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A High Performance PWM Inverter Voltage-Fed Induction Machines Drive with an Alternative Strategy for Speed Control

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Abstract: In this paper, the method of behaviour model control is applied to an induction motor with its mechanical load in order to increase the robustness of the vector control and to keep its performances despite the presence of perturbations (parameters variations, abrupt load variations, etc.). The idea of the proposed control is interesting and useful. It induces adding supplementary control inputs, which yield the process to follow the model. The distinguished feature of this control design is that it achieves the same performances as the Field Oriented Control without the need for heavy and expensive gain tuning. The proposed strategy minimizes the energy used for the control and ensures the stabilization and excellent tracking performance of the system. Simplicity of the method, minimisation of the required energy and the elimination of the need for gain tuning are the main positive features of the proposed approach.

Keywords: Induction motor drive, Behaviour model control, Vector control, Robust control.

1 Introduction

Electrical drives based on induction motors are the most widely used electromechanical systems in modern industry. Due to their reliability, ruggedness, simple mechanical structure, easy maintenance and relatively low cost, induction motors are attractive for use in a new generation of electrical transportation systems, such as cars, buses and trains. However, from the control point of view, they represent a complex multivariable nonlinear problem and constitute an important area of application for control theory. In fact, induction motors constitute a class of highly coupled and multivariable systems with two control inputs (stator voltages) and two output variables (rotor speed and rotor flux modulus), required to track desired reference signals [1, 2].

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Traditionally, field-orientated control (FOC) has permitted fast transient response by decoupling torque and flux control [3]. This control strategy exploits the fact that in a suitable rotating frame, aligned with the rotor flux space vector, the torque and flux dynamics are decoupled and the induction motor can be efficiently controlled using linear techniques.

Conventional proportional integral (PI) controllers are used to regulate the motor state in the new reference [3]. This traditional controller has the main advantage in comparison with others as nonlinear controller [4-6], adaptive controller [7], robust controller [8, 14] and fuzzy controller [10] to be simply implemented and less expensive.

However, the performance of the (FOC) critically depends on the tuning of the PI controller gains, a task rendered difficult by the high uncertainty on the rotor time constant [11]. In addition, considering the issue of power efficiency of induction motors, the performance of the field oriented control can be improved by using the method of behaviour model control (BMC) [15]. In fact, this method control theory has been used to obtain solutions to the resonant load in machine tool [16], control of machine tool for high speed machining [17], optimisation of control structure of an electric vehicle [18] and holds great promise for other problem areas as well [19, 20].

The main result of this paper is the design of a structure based on the behaviour model control that can be applied to induction motors allowing the synthesis of a robust control law in such a way as to force the angular speed to track given reference values while using the minimum possible control energy. Moreover, the proposed control strategy allows an exact decoupling between speed and flux in all speed ranges and achieves good performance in the presence of perturbations. In addition, the proposed approach achieves comparable results as the FOC strategy without the need for gain tuning.

This paper is organised as follows. The proposed approach is detailed and developed in section 2. The behaviour model of an induction machine is presented in section 3 and is retained for the control strategy. The global structure of the proposed control is explained in section 4. The effectiveness of the approach is examined in section 5 using computer simulation experiments. Some concluding remarks and research perspectives are given in section 6.

2 The Behaviour Model Control

In this paper, the "mod" subscript is used to define model values (e.g. Y_{mod}, the model output). $P(s)$ and $M(s)$ are transfer functions and s the time derivative operator.

2.1 Principle of behaviour model control

The behaviour model control (BMC) needs two controllers and a model of the process (Fig. 1).

