

ANTI-SATURATION CASCADE GENERALIZED PREDICTIVE CONTROL APPLIED IN INDUCTION MOTOR

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Abstract: *Owing to the physical limitations in the variable speed induction motor drives, the windup phenomenon appears and results in performance degradation when the speed controller output is saturated by the current limiter, this happens often when a large change in the speed reference occurs. This paper's aim is to prevent the undesirable effects of the saturation controller by employing a retuned generalized predictive speed controller based on Youla parametrization. The synthesis of an initial GPC controller based on its polynomial equivalent structure is required in the first step. Then, thanks to the Youla parameterization, this controller is retuned considering these saturation constraints reflected on two specifications. The first is a frequency specification on the quadratic component of stator current response to the speed reference which assures a minimal control action in transient periods. And the second is a time domain constraint on the measured speed response to the speed reference that enforce the output response to be within an imposed envelope. These constraints are stated within a linear convex optimization problem. The simulation results proved the efficiency of the resulting predictive speed controller.*

1. INTRODUCTION

The Generalized Predictive Control (GPC) [1] is a modern technique of process control. It has been successfully implemented in many industrial applications for last two decades [2], it is due to its robustness, optimality and ability to face uncertainty. Therefore, GPC applications to electric drives are developed and become interesting in many research

laboratories [3], [4], [5], [11] and [12]. Some authors use the GPC algorithm in cascade form to control multiple loops of the electric motor [14], also others adopt the multivariable GPC formulation in order to control different variables of the system [15].

Even today, the induction motor (IM) stands out as the motor of choice in wide range of applications especially in industrial drives. This, thanks to its higher efficiency, low cost, lower inertia, high initiate torque, and robust architecture. In high performance drive system, the field-oriented control (FOC) is commonly employed and it is the famous technique used in the speed control of the induction motor. It is due to its unique characteristics like high efficiency, good power factor and perfectly reliable [16]. The FOC method guarantees the decoupling of torque and flux control commands of the induction motor, so that the induction motor can be controlled linearly like a separate excited DC motor by means of the FOC scheme which guarantees the decoupling of torque and flux commands of the induction motor. Usually, VSI (Voltage Source Inverters) and (PWM pulse width modulation) are used to drive IM. Thus, the overall control system consists typically of an outer speed control loop and an inner current control loop in cascade. The studies that employed GPC laws to control induction motors, provide very satisfactory results in terms of robustness, optimality and ability to face uncertainty, compared to the classical regulators such as PI/PID controllers. Therefore, without system constraints, the optimal solution in an analytic (explicit) form can be obtained easily and hence the GPC control laws are pre-calculated off-line. As final step, the GPC controller can be synthesized in its numerical polynomial equivalent structure RST. In practice, the mentioned electrical drive is a difficult engineering problem, especially in high speed control, since it suffers from some limits, such as limits on power inverter's and maximum allowable motor current. If the GPC speed controller is designed in a linear region without regard to any constraint, it can generate an exceeded q-axis current reference for the GPC current controller in dynamic and high speed profiles which leads to an over modulation in the inverter. Also, in reality, this current command is limited to a prescribed maximal value depended on the inverter maximum current limit, the overheating of stator windings, and the magnetic saturation.

Consequently, if the GPC speed controller is saturated, the so-called windup phenomenon arises and the close-loop performance will worsen with respect to the expected linear performance, which leads to a slow settling time, a big overshoot on the speed response, and even an instability in the system. Hence, to safeguard the motor and the power electronics, the control of the IM needs the use of sophisticated control that respects these restrictions, while keeping a simple structure designing.

A classical method to handle the windup problem is to take constraints into account at the phase of control design. Nevertheless, this a priori design method has a high computational cost and the obtained control law can be difficult to implement [2], [4].

Another common design strategy is a two-step paradigm: in the first step, a linear control design is performed in which the nominal performance specifications are satisfied

ignoring the saturation constraints, then a supplementary compensator to the initial controller is achieved to minimize the windup impacts on the closed-loop performance which can take place during saturation [18]. The design based on this second method is not complicated design scheme from the viewpoint that the linear performance recovers when the saturation does not occur.

