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Fuzzy Adaptive Control of a Series Connected Tow-Motor Six-Phase Driver System with Seven-Level Single Inverter Supply

***Abstract**. This paper presents a method of behaviour model control (BMC) combined with fuzzy logic and a seven-level six-phase inverter to achieve mainly a high performance and to increase the robustness of the vector control and to keep its performances despite the presence of perturbations (parameters variations, abrupt load variations, etc.) of a series-connected two motor six phase drive system with single inverter supply. The idea of the proposed control is to induce adding supplementary control inputs, which yield the process to follow the model. The best of 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 effectiveness of the proposed behaviour model control in conjunction with the fuzzy logic is confirmed through the application of different load torques for wide speed range operation. Comparison between fuzzy behaviour model control, fuzzy logic controller (FLC) and conventional controller (PI) of the proposed two-motor drive is provided. The simulation results confirm also that, the validity and effectiveness of the control strategy proposed in both terms of performance and robustness (rotor inertia variations J1=5J1nominal) of the provision of such an adaptive control for electrical drives with the two machines of the system.*

Streszczenie.** W artykule opisano wykorzstanie metody BMC (behaviour model control) wspartej logiką romytą w siedmiopoziomowym sześciofazowym przekształtniku w celu osiągnięcia dobrych parametrów I zwiększenia odpoenośei w obecności zakłóceń. Osiąnioęto podobne parametry co w przypadku stosowania metody FOC (field oriented control) bez konicznośei stosowania dopasowania wzmocnienia. **Siedmiopoziomowy sześciofazowy przekształtnik użyty do sterowania szeregowo połaczonych dwóch silników.

Keywords: Multi-Machines System (MSCS), Vector Control, Fuzzy Adaptive Control, Behaviour Model Control (BMC).
Słowa kluczowe: system wielomaszynowy, przekształtnik, logika rozmyta, sterowanie adapatacyjne.

Introduction

Many industrial applications, such as the textile industry, paper mills, robots, require several electric motors. The first generation of multi-machine systems is limited in two configurations. The first configuration consists of a continuous stage which supplies multiple three phase inverters connected in parallel, where each inverter supplies a three-phase machine [1-3]. The control of each machine is independent via its inverter and its control algorithm while the second one is a system with parallel-connected three-phase motors fed from a single three-phase VSI, for this structure, the machines must have the same rotation speed and undergo the same load torque. These conditions make the system usable in a limited number of applications [4-6]. Furthermore, the independent control of each machine is impossible in this configuration. As a solution to this problem, this paper shows the possibility of independently controlling a multi-machine power system with the supply coming from only one inverter. Such an independent control is enabled by using an appropriate series connection of stator windings of multi-phase motors and vector control principles. The basic principles of the concept emerge from the fact that multi-phase machines, regardless of the number of phases, require only two currents for independent flux and torque control. Therefore, there are additional currents, which can be used to control other machines in the group

An appropriate series connection of the stator windings converts the flow/ torque production currents of a machine into flow / torque production currents for other machines, thus allowing independent control of each motor in the group using vector control. An appropriate series connection of the stator windings converts the flux/torque producing currents of one machine into the non-flux/torque producing currents for the other machines, allowing independent control of each multi-phase motor using a vector control scheme [7-12]

The concepts of the two multi-phase two-motor series-connected drive configurations are considered in detail in this paper. The presented drive system comprising of a symmetrical six-phase machine, the second is three-phase machine and a six-phase inverter (VSI). In[12], a speed adjustment is carried out using a fuzzy controller but its capacity of robustness remains moderately limited, because it loses its property for the large ranges of parametric variations (speed for example) given this disadvantage. The main result of this work is to present a new form of adaptive control by fuzzy logic controllers (currents and speed) to increase its robustness (rotor inertia variations) in such a way as to force the angular speed to track given reference values while using the minimum possible control energy. Moreover, an exact decoupling between the speed and the flux is realized by this strategy in all the speed ranges and makes it possible to obtain best performances in the presence of disturbances.

This paper consists of six sections. The first section will be devoted to modeling of the two-machines drive. Application of the indirect control by orientation of the rotor flux in order to realize the decoupling between the torque and the flux of the two machines, by using regulators and FLC is represented in section 3.The proposed approach is detailed and developed in section 4. The effectiveness of the approach is examined in section 5 using computer simulation experiments. Some concluding remarks are given in section 6.

Modeling of Multi-machine System

The drive system is composed by two induction machines. The first one is a symmetrical six-phase induction motor I.M1 which its windings are series connected with that of a second three-phase induction motor I.M(2). The two motors are supplied by a single power converter which is a six-phase Voltage Source Inverter (VSI).

Fig. 1 presents the connecting and suppling schematic of the two motors and the converter [7], [9]. The six-phase machine has the spacial displacement between any two consecutive stator phases equal to 60° (i.e. $\alpha=2\pi/6$).

Only phases 1, 3 and 5 are used by the second machine I.M2, this phases are electrically displaced to each other by and angle of $2\pi/3$.

We note that a simple series connection of stator windings fails to ensure the desired performances. A solution is adopted to overcome this constraint consists of using an adequate stator windings transposition [10], [11]. This transposition resides of connecting in one point each two (electrically displaced to each other by π) of six-phase windings and connect them in series with the windings of the I.M1 [12], [14].

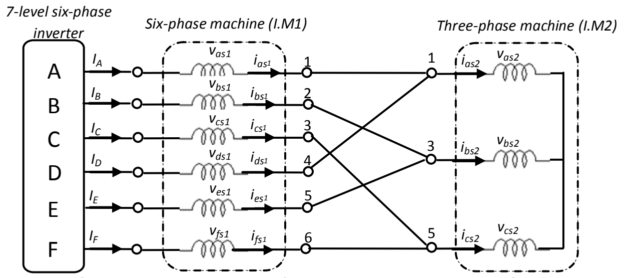

Fig. 1. Connection diagram for series connection of a six-phase and a three-phase machine

In this way, currents pass through the six-phase windings going to neutralize at the connecting point. And in the same context, the current passing through the one winding of I.M2 will be the half when passing through the windings of I.M1. This will generate in air-gap of the I.M1 a two (equal in magnitude and opposed in phase) Magneto-Motive Force (MMF). Therefore, a natural decoupling of the two motors will be possible by adopting the connection diagram shown in Fig. 1.