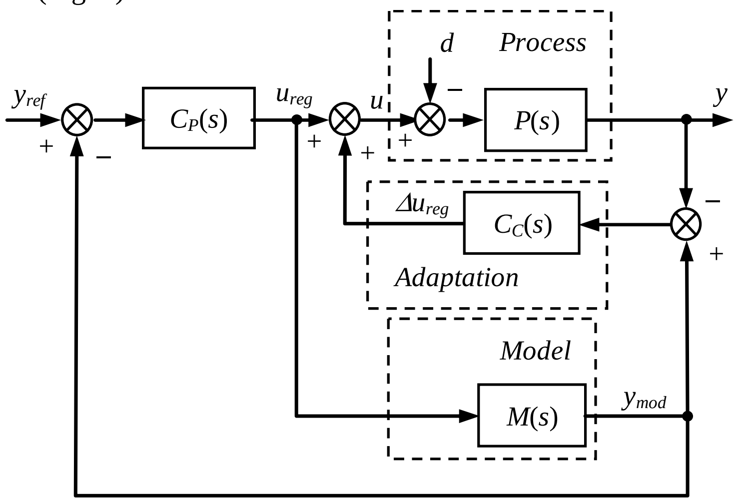

Fig. 1 – *Behaviour model control structure.*

The model $M(s)$ works in parallel with the controlled system (plant) $P(s)$. The main controller $C_P(s)$ is the same as in classical control. The control loop is augmented by a block $C_C(s)$; the so-called secondary controller. The main controller $C_P(s)$ leads to define the control input $U_{reg}(s)$ for the wished output $Y_{ref}(s)$. The secondary controller $C_C(s)$ takes into account the output difference between the real plant and its model. It defines a supplementary control variable $\Delta U_{reg}(s)$, which is added to the process input, in order to suppress this error. Thus, the plant follows the model. That is the reason why this control has been called behaviour model control.

Generally, some precautions are made for the model choice [17 - 20]: model dynamics close to the real plant one and the same transfer function degree for both model and plant.

In our study, one considers the Park model of the rotor field-oriented.

2.2 Controllers tuning

According to the functional diagram (Fig. 1), one can deduce the following expressions:

$$\begin{cases} Y(s) = P(s)\left[U_{reg}(s) + \Delta U_{reg}(s) - D(s)\right] \\ \Delta U_{reg}(s) = C_C(s)\left[M(s)U_{reg}(s) - Y(s)\right] \end{cases}. \tag{1}$$

The calculations deduced from the relations (1) lead to the following expressions:

$$\begin{cases} Y(s) = \dfrac{P(s)\left(1 + M(s)C_C(s)\right)}{1 + P(s)C_C(s)} U_{reg}(s) - \dfrac{P(s)}{1 + P(s)C_C(s)} D(s) \\ Y_{\text{mod}}(s) = M(s)U_{reg}(s) \end{cases}. \tag{2}$$

These relations express the plant output $Y(s)$ and that of the model $Y_{\text{mod}}(s)$ according to the entries, in the presence of the perturbation $D(s)$.

In order to simplify the expression (2), the secondary controller, $C_C(s)$, has to accomplish the assumptions [21]:

$$\begin{cases} \left|M(s)C_C(s)\right| >> 1 \\ \left|P(s)C_C(s)\right| >> 1 \end{cases}. \tag{3}$$

After the simplification of the expression (2), one finds:

$$\begin{cases} Y(s) = M(s)U_{reg}(s) - \dfrac{1}{C_C(s)} D(s) \\ Y_{\text{mod}}(s) = M(s)U_{reg}(s) \end{cases}, \tag{4}$$

this leads to the following result:

$$Y(s) = Y_{\text{mod}}(s) - \frac{1}{C_C(s)} D(s). \tag{5}$$

If the $C_C(s)$ controller is well tuned, the plant output is the same as that of the model, except for the total perturbation. If this equivalent perturbation ($D(s)/C_C$) is negligible, the process output follows the model output perfectly. This condition is written:

$$\frac{D(s)}{C_C(s)} << M(s)U_{reg}(s). \tag{6}$$

From equations (2) and using assumption (3), the output of the real plant becomes:

$$Y(s) = \frac{M(s)C_P(s)}{1 + M(s)C_P(s)} Y_{ref} - \frac{1}{C_C(s)} D(s). \tag{7}$$

The first term of (7) yields to the main loop (Fig. 2), a closed-loop composed by $C_P(s)$ controller and the behaviour model $M(s)$. The second term of (7) shows the attenuation of the external perturbation $D(s)$.