Based on this second approach, the present work intends to employ an effective and simple technique to prevent the controller saturation by relying on the Youla parametrization and based on the results provided by P. Rodriguez and D. Dumur [6]. The authors in [6] presents a unified off-line method using the Youla parametrization to re-tune an initial GPC law while keeping its two degrees of freedom form. This parametrization is fulfilled via convex optimization in terms of two free parameters Q1 and Q2 in which a separation is made between the tracking behavior and the closed loop features. As a result, the Q2 can modify only the input-output transfer function without influencing on the closed loop performances. In this paper, we will investigate this feature on our electrical drive to prevent the wind-up phenomenon.

2. SYSTEM MODELING AND GPC LAW DESIGN

By adopting the traditional assumptions of the vector control of the induction machine, the model in the reference axes d, q related to the rotating field is presented below [7]:

$$\begin{cases} v_{ds} = (R_s + \sigma L_s s) i_{ds} - e_{ds} \\ v_{qs} = (R_s + \sigma L_s s) i_{qs} - e_{qs} \\ \varphi_r = \left(\frac{L_m}{1 + T_r s} \right) i_{ds} \\ C_{em} = \frac{p L_m}{L_r} \varphi_r i_{qs} \\ \Omega = \left(\frac{1}{f_r + J s} \right) (C_{em} - C_r) \end{cases} \quad (1)$$

where:

s	Laplace operator,
v_{ds}, v_{qs}	Stator voltages,
i_{ds}, i_{qs}	Stator currents,
φ_r	Rotor flux,
Ω	Rotation speed,
C_{em}, C_r	Electromagnetic and load torques respectively,
J	Moment of inertia,
L_s, L_r, L_m	Stator inductance, rotor inductance and mutual inductance respectively,

f_r	Friction coefficient
R_s, R_r	The resistance of the stator and of rotor the respectively,
σ	Blondel's dispersion coefficient
$T_r = \frac{L_r}{R_r}$	Rotor time constant

and e_{ds}, e_{qs} represent the voltage compensation terms:

$$\begin{cases} e_{ds} = \sigma w_s L_s i_{qs} \\ e_{qs} = -w_s \frac{L_m}{L_r} \phi_r - \sigma w_s L_s i_{ds} \end{cases} \quad (2)$$

where w_s is the synchronous speed.

We can neglect the electric pole of the induction motor because it is faster than the mechanic pole. Also we consider that the block inverter has neither gain nor dynamics in block scheme of *fig.1*, it is possible to deduce the following transfer functions corresponding respectively to the electric and mechanics modes:

$$\begin{cases} G_i(s) = \frac{1}{(R_s + \sigma L_s s)} \\ G_s(s) = \frac{1}{(f_r + J s)} \end{cases} \quad (3)$$

As the GPC controllers are of discrete type, the transfer functions (3) must be converted into a discrete time transfer functions. Thus, using the ZOH (Zero Order Hold) discretization method, the z-transform of the system transfer functions (3) can be given as follows:

$$\begin{cases} G_i(q^{-1}) = \frac{q^{-1} A_i(q^{-1})}{B_i(q^{-1})} \\ G_s(q^{-1}) = \frac{q^{-1} A_s(q^{-1})}{B_s(q^{-1})} \end{cases} \quad (4)$$

The transfer functions derived above are the models used in the GPC controllers design for the speed and currents respectively. *Fig. 1* illustrates the block diagram of the induction motor drive based on the FOC scheme. The field weakening guarantees that the flux reference diminishes when the motor speed exceeds its nominal value. Using the Park's transformation, the $abs \Rightarrow dq$ block get the i_{as}, i_{bs} et i_{cs} motor stator currents, and the $dq \Rightarrow abs$ block makes the reverse Park's transformation.