According to Fig. 1, the stator and rotor voltages of the two machines can be written as follows [1], [4]:

(1) $$[v_s]=\begin{bmatrix} V_A \\ V_B \\ V_C \\ V_D \\ V_E \\ V_F \end{bmatrix}=\begin{bmatrix} v_{as1}+v_{as2} \\ v_{bs1}+v_{bs2} \\ v_{cs1}+v_{cs2} \\ v_{ds1}+v_{as2} \\ v_{es1}+v_{bs2} \\ v_{fs1}+v_{cs2} \end{bmatrix}$$

The relationship between the current source and the stator currents of each machine are given as follows:

(2) $$\begin{aligned}[i_s]&=\begin{bmatrix} I_A & I_B & I_C & I_D & I_E & I_F \end{bmatrix}^T \\ &=\begin{bmatrix} i_{as1} & i_{bs1} & i_{cs1} & i_{ds1} & i_{es1} & i_{fs1} \end{bmatrix}^T \\ &=[i_{s1}]\end{aligned}$$

(3) $$[i_{s2}]=\begin{bmatrix} i_{as2} \\ i_{bs2} \\ i_{cs2} \end{bmatrix}=\begin{bmatrix} I_A+I_D \\ I_B+I_E \\ I_C+I_F \end{bmatrix}$$

The electrical equations:

(4) $$\begin{cases} [V_{sk}]=[R_{sk}].[i_{sk}]+\dfrac{d}{dt}[\varphi_{sk}] \\ [0]=[R_{rk}].[i_{rk}]+\dfrac{d}{dt}[\varphi_{rk}] \end{cases}$$

Where

(5) $$\begin{cases} [\varphi_{sk}]=[L_{ssk}].[i_{sk}]+[M_{srk}].[i_{rk}] \\ [\varphi_{rk}]=[L_{rrk}].[i_{rk}]+[M_{rrk}].[i_{sk}] \end{cases}$$

Knowing that k=1 for the I.M1 and k=2 for the I.M2
With:

$$[R_{seq}]=[R_{s1}]+\begin{bmatrix} [R_{s2}] & [R_{s2}] \\ [R_{s2}] & [R_{s2}] \end{bmatrix};$$

$$[L_{seq}]=[L_{s1}]+\begin{bmatrix} [L_{s2}] & [L_{s2}] \\ [L_{s2}] & [L_{s2}] \end{bmatrix}$$

Modeling of Multimachines System into three subspaces(α,β), (x,y), ($o+,o-$):

The original six dimensional systems of the MSCS can be decomposed into three orthogonal subspaces, (α, β), (x, y) and ($o+$, $o-$) [1], using the following transformation $X_{\alpha\beta o}=[T_6(\alpha)]^{-1}.X_{abc}$ and $X_{dqo}=[T_6(\alpha)]^{-1}X_{\alpha\beta o}$

Where: X represents stator currents, stator flux, stator voltages in MSCS.

The matrix $[T_6(\alpha)]$ is given by:

(6) $$[T_6]=\frac{1}{\sqrt{3}}\begin{bmatrix} 1 & cos(\alpha) & cos(2\alpha) & cos(3\alpha) & cos(4\alpha) & cos(5\alpha) \\ 0 & sin(\alpha) & sin(2\alpha) & sin(3\alpha) & sin(4\alpha) & sin(5\alpha) \\ 1 & cos(2\alpha) & cos(4\alpha) & cos(6\alpha) & cos(8\alpha) & cos(10\alpha) \\ 0 & sin(2\alpha) & sin(4\alpha) & sin(6\alpha) & sin(8\alpha) & sin(10\alpha) \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1/\sqrt{2} & -1/\sqrt{2} & 1/\sqrt{2} & -1/\sqrt{2} & 1/\sqrt{2} & -1/\sqrt{2} \end{bmatrix}$$

(7) $$[T_3]=\sqrt{\frac{2}{3}}\begin{bmatrix} 1 & cos\,2\alpha & cos\,4\alpha \\ 0 & sin\,2\alpha & sin\,4\alpha \\ \dfrac{1}{\sqrt{2}} & \dfrac{1}{\sqrt{2}} & \dfrac{1}{\sqrt{2}} \end{bmatrix}$$

Application of the transformations matrix (6) and (7) in conjunction with the first row of (4) lead to the decoupled model of the six-phase two-motor drive system. Source voltage equations that include equations of the two stator windings connected in series can be given as:

$$\begin{cases} V_{s\alpha}=R_{s1}i_{s\alpha1}+L_{s1}\dfrac{di_{s\alpha1}}{dt} \\ \quad +M_1\dfrac{d}{dt}\left(\cos(\theta_1)i_{r\alpha1}-\sin(\theta_1)i_{r\beta1}\right) \\ V_{s\beta}=R_{s1}i_{s\beta1}+L_{s1}\dfrac{di_{s\beta1}}{dt} \\ \quad +M_1\dfrac{d}{dt}\left(\sin(\theta_1)i_{r\alpha1}+\cos(\theta_1)i_{r\beta1}\right) \\ V_{sx}=R_{eq}i_{sx1}+(l_{s1}+2L_{s2})\dfrac{di_{sx1}}{dt} \\ \quad +\sqrt{2}\left[M_2\dfrac{d}{dt}\left(\cos(\theta_2)i_{r\alpha2}-\sin(\theta_2)i_{r\beta2}\right)\right] \\ V_{sy}=R_{eq}i_{sy1}+(l_{s1}+2L_{s2})\dfrac{di_{sy1}}{dt} \\ \quad +\sqrt{2}\left[M_2\left(\sin(\theta_2)i_{r\alpha1}+\cos(\theta_2)i_{r\beta2}\right)\right] \\ V_{so+}=R_{eq}i_{so+1}+(l_{s1}+2L_{s2})\dfrac{di_{so+1}}{dt} \\ V_{so-}=R_{eq}i_{so-1}+l_{s1}\dfrac{di_{so-1}}{dt} \end{cases}$$

(8)

Rotor voltage equations of six-phase machine and three-phase machine are:

(9) $$\begin{cases} 0=R_{r1}i_{r\alpha1}+L_{m1}\dfrac{di_{s\alpha1}}{dt}+L_{r1}\dfrac{di_{r\alpha1}}{dt}+\omega_{r1}(L_{m1}i_{s\beta1}+L_{r1}i_{r\beta1}) \\ 0=R_{r1}i_{r\beta1}+L_{m1}\dfrac{di_{s\beta1}}{dt}+L_{r1}\dfrac{di_{r\beta1}}{dt}+\omega_{r1}(L_{m1}i_{s\alpha1}+L_{r1}i_{r\alpha1}) \end{cases}$$