The secondary controller designer uses the secondary closed-loop (Fig. 3), to tune the controller, in order to obtain:

$$Y(s)=\frac{P(s)C_C(s)}{1+P(s)C_C(s)}Y_{\text{mod}}(s)+\frac{P(s)}{1+P(s)C_C(s)}\left(U_{reg}(s)-D(s)\right). \qquad (8)$$

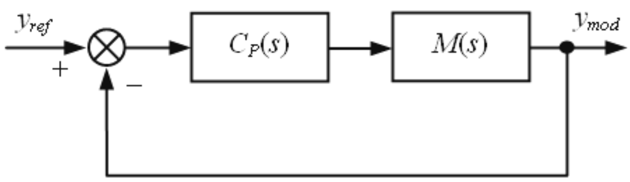

Fig. 2 – *Equivalent BMC main loop.*

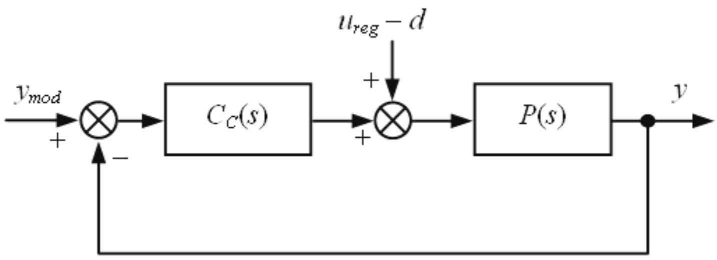

Fig. 3 – *Equivalent BMC secondary loop.*

So, the main controller should be designed based on the plant model. It is more efficient because the model has non-varying and well-known parameters. Therefore, this control yields an important robustness, which is the main advantage of BMC.

3 Application to Induction Motor Control

3.1 Dynamic model of the induction motor

The mathematical model of the induction machine can be defined by the following state representation:

$$\begin{cases} \dfrac{di_{ds}}{dt} = -\gamma i_{ds} + \dfrac{K_L}{T_r}\Phi_r + \omega_s i_{qs} + \dfrac{1}{\sigma L_s}V_{ds} \\ \dfrac{di_{qs}}{dt} = -\gamma i_{qs} - pK_L\omega\Phi_r - \omega_s i_{ds} + \dfrac{1}{\sigma L_s}V_{qs} \\ \dfrac{d\Phi_r}{dt} = -\dfrac{1}{T_r}\Phi_r + \dfrac{L_{sr}}{T_r}i_{ds} \\ \dfrac{d\omega}{dt} = \dfrac{1}{J}\left(T_{em} - F_v\omega - T_L\right) \end{cases} . \tag{9}$$

In which Φ_r is the rotor flux (for rotor field-oriented control), (i_{ds}, i_{qs}) are the stator currents and (V_{ds}, V_{qs}) are the stator voltages which constitute the control inputs, the output to be controlled are the rotor speed ω.

$T_r = \dfrac{L_r}{R_r}$ is the rotor time constant, $\sigma = 1 - \dfrac{L_{sr}^2}{L_s L_r}$ is the total leakage coefficient, $K_L = \dfrac{L_{sr}}{\sigma L_s L_r}$ and $\gamma = \left(R_s + R_r\dfrac{L_{sr}^2}{L_r^2}\right)\Big/\sigma L_s$.

The electromagnetic torque is expressed according to the rotor flux by:

$$T_{em} = p\frac{L_{sr}}{L_r}\Phi_r i_{qs} . \tag{10}$$

From the expression of the speed in (9) and the expression (10), the transfer function of the angular speed is:

$$\Omega(s) = \left(p\frac{L_{sr}}{L_r}\Phi_r I_{qs}(s) - T_L(s)\right)\left(\frac{1}{F_v + Js}\right). \tag{11}$$

In addition, the transfer functions of the stator currents on the *dq*-axes are:

$$\begin{cases} I_{ds}(s) = \left(\dfrac{V_{ds}(s) + E_{ds}(s)}{\sigma L_s\gamma}\right)\left(\dfrac{1}{1+\gamma s}\right) \\ I_{qs}(s) = \left(\dfrac{V_{qs}(s) + E_{qs}(s)}{\sigma L_s\gamma}\right)\left(\dfrac{1}{1+\gamma s}\right) \end{cases}, \tag{12}$$

where E_{ds} and E_{qs} are the stator currents disturbances on the d and q axis respectively and are given by:

$$\begin{cases} E_{ds}(t) = \sigma L_s \left(\dfrac{K_L}{T_r} \Phi_r + \omega_s i_{qs} \right) \\ E_{qs}(t) = -\sigma L_s \left(p\omega K_L \Phi_r + \omega_s i_{ds} \right) \end{cases} . \tag{13}$$

The expressions (11) and (12) are the behaviour models of the angular speed and the stator currents that are used in the proposed control.