In both cases (speed loop or current loops), Tthe GPC control strategy uses for the prediction the CARIMA model (Controlled Autoregressive Integrated Moving Average):

$$A(q^{-1})y(t) = B(q^{-1})u(t) + \frac{\xi(t)}{\Delta(q^{-1})} \quad (5)$$

where $u(t)$ is the control signal, $y(t)$ is the process output, $\xi(t)$ is the zero mean white noise, $\Delta(q^{-1}) = 1 - q^{-1}$, and A and B are polynomials in backward shift operator q^{-1} derived from (5). The predicted output in the j -th prediction step over the costing horizons $N_1 \leq j \leq N_2$ is done by:

$$y(t+j) = \underbrace{F_j(q^{-1})y(t) + H_j(q^{-1})\Delta u(t-1)}_{\text{free response}} + \underbrace{G_j(q^{-1})\Delta u(t+j-1) + J_j(q^{-1})\xi(t+j)}_{\text{forced response}} \quad (6)$$

F_j, G_j, H_j are polynomials determined from solving iteratively Diophantine equations.

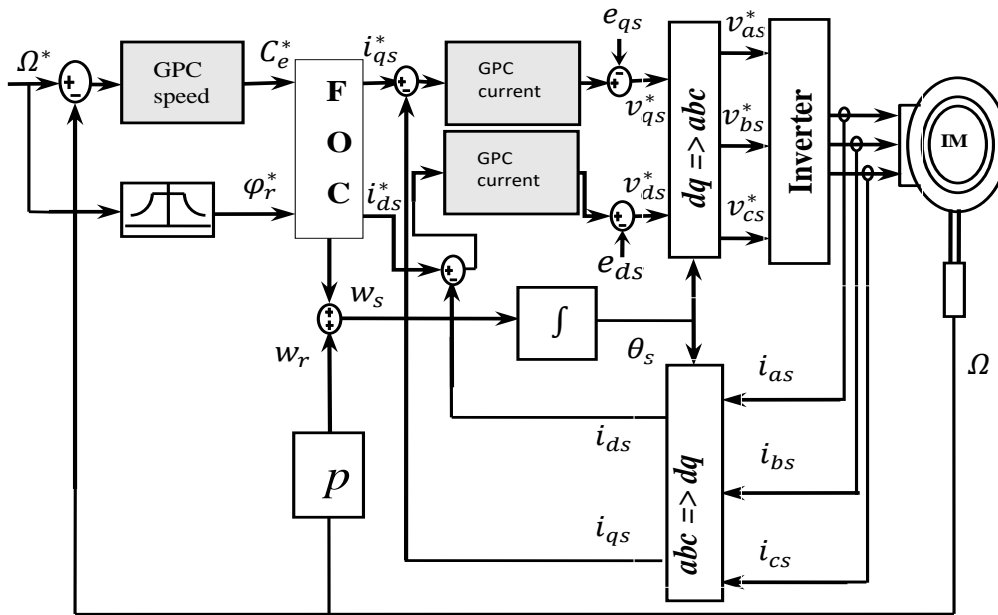


Fig. 1 Block diagram of an induction motor drive based on GPC controllers

The GPC control law is obtained by minimizing the cost function given by:

$$J(N_1, N_2) = \sum_{j=N_1}^{N_2} [\hat{y}(t+j) - w(t+j)]^2 + \lambda \sum_{j=1}^{N_u} [u(t+j-1)]^2 \quad (7)$$

$$\Delta u(t+j) = 0 \text{ for } j \geq N_u$$

where λ is the control weighting factor, w the set-point, N_1 and N_2 are the minimum and maximum costing horizons, and N_u is the control horizon.

The obtained GPC control law can be transformed to a two degrees of freedom RST structure that given as:

$$S(q^{-1})\Delta(q^{-1})u(t) = -R(q^{-1})y(t) + T(q)w(t) \quad (8)$$

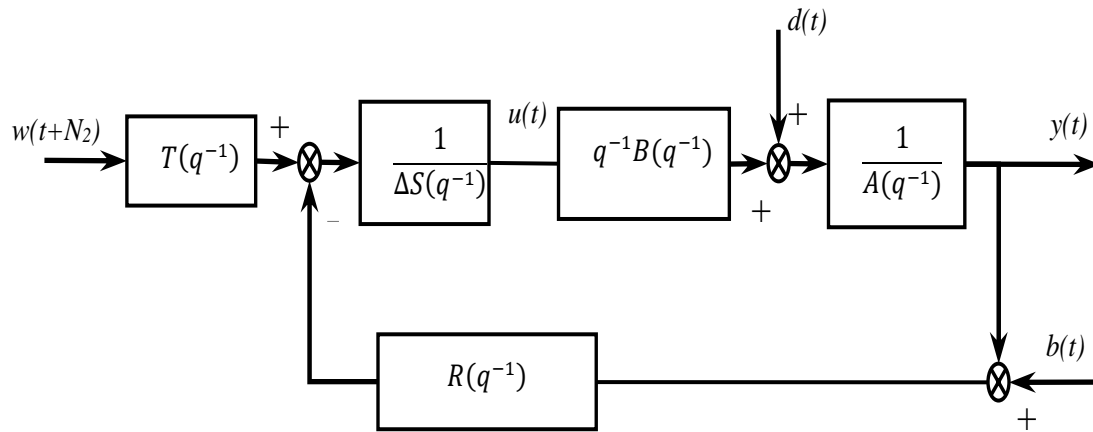


Fig. 2. GPC Equivalent polynomial RST controller

So, three GPC-RST controllers will be synthesized, one for the outer speed control loop denoted (GPC speed) and a pair for the inner current loops denoted (GPC current).

Assuming the design of the initial GPC speed controller has been performed with R_0, S_0, T_0 and N_1, N_2, N_u, λ are tuned to satisfy certain closed loop performance. We notice in the Fig. 1, that the GPC speed controller does not have output magnitude limiter, and therefore, the reference of the electromagnetic torque C_{em}^* , subsequently the current command i_{qs}^* can take values relatively large in transient regimes, especially, in high speed profiles, and, as a consequence, the system drive can be damaged by the large control action. The main goal of this work is to avoid the over values of the currents without incorporating a limiter at the output and without losing the close loop performance obtained by the initial GPC speed controller. It will be carried out by retuning the initial controller based on Youla parameterization. the resulting controller must respect the prescribed limits.

3. RETUNED GPC SPEED USING YOULA PARAMETERIZATION

According to the work presented in [6], The Youla parameterization of the initial GPC speed controller (R_0, S_0, T_0) leads to the following stabilizing polynomials:

$$\begin{cases} T(q^{-1}) = T_0(q^{-1}) - A_0(q^{-1})Q_2(q^{-1}) \\ R(q^{-1}) = R_0(q^{-1}) + \Delta A(q^{-1})Q_1(q^{-1}) \\ S(q^{-1}) = S_0(q^{-1}) - q^{-1}B(q^{-1})Q_1(q^{-1}) \end{cases} \quad (9)$$

where Q_1 and Q_2 are stable transfer functions. The corresponding block diagram is shown in fig. 3.

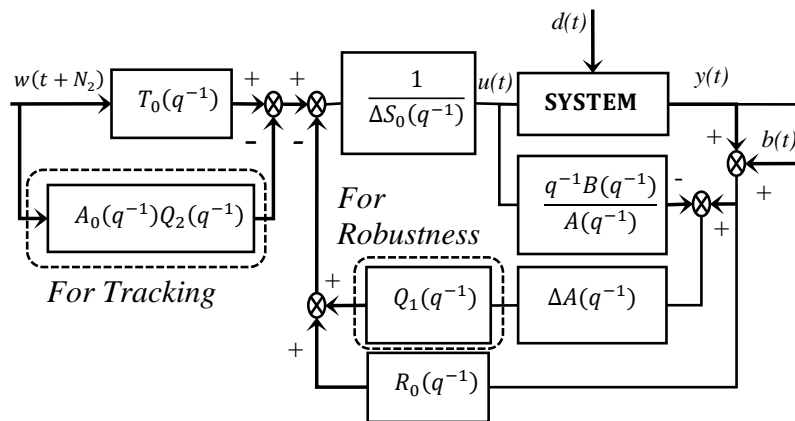


Fig. 3 GPC RST controller with Youla parameterization

From the diagram of the Fig.3, we can conclude that the parameter Q_1 modifies the closed loop features preserving the input-output transfer unchanged, whereas the parameter Q_2 modifies only the input-output transfer function [18]. In the following we set Q_1 to zero since the closed loop performance is fulfilled by the initial controller design, then Q_2 will be used to retune this initial controller modifying the input-output behavior in order to prevent the undesired high control signal at the output of the GPC speed controller.