(10) $$\begin{cases} 0=R_{r2}i_{r\alpha2}+\sqrt{2}L_{m2}\dfrac{di_{sx1}}{dt}+L_{r2}\dfrac{di_{r\alpha2}}{dt}+\omega_{r2}(\sqrt{2}L_{m2}i_{sy1}+L_{r2}i_{r\beta2}) \\ 0=R_{r2}i_{r\beta2}+\sqrt{2}L_{m2}\dfrac{di_{sy1}}{dt}+L_{r2}\dfrac{di_{r\beta2}}{dt}-\omega_{r2}(\sqrt{2}L_{m2}i_{sx1}+L_{r2}i_{r\alpha2}) \end{cases}$$

with:

$$(11)\qquad \begin{cases} L_{s1} = l_{s1} + \frac{3}{2} L_{ms1} \\ M_1 = \frac{3}{\sqrt{2}} L_{sr1} \\ L_{r1} = l_{r1} + \frac{3}{2} L_{mr1} \end{cases} ; \quad \begin{cases} L_{s2} = l_{s2} + \frac{3}{2} L_{ms2} \\ M_2 = \frac{3}{\sqrt{2}} M_{sr2} \\ L_{r2} = l_{r2} + \frac{3}{2} L_{mr2} \end{cases}$$

Application of (6) in conjunction with (1) yields:

$$(12)\qquad \begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \\ v_{sx} \\ v_{sy} \\ v_{so+} \\ v_{so-} \end{bmatrix} = [T_6] \begin{bmatrix} v_{sa1} + v_{sa2} \\ v_{sb1} + v_{sb2} \\ v_{sc1} + v_{sc2} \\ v_{sd1} + v_{sa2} \\ v_{se1} + v_{sb2} \\ v_{sf1} + v_{sc2} \end{bmatrix} = \begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \\ v_{sx1} + \sqrt{2} v_{s\alpha 2} \\ v_{sy1} + \sqrt{2} v_{s\beta 2} \\ v_{so+} \\ v_{so-} \end{bmatrix}$$

And

$$(13)\qquad \begin{cases} i_{s\alpha} = i_{s\alpha 1} \\ i_{s\beta} = i_{s\beta 1} \end{cases} ; \quad \begin{cases} i_x = i_{sx1} = \frac{i_{s\alpha 2}}{\sqrt{2}} \\ i_y = i_{sy1} = \frac{i_{s\beta 2}}{\sqrt{2}} \end{cases} ; \begin{cases} i_{o+} = i_{so+1} \\ i_{o-} = i_{so-1} \end{cases}$$

Torque equations of the two machines are:

$$(14)\qquad \begin{cases} T_{em1} = P_1 M_1 (i_{rd1} i_{sq1} - i_{sd} i_{rq1}) \\ T_{em2} = P_2 M_2 (i_{rd2} i_{sy1} - i_{sx} i_{rq2}) \end{cases}$$

As can be seen to equations (10)-(14) and (18), that flux/torque producing stator currents of the six-phase machine are the source (α, β) current components, while the flux/torque producing stator currents of the three-phase machine are the source (*x, y*) current components. This indicates the possibility of independent vector control of two machines. It therefore follows that independent vector control of the two machines can be realized with a single six-phase inverter.

Vector Control of the two-Motor Drive

With the transformation (8), the components of the plane (α, β) to equations (10)-(14) can be expressed in the (*d, q*) plane. The two series-connected machines can be controlled independently using rotor-flux oriented control principles (Fig. 2).

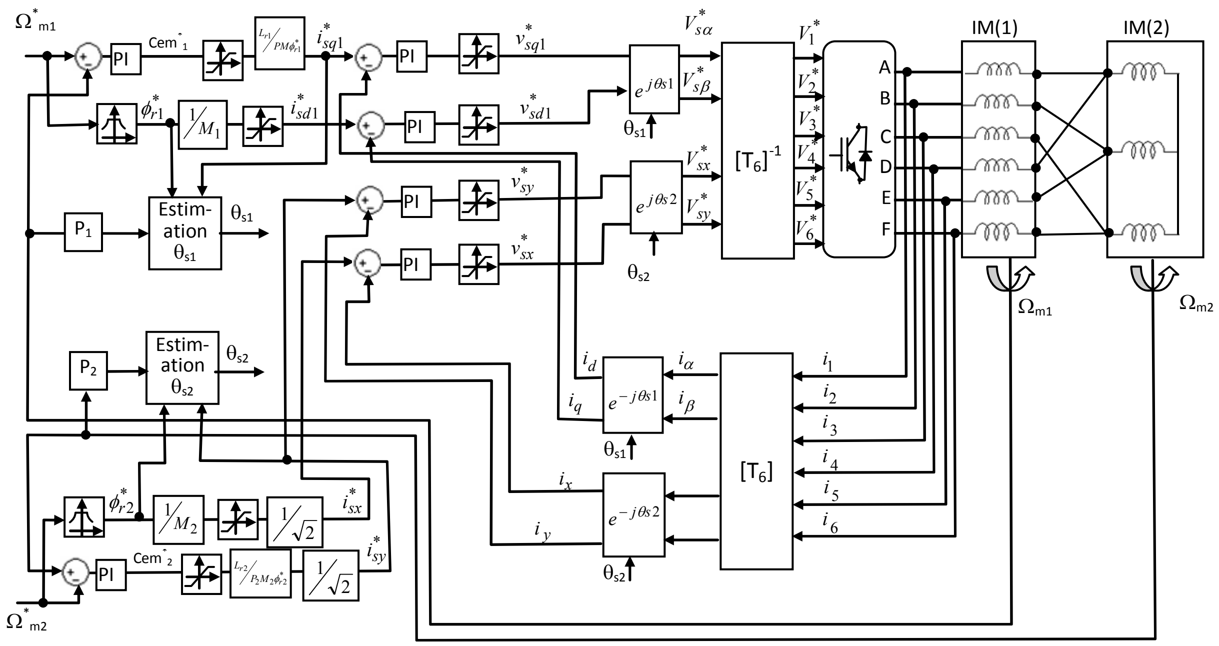

Fig. 2. Indirect rotor flux oriented controller for the two-motor drive.