3.2 Speed and currents controllers tuning

The main controller has to impose the dynamics of the main loop (Fig. 2). So, it imposes the closed-loop response time, T_{RM} . The secondary controller has to impose the dynamics of the secondary loop, in order to obtain (5). So, it yields the secondary closed-loop response time T_{RP} . In order to have good performances and an easier tuning, we made the assumption of decoupling modes. That means there is a separation in dynamics between current and speed:

$$\begin{cases} (T_{RM})_I < (T_{RM})_\Omega \\ (T_{RP})_I < (T_{RP})_\Omega \end{cases} . \tag{14}$$

But there is another condition on the secondary controller tuning, the inequality (3). It means that the response time (T_{RP}) of the secondary closed-loop must be smaller than the response time (T_{RM}) of the main closed-loop. This condition must be checked in the case of two overlap loops (in our case the speed loop and that of current on *q*-axis), one will thus write:

$$(T_{RP})_I < (T_{RM})_I < (T_{RP})_\Omega < (T_{RM})_\Omega . \tag{15}$$

For the behaviour corrector of the internal speed closed-loop and that of i_{ds} , one can use a PI loop of the form:

$$C_C = -K_P \left(y - y_{\text{mod}} \right) - K_I \int_0^t \left(y(\tau) - y_{\text{mod}}(\tau) \right) d\tau . \tag{16}$$

The gains K_P and K_I can be tuned in order to guarantee specific performances.

According to (13), the disturbance on the q -axis is a slope, the behaviour corrector that one proposes is an action proportional and integral following form:

$$C_C(s) = \frac{C_2 s^2 + C_1 s + C_0}{s^2} . \tag{17}$$

For the speed and the two currents (i_{ds} and i_{qs}), the $C_P(s)$ parameters are imposed by pole placement.

3.3 Structure of the behaviour model control of an induction machine

The structure of the BMC control of the induction machine is represented by the Fig 4. One notes on this figure the appearance of the blocks based on the BMC for the speed and the two stator currents on the d and q axis. It is noted that $C_C(s)$ for the two currents loops are not the same ones, because the disturbances are different on the two axes.

By comparison with the field-oriented control, the decoupling block between the two stator currents is not represented because the structure of BMC compensate the effect of the electric disturbances (E_{ds} and E_{qs}) on the dq-axes.

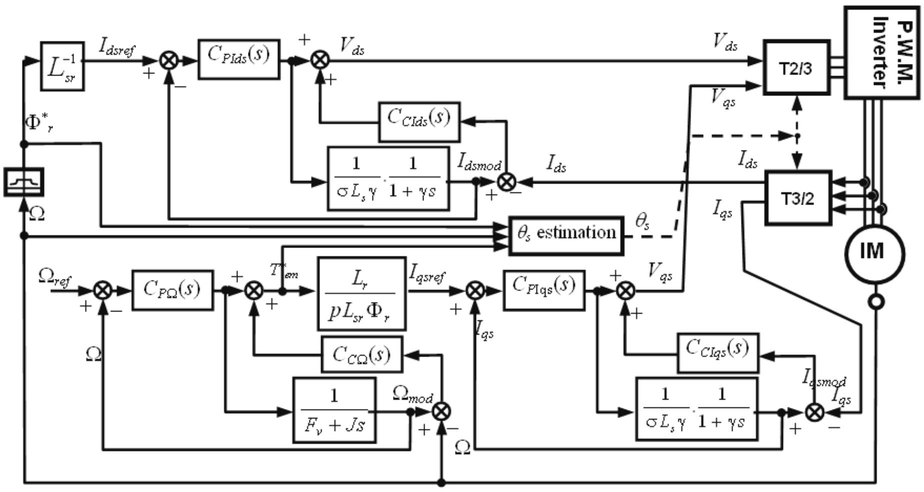

Fig. 4 – *Behaviour model control of an induction machine.*

The estimation of the angle θ_s is important for the orientation control of the reference frame. Its estimate is based on the measurement of the angular speed and the knowledge of slip frequency, which is not accessible but can be estimated (Fig. 5). The following relation defines this estimator:

$$\omega_{sl} = \frac{R_r}{p} \frac{T_{em}^*}{\left(\Phi_r^*\right)^2} . \tag{18}$$

This expression has the advantage of being an estimator of slip frequency, less related to the machine parameters, except rotor resistance. That is due to the knowledge of the references of the electromagnetic torque, resulting from BMC speed loop, and rotor flux.