Q_2 is designed to satisfy two types of specifications: time-domain and frequency specifications. Where the signals w , y , and u are correspond receptively to the speed reference Ω^* , measured speed Ω , and the current command i_{qs}^* .

3.1. Time-domain specifications

Starting from the Fig. 3, let define the close loop transfer function between the input reference w and the output y :

$$H_{yw} = \frac{y}{w} = \frac{T_o q^{-1} B}{A_o A_c} - \frac{q^{-1} A_o B}{A_o A_c} Q_2 \tag{10}$$

$A_o A_c = \Delta A S_0 + q^{-1} B R_0$ represent the closed loop characteristic polynomial partitioned into an observer polynomial A_o and a control polynomial A_c .

Let $s(t)$ be the step response of H_{yw} to the input reference $w(t)$. The main objective of the retuned controller is to make $s(t)$ constrained inside in an imposed template. This template is chosen according to some requirements on the set point response signal. The template is specified by the minimal amplitude $s_{min}(t)$ and maximal amplitude $s_{max}(t)$. Hence, the set of all Q_2 parameters that satisfy this constraint is:

$$C_{env} = \{Q_2 / \Phi_{env}(Q_2) \leq 0\} \tag{11}$$

with:

$$\Phi_{env}(Q_2) = \max(\max_{t \geq 0}(s(t) - s_{max}(t)), s_{min}(t) - s(t)) \tag{12}$$

The time-domain specifications are convex in Q_2 [9] because there is a linear dependency between the transfer function H_{yw} and the Youla parameter Q_2 as given in equation (10).

3.2. Frequency specifications

Let H_{uw} be the transfer function between the input reference w and the control signal u :

$$H_{uw} = \frac{u}{w} = \frac{T_0 A}{A_o A_c} - \frac{A A_o}{A_o A_c} Q_2 \quad (13)$$

The frequency specifications lie in the minimization of the effort control in transient response, these specifications are also convex in Q_2 . This is carried out by minimizing an H_∞ norm of the transfer function H_{uw} weighted by a transfer function W .

In the steady state, the output y must equal the reference w . To ensure that, it is necessary to validate the following relation:

$$\frac{y}{w} = \frac{T_0 q^{-1} B}{A_o A_c} - \frac{q^{-1} B}{A_c} Q_2 \Big|_{q=1} = 1 \quad (14)$$

It is thus necessary that $Q_2 = 0$ for $q = 1$. This can be obtained simply by forcing in Q_2 a term $\Delta = 1 - q^{-1}$ in the numerator:

$$Q_2 = \Delta \cdot Q_2' \quad (15)$$

Finally, the Q_2 design is achieved by a H_∞ norm minimization of the transfer control/set-point under constraints imposed by the time-domain specifications (11):

$$\min_{\substack{Q_2' \in RH_\infty \\ \Phi_{env}(Q_2) < 0}} \left\| \left(\frac{T_0 A}{A_c} - \frac{A \Delta}{A_c} Q_2' \right) W \right\|_\infty \quad (16)$$

Where RH_∞ is the space of all stable and proper transfer functions. This convex optimization problem leads to a Q_2 parameter varying in an infinite dimensional space. A sub-optimal solution of this optimization problem is obtained by considering a finite dimensional sub-space generated by an orthonormal base of discrete stable transfer functions such as a polynomial or FIR (Finite Impulse Response) filter. In this manner, the time-domain constraints and the H_∞ norm minimization can be approximated by a minimization under linear inequality constraints [6].

4. SIMULATION TESTS

In the next, the retuned GPC predictive controller has been applied on the induction motor drive. The induction machine is a squirrel-cage type of 1.1kW and 1500 rpm.