Model Simplifies of the Multimachines System

If the plane (*d, q*) is perfectly directed, we suppose that the component $\varphi_{rq,k} = 0$. This simplifies the model of the MSCS as follows:

$$(15)\qquad \begin{cases} \frac{d\varphi_{r\alpha,k}}{dt} = A \frac{M_k}{T_{r,k}} X - \frac{1}{T_{r,k}} \varphi_{r\alpha,k} \\ \frac{d\varphi_{r\beta,k}}{dt} = \frac{M_k}{T_{r,k}} Y - (\omega_{s,k} - P_k \Omega_{m,k}) \varphi_{r\alpha,k} \\ \frac{d\Omega_{m,k}}{dt} = A \frac{p_k M_k}{J_k L_{r,k}} \varphi_{r\alpha,k} . i_{s\alpha,k} - \frac{1}{J_k} T_{l,k} \end{cases}$$

Where:

Six-phase machine (k=1):

$$(16.a)\qquad \begin{cases} X = i_{s\alpha 1} \\ Y = i_{s\beta 1} \\ A = 1 \end{cases}$$

Three -phase machine (k=2) :

$$(16.b)\qquad \begin{cases} X = i_{sx} \\ Y = i_{sy} \\ A = \sqrt{2} \end{cases}$$

By introducing the angular speeds of sliding, the following equation is obtained:

$$(17)\qquad \frac{d\theta_{sl,k}}{dt}=\omega_{sl,k}=(\omega_{s,k}-p_k\Omega_{m,k})=\frac{M_k}{T_{r,k}}\cdot\frac{i^s}{\varphi_{rd,k}}$$

Where : $\theta_{s,k}=\theta_{sl,k}+\theta_{m,k}$

With: $i^s=\begin{cases} i_{s\beta} & for \quad k=1 \\ \sqrt{2}.i_{sy} & for \quad k=2 \end{cases}$

With this condition, the fluxes and torques for MSCS are:

$$(18)\qquad \begin{cases} \varphi_{r\alpha1}=\dfrac{M_1}{1+T_{r1}S}i_{s\alpha1} \\ T_{em1}=\dfrac{p_1M_1}{L_{r1}}\varphi_{r\alpha1}i_{s\beta1} \end{cases}$$

$$(19)\qquad \begin{cases} \varphi_{r\alpha2}=\sqrt{2}\dfrac{M_2}{1+T_{r2}S}i_{sx} \\ T_{em2}=\sqrt{2}\dfrac{p_2M_2}{L_{r2}}\varphi_{r\alpha2}i_{sy} \end{cases}$$

Where: S is the time derivative operator.
According to 18, six-phase machine's (I.M1) flux/torque are controllable by inverter ($i_{s\alpha}$, $i_{s\beta}$) axis current components, while flux and torque of the three-phase machine (I.M2) can be controlled using inverter (i_{sx} , i_{sy}) current components.

Behavior Model Control Principals

In this paper, k is the machine number (k=1: first machine, k=2: second machine) and the "mod" subscript is used to define model values (e.g. $Y_{mod,k}$, the model output). $P_k(S)$ and $M_k(S)$ are transfer functions. The Behavior Model Control [17, 22] aims to define a complementary control function (value) to that defined by the main controller in order to enhance the control algorithm performance. It requires at least two controllers, a model and the process to be controlled (Fig. 3).

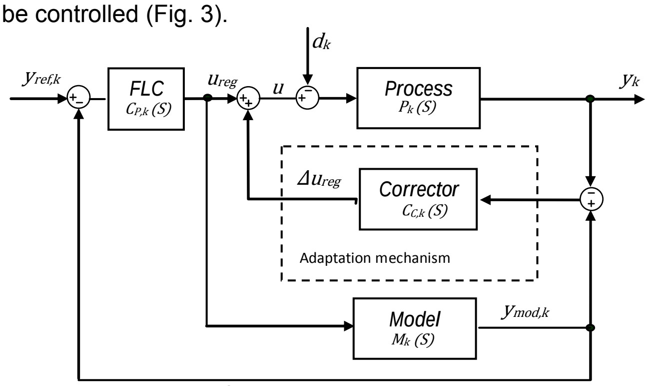

Fig. 3. Behavior Model Control (Functional Block)

According to the set-point (reference) $Y_{ref,k}$, the main controller $C_{p,k}(S)$ delivers an output value $u_{reg,k}$. The issued control value $u_{reg,k}$ from the main controller is applied to a predefined model $M_k(S)$ called the Behavior Model(BM). The BM define a model output value $Y_{mod,k}$. The second controller called "The Behavior Controller or the Adaptation Corrector" uses the gap between the process output value Y_k and the model output value $Y_{mod,k}$, to define the complementary control function (value) $\Delta u_{reg,k}$.

By cancelling the error (Y_k - $Y_{mod,k}$), the process behaviour becomes similar to that of the predefined model, hence the term Behavior Model Control.

The complementary control function (value) $\Delta u_{reg,k}$ is added to the output value $u_{reg,k}$ and then applied to the process $P_k(S)$ as an input value.

As a result, this auxiliary control algorithm increases the robustness of the global control algorithm [17,18], it indirectly rejects various disturbances [19] , facilitates the synthesis of classical control [23] and allows the linearization of a non-linear processes through a linear model [24].

It should be noted that the main controller serves to eliminate the error between the set-point (reference) $Y_{ref,k}$ and output value. this later, can be the output of the model $Y_{mod,k}$ or that of the process Y_k. Therefore, one can define two structure of BMC: BMC based on the predefined model output, and BMC based on the process output.

