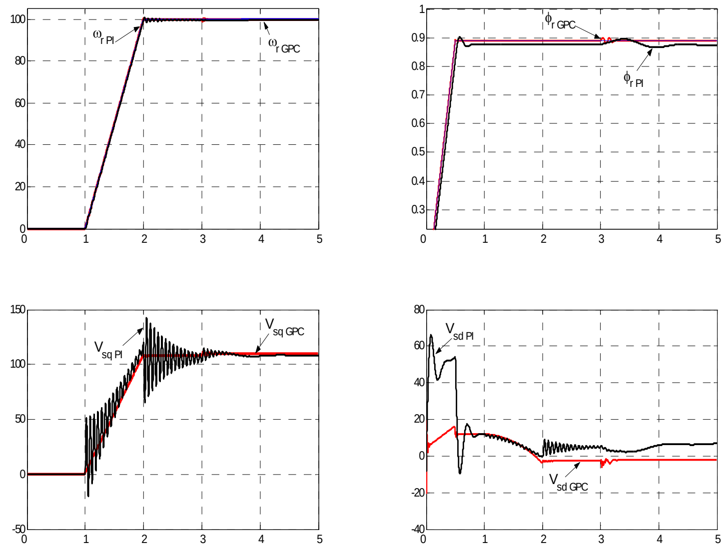

Figure 5: Simulation results for an induction motor operating under DFOC with PI/GPC regulators including a sliding mode observer with rotor resistance adapter. Legend: same as in figure 3.

Considered simulations results, we remark that the GPC response time control is evaluated at 1,94s for the mechanical speed and at 0.47s for the rotor flux, these PI control time is equal to 1,99s and 0.51s for the mechanical speed and the rotor flux respectively. The GPC overshoot is evaluated at 2% for the mechanical speed and at 0.32% for the rotor flux, these PI overshoot is equal to 7% and 1.44% for the mechanical speed and the rotor flux respectively.

One notes according to the simulation results that the variations of the real and reference speed present a similar dynamic in terms of continuation and establishment.This dynamic is insensitive for the parameters variations.

The dynamics of rotor flux depends on the parameters variations. The simulation results show that an increase of 50% of rotor resistance results in a tracking error of 8% for PI regulator and 3% for GPC regulator. The introduction of the algorithm adapter made it possible to correct the imperfections introduced by the rotor resistance variations. It makes it possible to reduce the tracking error to 3% for PI regulator and of less than 1% for the GPC controller.

8. Conclusion

In this paper, we could highlight the FOC performances of induction motor by using two control strategies. The first uses two PI regulators and the second uses two predictive controller for mechanical speed and rotor flux.

The two proposed strategies were applied to an induction motor operating under FOC technique. In these control strategies, a rotor flux observer is developed. The proposed observer is based on the sliding mode technique. In order to deal with parameters variations and the externals perturbations, we have developed a rotor resistance adaptation algorithm based on the *Lyapunov* theory.

The obtained simulation results show a clear improvement on the level of the machine response by use of the predictive controller. High robustness with respect to the parameter variations is checked during integration of rotor resistance algorithm adaptation. Actually, we oriented our work to the implementation the real time control strategy devoted for induction machine.

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Appendix

A.1 Motor Parameters

- Rotor resistance $R_r = 4.282\,\Omega$,
- Stator resistance $R_s = 5.717\,\Omega$,
- Stator inductance $\ell_s = 0.464H$,
- Rotor inductance $\ell_r = 0.464H$,
- Mutual inductance $M = 0.4417H$,
- Total leakage factor $\sigma = 0.09838$,
- Moment of inertia $J = 0.0049 kgm^2$,
- Mechanical viscous damping $f = 0.029$,
- Number of pole pairs $np = 2$.

A.2 Discrete transfer functions parameters

$a_{\varphi 1} = -1.802$ $a_{\varphi 2} = 0.8034$

$b_{\varphi 0} = 4.15e\text{-}5$ $b_{\varphi 1} = 3.862e\text{-}5$

$a_{\omega 1} = -1.769$ $a_{\omega 2} = 0.79$

$b_{\omega 0} = 0.00535$ $b_{\omega 1} = 0.0049$

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