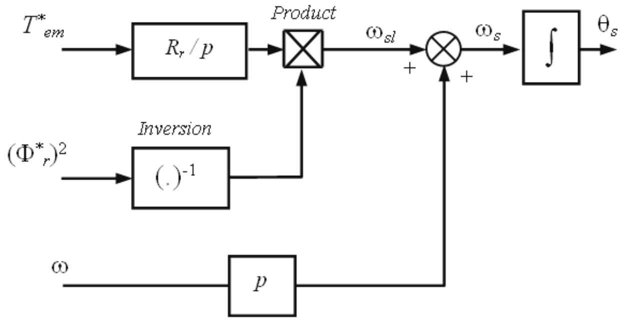

Fig. 5 – *Block of* θ_s *estimation.*

4 System Performances Evaluation

Initially, one applies changes of references to highlight the effectiveness of the control speed by the BMC. Fig. 6 shows the performances of the behaviour model control of a three-phase induction machine. On this figure, the speed follows its reference and the control of the currents on the two axes makes it possible to control rotor flux and the torque under these operating conditions.

One notice for a well-known process with constant parameters and well-compensated disturbances, a conventional control (PI) is as powerful as the behaviour model control. The difference comes from the capacity to keep these trajectories in spite of the external perturbations and the parameters variations.

The speeds trajectories in the case of the rotor inertia variations are illustrated by Figs. 7 and 8. According to Fig. 7, with a conventional control (PI), the speed trajectory changed and does not present the same response time. On the other hand, the BMC preserve better its trajectory (Fig. 8).

Fig. 9 shows that with BMC the process is disturbed less by an external perturbation (load torque), compared to the conventional control (PI). Moreover, it is noticed that the BMC bring means for better controlling the transitory error due to an external disturbance.

Figs. 10 and 11 demonstrate the influence of the dry torque. The influence of this resistive torque is significant in the area of null speed, in particular during the change of speed sign, which influences considerably in the case of the position control of the induction machine. To highlight this phenomenon, one carries out a change of instruction, the speed passing from 100 rad/s to 0 rad/s. If the adjustment of both correctors is made to have a going beyond, speed must become negative before being stabilised on the end value. One thus carries out a variation speed in the area of zero speed. The traditional control does not have the dynamic required by the adjustment (Fig. 10). The dry torque manages to slow down the machine, in spite of correctors adjustment. On the other hand, behaviour model control keeps the same performances despite the existence of the dry torque (Fig. 11). Thus, this control has an effect of the "linearization" of the process response [23], because the behaviour model selected is linear.