Behaviour loop analysis

According to the scheme of Fig. 3, the following expressions, [24 ,26] are deduced:

$$(20)\qquad \begin{cases} Y_k(S)=P_k(S)\left[U_{reg,k}(S)+\Delta U_{reg,k}(S)-d_k\right] \\ \Delta U_{reg,k}(S)=C_{C,k}(S)\left[M_k(S)\cdot U_{reg,k}(S)-Y_k(S)\right] \end{cases}$$

Following the calculation, one arrive at the system that expresses the process output Y_k as well as that of the model $Y_{mod,k}$.

$$(21)\qquad \begin{cases} Y_k(S)=\dfrac{P_k(S)\left(1+M_k(S)\cdot C_{C,k}(S)\right)}{1+P_k(S)\cdot C_{C,k}(S)}\cdot U_{reg,k}(S)-\dfrac{P_k(S)}{1+P_k(S)\cdot C_{C,k}(S)}\cdot d_k \\ Y_{mod}(S)=M(S)\cdot U_{reg}(S) \end{cases}$$

Where d_k : represents the disturbance.
In order to simplify this system, the correction controller $C_{C,k}(S)$ must satisfy the following assumptions:

$$(22)\qquad \begin{cases} \left|M_k(S)\cdot C_{C,k}(S)\right|>>1 \\ \left|P_k(S)\cdot C_{C,k}(S)\right|>>1 \end{cases}$$

After simplification, the following is obtained:

$$(23)\qquad \begin{cases} Y_k(S)=M_k(S)\cdot U_{reg,k}(S)-\dfrac{1}{C_{C,k}(S)}d_k \\ Y_{mod,k}(S)=M_k(S)\cdot U_{reg,k}(S) \end{cases}$$

Which gives the following result:

$$(24)\qquad Y_k(S)=Y_{mod,k}(S)-\frac{1}{C_{C,k}(S)}d_k$$

The process output (Y_k) is the same as that of the model ($Y_{mod,k}$) at small disturbance. If this disturbance $\dfrac{d_k}{C_{C,k}(S)}$ is negligible compared to the process output Y_k, it follows perfectly the model output. This condition is written:

$$(25)\qquad \frac{d_k}{C_{C,k}(S)}<<M_k(S)\cdot U_{reg,k}(S)$$

Considering the return coming from the model output $Y_{mod,k}$, the system (21) becomes:

$$(26)\qquad \begin{cases} Y_k(S)=\dfrac{P_k(S)\left(1+M_k(S)\cdot C_{C,k}(S)\right)}{M_k(S)\cdot\left(1+P_k(S)\cdot C_{C,k}(S)\right)}\cdot Y_{mod,k}(S)-\dfrac{P_k(S)}{1+P_k(S)\cdot C_{C,k}(S)}\cdot d_k \\ Y_{mod,k}(S)=\dfrac{M_k(S)\cdot C_{P,k}(S)}{1+M_k(S)\cdot C_{P,k}(S)}\cdot Y_{ref,k}(S) \end{cases}$$

To simplify the transfer function $\dfrac{Y_k}{Y_{mod,k}}$, let us suppose the following hypothesis:

$$(27)\qquad \left|M_k(S)\cdot C_{C,k}(S)\right|>>1$$

This hypothesis could be the result of the process evolution, so that the second condition of the hypothesis (22) would not be verified. We obtain:

$$Y_k(S)=\frac{P_k(S)\cdot C_{C,k}(S)}{\left(1+P_k(S)\cdot C_{C,k}(S)\right)}\cdot Y_{\text{mod},k}(S)-\frac{P_k(S)}{1+P_k(S)\cdot C_{C,k}(S)}\cdot d_k \quad (28)$$

The transfer function $\frac{Y_{\text{mod},k}}{Y_{ref,k}}$ needs no simplifying assumption. It is the expression of a closed loop of the model $M_k(S)$ and the corrector $C_{p,k}(S)$. The synthesis of the corrector $C_p(S)$ is done according to the model $M_k(S)$. This closed loop not being affected by any external or internal disturbance obviously appears very robust.

These transfer functions ($\frac{Y_k}{Y_{\text{mod},k}}, \frac{Y_{\text{mod},k}}{Y_{ref,k}}$) put highlight control loop illustrated by the functional block diagram of Fig. 4:

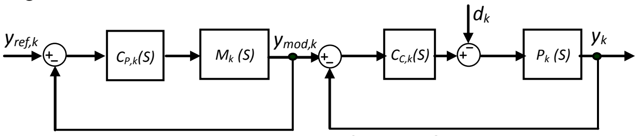

Figure 4. Functional block diagram of the BMC based on the model output.

On this equivalent diagram, it is clear that the $C_{C,k}(S)$ corrector does not only intervene on the perturbation, but also on the continuation of the reference (Input $Y_{mod,k}$).
As the two loops are in series, the slowest will impose the overall dynamics of the setting. In this case, we choose the first loop (M_k - $C_{p,k}$) slower than the second (P_k - $C_{C,k}$).
The transfer function $\frac{Y_k}{Y_{\text{mod},k}}$ can be further simplified by using the following expression:

$$\left|P_k(S)\cdot C_{C,k}(S)\right| >> 1 \quad (29)$$

we then obtain

$$Y_k(S)=Y_{\text{mod},k}(S)-\frac{d_k}{C_{C,k}(S)} \quad (30)$$

Equation 30 result is in agreement with that of equation 24. The choice of the model, for the speed loop, focused on the transfer function of two motors which links the mechanical speed to the electromagnetic torque (equation 31). For the current loop, the model is equivalent to the transfer function (32) in order define the parameters of the current controllers (equation 32)[27].

$$M_{\Omega,k}(S)=\frac{1}{J_k\cdot S+F_k} \quad (31)$$

F_k: viscous friction coefficient

$$M_{i,k}(S)=\frac{1}{\sigma_k\cdot L_{s,k}\cdot S+K_{t,k}} \quad (32)$$

Where: $\sigma_k=1-\frac{M_k^2}{L_{s,k}\cdot L_{r,k}}$; $K_t=R_{s,k}+\frac{M_k^2}{L_{r,k}\cdot T_{r,k}}$

Correction controller (used in the speed loop)

The correction controller used in the speed loop allows to cancel the error between the output speed of the machine Ω_k and that of the model $\Omega_{mod,k}$. So it is very convenient to use this error as well as its derivative as inputs to this corrector. By integrating the output of the latter, we obtain the correction signal $\Delta T_{em,k}$ which allows the two machines to have a similar behavior to that of the model.

The internal structure of the "Correction Controller" block FLCR is identical to that of an FLC (Main Controller), i.e it consists of three blocks: Fuzzification (F), Knowledge Base and the Inference (I) and defuzzification (D) (Fig. 5).
where : u represents the stator currents ($i_{ds,k}$ and $i_{qs,k}$) and speed Ω_k for the two machines.