As a first step, three initial GPC controllers have been designed, one for the outer speed control loop (GPC speed) and two for the inner current loops (GPC current) *fig. 1*. The inner system is sampled at $T_i = 0.05\text{ms}$, and the outer system is sampled at $T_s = 1\text{ms}$. According to the rules given in [8], the following tuning parameters are selected:

$$N_1 = 1, N_2 = 30, N_u = 1 \text{ for the speed loop;}$$

$$N_1 = 1, N_2 = 5, N_u = 1 \text{ for the current loop.}$$

Using the simplified model (4), the simulation shows that the GPC speed controller gives a fast tracking response but the control action is very high (reaches about 400% for the nominal value), which is unacceptable in electrical drives. This inconvenient justifies to redesign the initial controller in order to make it able to reduce the control action in the transient response. For that purpose, minimization problem (16) must be solved considering, as temporal specifications, a time domain template for which the system preserves the time response obtained with the initial controller. We chose $N_t = 200$ the number of points of the temporal response taken into account by the template (i.e. horizon of 0.2s). In order to give more weight to high frequencies, a high pass filter is chosen as weighting fundion:

$$W = \frac{1-0.8q^{-1}}{0.2} \quad (17)$$

The research of the Q_2 parameter has been accomplished with 180 points for the transfer norm calculus and with 100 -order polynomial. The resulting polynomial is then approximated by a transfer function [10].

Fig. 4 shows the step response of the simplified model connected to both the initial and modified controller and it also shows the template which must be respected.

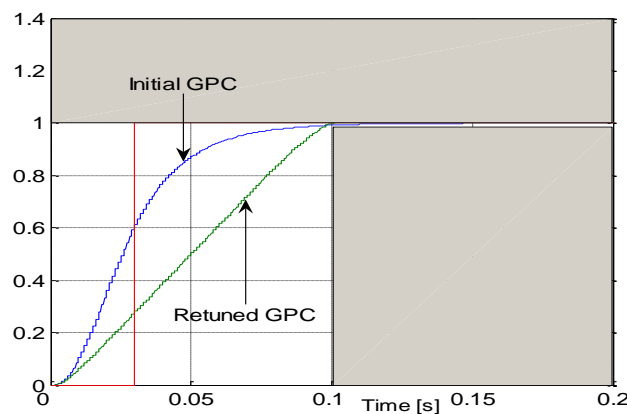


Fig. 4 Step response and time domain template

The temporal response to a step input and to the perturbation at $0.2s$ is shown in *fig. 5*. And *fig. 6* depicts the control signal where the results obtained with the initial controller are superimposed in comparison. We note that the dynamic of the disturbance rejection remains unchanged. In addition, the transitory command is reduced and the dynamics of the input/output behavior respects the imposed template.