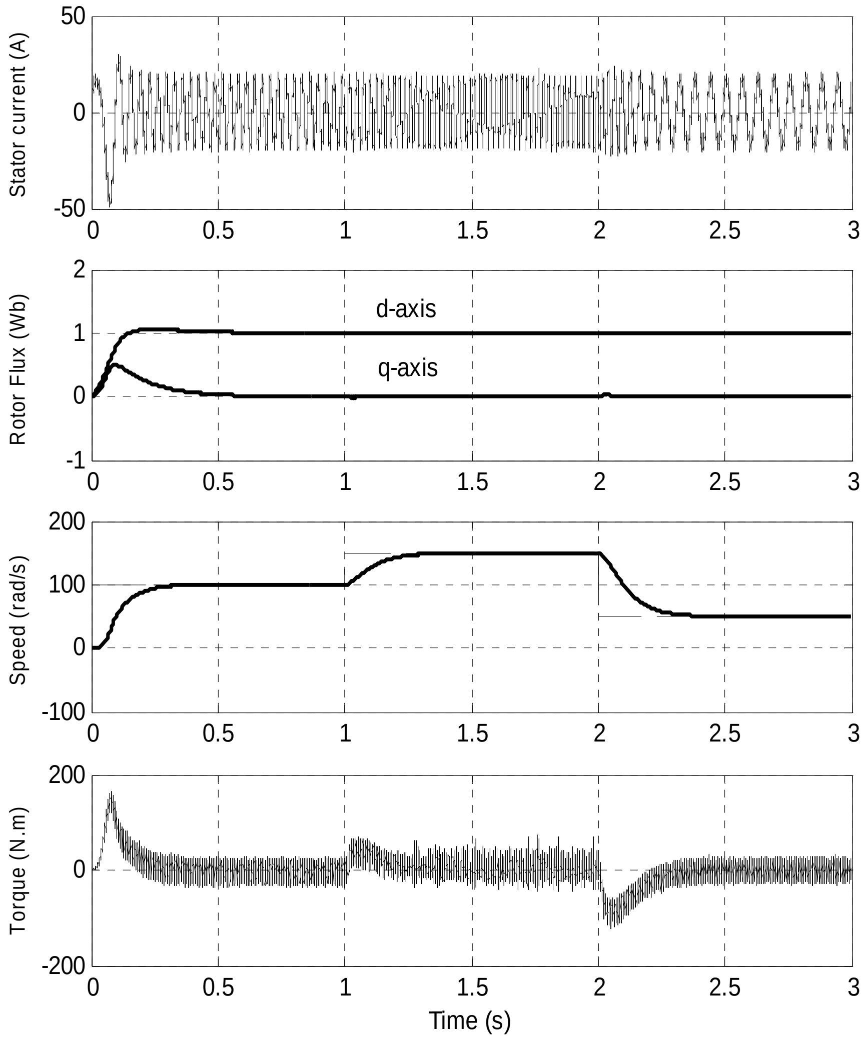

Fig. 6 – *BMC control of induction machine with speed reference variation (100, 150 and 50 rad/s).*