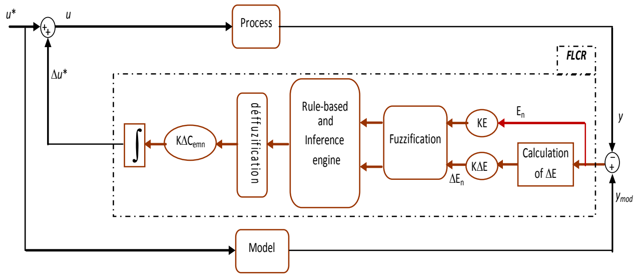

Figure 5. Structure of fuzzy correction controller FLCR.

As with the design process of the main controller FLC, each input is represented by seven fuzzy sets. This leads to a rule base of forty-nine (49) rules. The inference method used is that of Mamdani (Max-Min). While defuzzification is achieved by the Center of Area (COA) method is employed [22] ,[28 ,32].
The membership functions of input variables E and ΔE rspectively and output variable, which are with conventional triangular shapes. Each membership is divided into seven and three fuzzy sets for speed and stator currents respectively (Fig.6)

The following table 1 shows the rules used for the design of the correction controller (behavior) (used in the speed loop of two machines).

Table 1. Shows the rule base for controlling the speed

EΩ / ΔEΩ	NH	NM	NS	ZE	PS	PM	PH
NH	NH	NH	NH	NH	NM	NS	ZE
NM	NH	NH	NH	NM	NS	ZE	PS
NS	NH	NH	NM	NS	ZE	PS	PM
ZE	NH	NM	NS	ZE	PS	PM	PH
PS	NM	NS	ZE	PS	PM	PH	PH
PM	NS	ZE	PS	PM	PH	PH	PH
PH	ZE	PS	PM	PH	PH	PH	PH

The different sets are noted as follows:
NH: Negative High; PS: Positive low; ZE: near Zero;
NM: Negative Medium; PM: Positive Average;
NS: Negative low; PH: Positive High.

According to Fig. 2, we have two behavioural controllers for the two current loops for each machine in the system. The first for the current loop $i_{ds,k}$, the second for the current loop $i_{qs,k}$. Each of these controllers makes it possible to cancel the difference between the process output two machines and that of the model $M_{i,k}(S)$; in order to impose on each machine a similar behaviour to that of the model $M_{i,k}(S)$.

As for the speed loop correction controller, the input of the FLCR is the error $E_{C,k}$, as well as its derivative. This error is equal to $E_{c,k}=i_{sd,k}-i_{sd,\text{mod},k}$ for the FLCR of the current $i_{sd,k}$ and $E_{c,k}=i_{sq,k}-i_{sq,\text{mod},k}$ for the FLCR controller of the current $i_{sq,k}$.

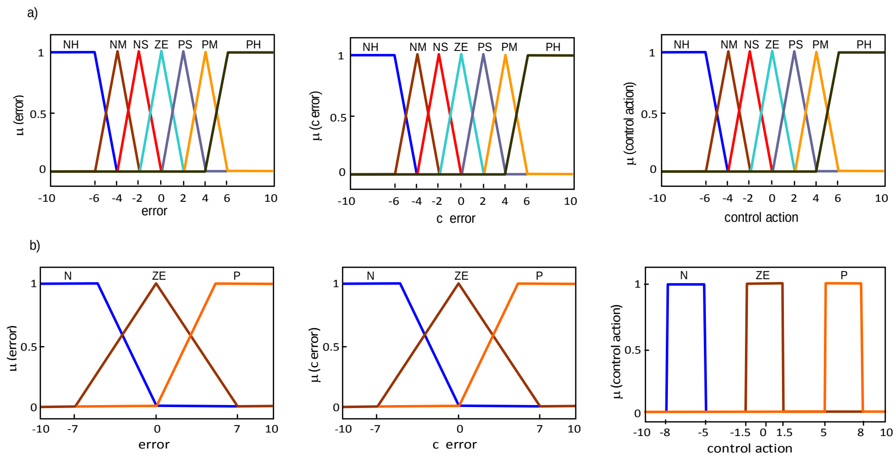

Figure. 6. Membership functions of input/output variables: a) speed, b) current

We will use three (03) sets for each controller input. This leads to nine (09) rules for each current ($i_{sd,k}$ and $i_{sq,k}$), controller , shown in Table 2. The inference is carried out by the Max-Min method, and the defuzzification by that of the gravity center.

Table2. Shows the rule base for controlling the currents

ΔE_i \ E_i	N	ZE	P
N	N	N	ZE
ZE	N	ZE	P
P	ZE	P	P

Speed and currents controllers tuning

The main controller has to impose the dynamics of the main loop (Fig. 4). So, it imposes the closed-loop response time, $t_{RM,k}$. The secondary controller has to impose the dynamics of the secondary loop, in order to obtain (24). So, it yields the secondary closed-loop response time $t_{RP,k}$. 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:

(34) $$\begin{cases} (t_{RM})_{I,k} < (t_{RM})_{\Omega,k} \\ (t_{RP})_{I,k} < (t_{RP})_{\Omega,k} \end{cases}$$

But there is another condition on the secondary controller tuning, the inequality (22). It means that the response time ($t_{RP,k}$) of the secondary closed loop must be smaller than the response time ($t_{RM,k}$) of the main closed-loop. This condition must be checked in the case of two overlap loops, one will thus write:

(35) $$(t_{RP})_{I,k} < (t_{RM})_{I,k} < (t_{RP})_{\Omega,k} < (t_{RM})_{\Omega,k}$$

Simulation Results

The simulation results are intended to simulate the application of fuzzy controllers on the speed control as well as the decoupled control of the two machines of the MSCS.

Case 1: Two motors turned in the same direction

Fig. 7 shows the operation of the two-machine drive system for many different speeds references with no load at starting up phase. At steady state condition, the two machines are loaded simultaneously or not by their nominal loads. At the beginning, the first machine is running at 20 rad/s; at t=0.5 s, it is accelerated to 50 rad/s, after that, its direction of rotation is reversed to -50 rad/s at t=2s and then stopped at t=3.5s. For the second machine the speed reference is set at 40 rad/s, 100 rad/s,-60 rad/s , and -40 rad/s at t=0 s, 1 s, 2.5 s, 3.5 s, respectively.It is clear too that the start of MAS-HP (I.M1) did not have an impact on the speed or on the electromagnetic torque of the MAS-TP (I.M2).