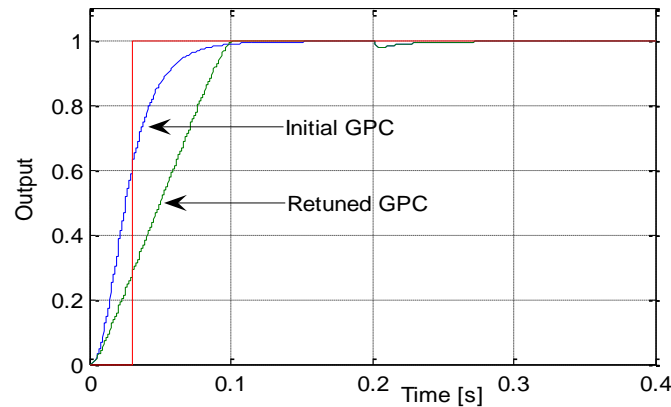


Fig. 5 Temporal response of the system connected to the initial and retuned GPC

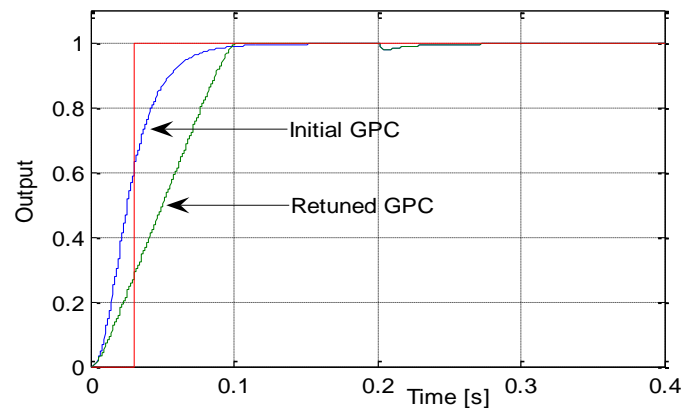


Fig. 6 Control signal of the system connected to the initial and retuned GPC

Now, let's examine this procedure on the complete drive system shown in *fig 1* under various situations. The drive system parameters are given in the Appendix. A rectangular form of motor speed set-point is chosen. Initially, the motor operates under unload conditions with a speed rotor equals 150 rd/s . Between $t = 0.3s$ and $t = 0.8s$, in order to examine the impact of the load condition the load torque is stepped to 5 Nm . The speed is reversed from 150 rd/s at $1s$ to -150 rd/s at $2s$. After that, the motor is stopped.

Figure 7 illustrates the rotor speed and *figure 8* depicts the quadratic component of the stator current i_{qs} that represents the control signal. It is easier to see that i_{qs} is minimized in the transient response thanks to the retuned GPC regulator, while this controller preserves the same time response of the system with zero steady-state error. Moreover, it has no effects on the disturbance rejection dynamic obtained by the initial GPC controller. These results prove that

the proposed design is efficient compared the conventional methods with anti-windup scheme that use a limiter in the output of the controller with compensators.

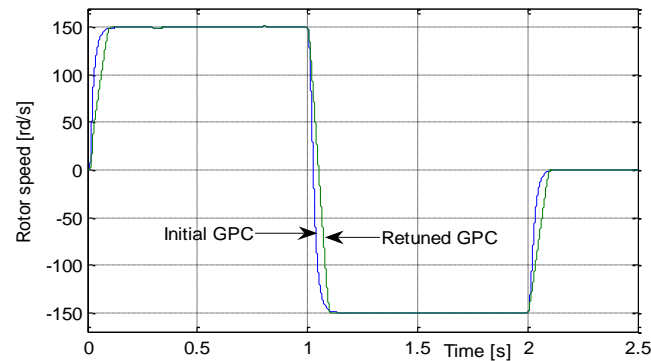


Fig. 7 Rotor speed of the induction motor drive

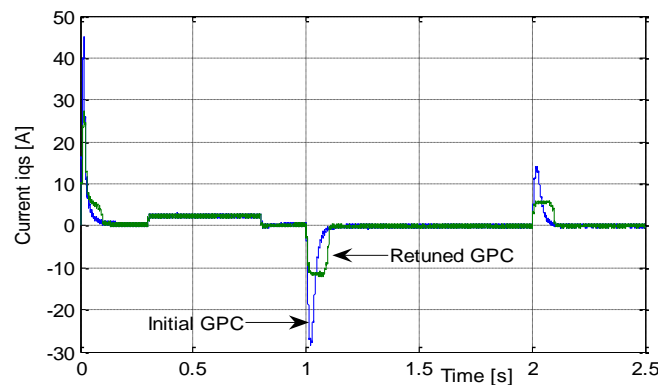


Fig. 8 quadratic component of stator current i_{qs}

5. CONCLUSION

This paper has stated the effect of the high transient current control of an induction machine generated by a GPC speed controller caused by the fast and large step change in the speed reference. For that, three generalized predictive controllers are used to drive the machine, two inner ones to control the currents and an outer one to control the speed. Then, by means of the Youla parameterization, the outer controller has been retuned such that both time-domain and frequency constraints are satisfied. These constraints are formulated as a convex linear optimization framework. We conclude that the resulting controller has two advantages. Firstly, it can minimize the current command in the transitory response, and secondly, it keeps the time response of the system obtained before the modification without changing the closed loop behavior. These results are validated by simulations.

Appendix (Motor parameters):

1.1 kW, 1500 rpm, 220/380 V, 3.5 A, 1.14 Wb, 7 Nm,
 $R_s = 8.1 \Omega$, $R_r = 3.2 \Omega$, $L_s = L_r = 0.48 H$, $L_m = 0.46 H$, $J = 0.006 Kg.m^2$

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