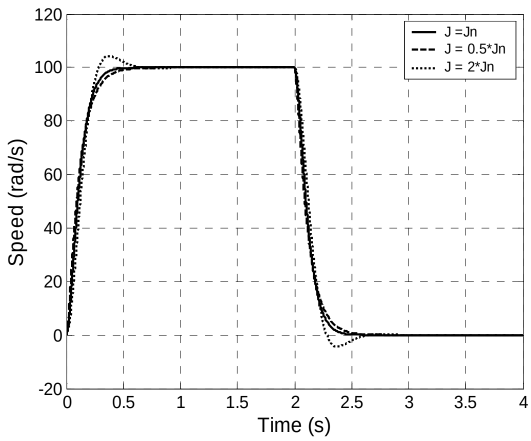

Fig. 7 – *Conventional speed control – rotor inertia variation.*

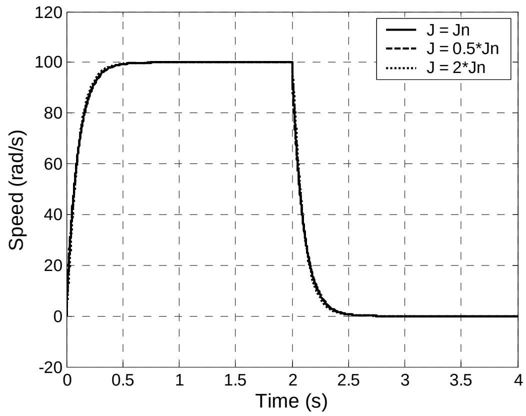

Fig. 8 – *BMC speed control – rotor inertia variation.*

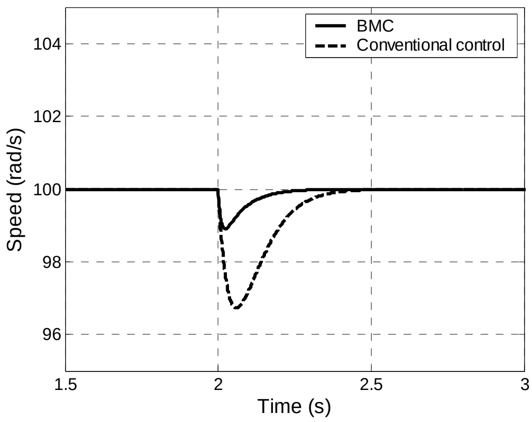

Fig. 9 – *Speed response –Application of a load torque (20 N.m) at t = 2s.*

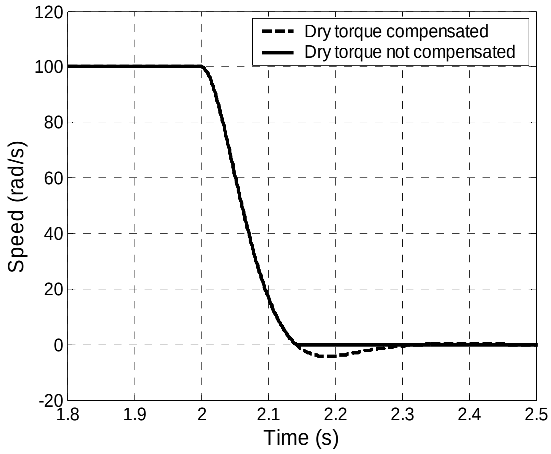

Fig. 10 – *Influence of the dry torque – conventional control.*

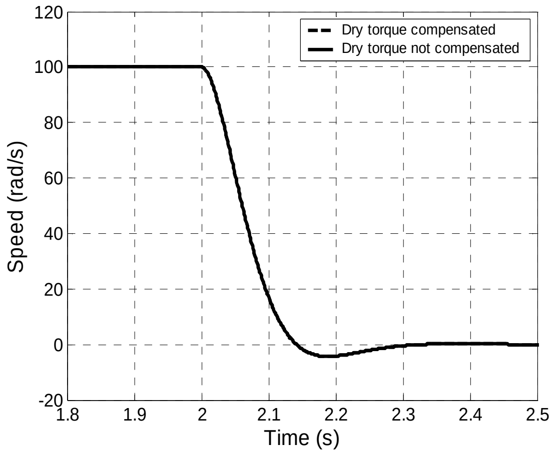

Fig. 11 – *Influence of the dry torque – BMC control.*

5 Conclusion

The present work has demonstrated that comparable performances to the classical Field Oriented Control can be obtained using the behaviour model control without the need for heavy and expensive gain tuning. The model is deduced from the Park model. The dynamic decoupling of the flux and speed variables was achieved using BMC structure. Based on simulation experiments results we have demonstrated the tracking performance and disturbance rejection capabilities of the proposed controller, a solution for current components decoupling without the need of compensation and the good dynamic behaviour of the overall system.

In our process, there is a separation in dynamics of at least one decade between current and speed, so the electrical dynamics can be disregarded, as seen from the mechanical dynamics. Hence, the BMC current control can be omitted in a speed control, but it is necessary in a torque control.

Simplicity of the overall scheme, minimization of the required energy and the elimination of the need for gain tuning are the main positive features of the proposed approach. In our future work, we shall focus on the estimation by the behaviour model structure of non-measurable parameters and state variables.

Appendix 1 – Nomenclature

(d,q)	Rotating reference frame
X_{dq}	Components of a vector in the rotating reference frame
V_{ds}, V_{qs}	Stator voltages
i_{ds}, i_{qs}	Stator currents
Φ_{dr}, Φ_{qr}	Components of the rotor flux
Φ_r	Rotor flux magnitude
R_s, R_r	Stator and rotor resistance
L_s, L_r	Stator and rotor inductance
L_{sr}	Mutual inductance
ω_s	Stator electric angular pulsation
ω_{sl}	Slip frequency
θ_s	Angular position of the reference frame
ω	Mechanical speed
T_{em}	Electromagnetic torque
T_L	Load torque
J	Rotor inertia
F_v	Viscous friction coefficient

Appendix 2 – Induction Motor Data

Rated power	P_n	3.7	KW
Rated speed		1460	rpm
Rated Voltage		220/380	V
Number of pole pairs	p	2	
Stator resistance	R_s	1.126	Ω
Rotor resistance	R_r	0.11	Ω
Stator inductance	L_s	0.17	H
Rotor inductance	L_r	0.015	H
Mutual inductance	L_{sr}	0.048	H
Rotor inertia	J	0.135	Kgm²
Viscous friction coefficient	F_v	0.0018	I.S

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