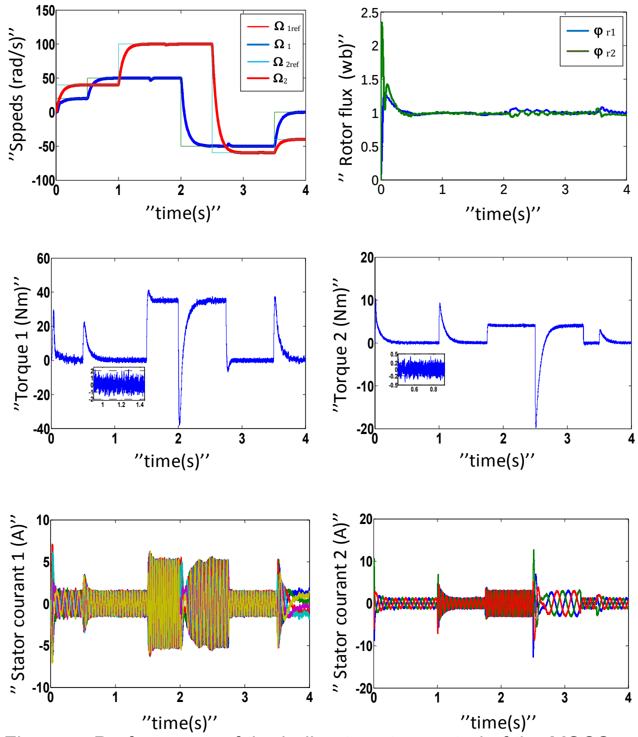

Figure 7. Performance of the indirect vector control of the MSCS using fuzzy adaptive regulators: operating at different reference speeds values.

Case 2: Two machines turned in opposite directions

The effect of the speed rotation reversion of one machine on the other machine performance is investigated in Fig. 8. In the stating phase, the first machine is rotating at +50 rad/s; the other is running at the opposite speed -100rad/s. After that Ω_1 is kept at standstill between 1.5 and 3 s, while Ω_2 is reversed from -100 to +100 rad/s at t=2 s, and it is returned to zero at t=3.5 s. At the subsequent test, the speed Ω_2 is held at zero, while Ω_1 is set at -50 rad/s at 3s. The results illustrated in Figs. 7,8 show once more that the control of the two machines is completely decoupled. Indeed, the speed of one machine and its electromagnetic torque remain completely undisturbed during the reversion of the other machine, indicating a complete decoupling of the control. So, there is hardly any evidence of torque disturbance of one machine during the reversal of the other one. As shown from Figs. 8, the starting and reversing transients of one machine do not have any tangible consequence on the operation of the second machine. The decoupled control is preserved and the characteristics of both machines are unaffected.

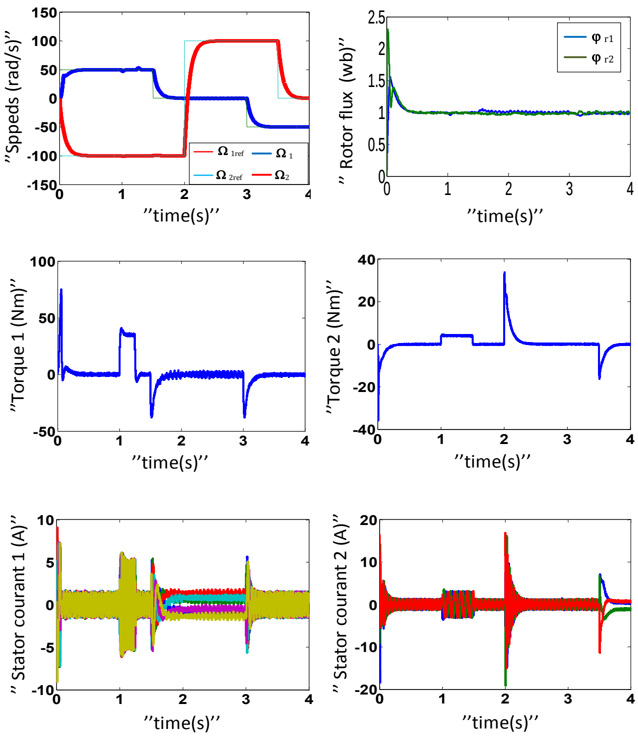

Figure 8. Dynamic responses of the MSCS using fuzzy adaptive regulators: when the two motors are operating in the opposite directions.

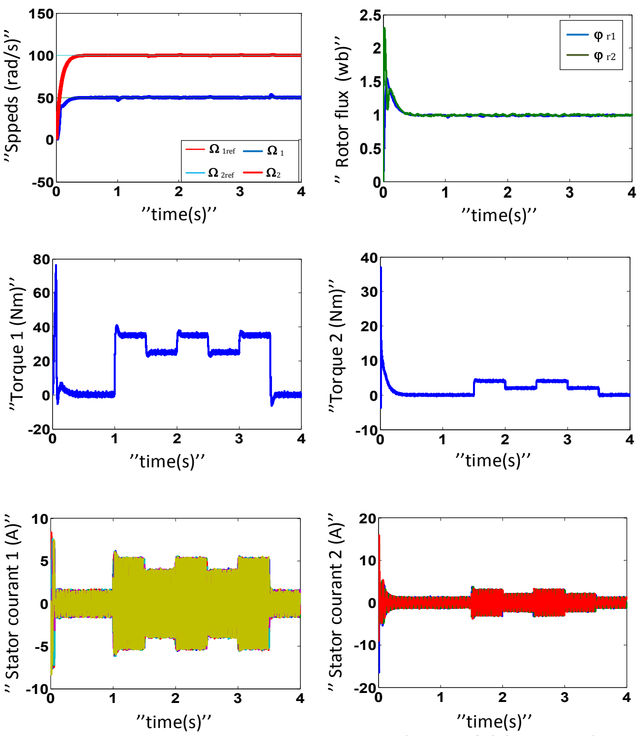

Figure 9. Dynamic responses controlled of the MSCS using fuzzy adaptive regulators and operating at different loading conditions.

Case 3: Operation at different loading conditions

Fig. 9 shows the performance of the two-machine drive controlled by BCM controllers under load torques variations condition. The reference of the first speed is set at 50 rad/s, while the speed reference of the second machine is set at 100 rad/s. A set of load torques from 39 Nm to 30 Nm and from 4 Nm to 2 Nm are applied on the two machines shafts in different times respectively.

It is clear from Fig. 9 that when one machine is either loaded or unloaded, the second machine performance is unaffected; which proves once again that both motors connected in series are totally decoupled. In case of BCM control, no variation whatsoever can be observed in the speeds responses of the both machines during these transients. It is worth noticing that there is no impact on the speed and electromagnetic torque of one machine when the speed or the load of the other machine in series -connected system changes. Thus, through proper phase transformation rules, the decoupled control of two machines connected in series can be achieved with a single supply from a seven-level six-phase voltage source inverter. In addition, measured and estimated speeds are in excellent agreement, with BCM as FLC and PI.

Robustness test

Figs. 10 represent the speed responses of I.M(1) machine driver controlled by BCM, FLC and conventional PI of the system under moment of inertia variations condition $J_1=J_{1nominal}$, $J_1=3J_{1n}$ and $J_1=5J_{1n}$ of the MAS-HP. The reference of the first speed is set at 50 rad/s to -50, while the speed reference of the second machine is set at 100 rad/s.

According to Fig. 10.c, with FLC and PI control, the speed path has changed and does not have the same response time for $J_1=5J_{1n}$. On the other hand, the BMC retains its trajectory better (Fig10.b,c). At the bottom of the Fig.10, it shows that with BMC, the process is less disturbed by an external disturbance than by compared to the conventional control (PI) and the FLC control. In addition, it is noted that the BMC provides means to better control the transient error due to external disturbance.

The second test consists of the robustness test of the system. An example of the robustness of the BMC compared with FLC and conventional PI controllers. We change the motor parameters (R_s,R_r and L_s) and without realizing anyadjustment in the controllers the speed regulation is tested in a motor control. The new six-pahse motor parameters are: R_r = 6Ω, R_s = 4.6Ω, L_m = 0.2H, L_r = 0.06H, L_s = 0.184H, J_1 = 0.12Kg.m2. Fig. 11 shows the speed of the six-phase machine controlled with PI, fuzzy logic controllers and BMC, a load torque of 39N.m is applied at t = 1s to the new machine, without readjusting the controllers. The Fig. 11 (a) shows the response of the six-phase machine controlled with PI controllers. The performance of the system becomes wrong when the load changes at t = 1s, the system becomes instable. But, with the BM, FLC the speed regulation is correct as shown in Fig. 11 (b) and (c).

Table 3. The values parameters for the Six-phase induction motor.

Rated power:	P_n = 5.5kw
Nominal currant:	I_n = 6A
Stator resistance:	R_s = 2.3 Ω
Rotor resistance:	R_r = 3 Ω
Stator inductance:	L_s = 0.203H
Rotor inductance:	L_r = 0.203H
Mutual inductance:	L_m =0.2H
Rated phase stator voltage:	V_n = 220v
Pole pair number:	P = 1
Rotor speed:	N = 1000 tr/min
Friction coefficient:	Kf = 0.006 Nms/rad
Moment of inertia:	J_1 = 0.06Kg.m²

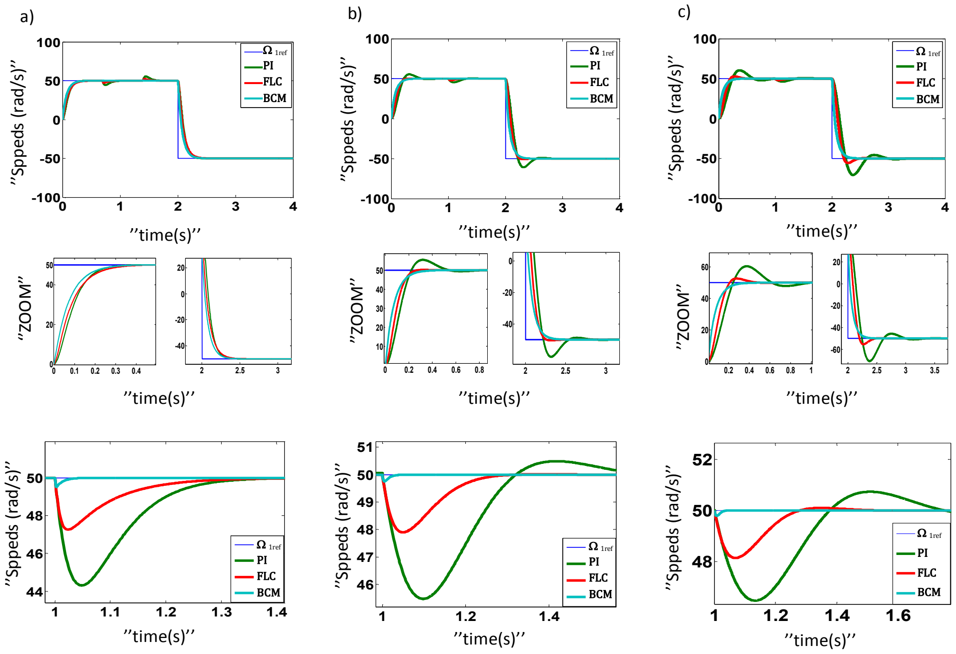

Figure 10. Speed response of I.M1 - rotor inertia variation and application of a load torque (39 N.m) at t = 1s:
a) J1=J1n; b) J1=3J1n; c) J1=5J1n

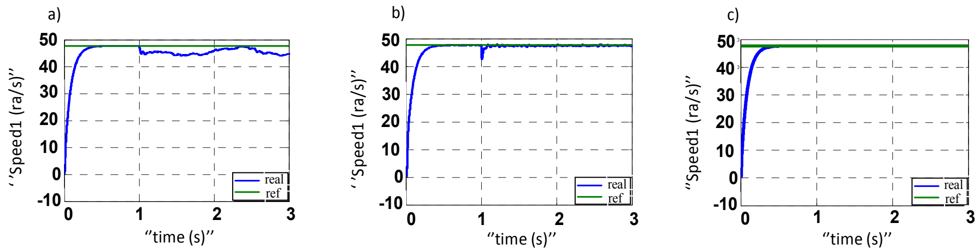

Figure 11.Test of robustness of the BMC ,fuzzy controller and PI with applied load torque 39Nm at *t* = 1s.
a) PI controllers, b) FLC , c) BMC.

Table 4.The values parameters for the Three-phase induction motor.

Puissance nominal:	P_{n2} = 1 kw
Stator resistance:	R_{s2} = 4.67 Ω
Rotor resistance:	R_{r2} = 8 Ω
Stator inductance:	L_{s2} = 0.374H
Rotor inductance:	L_{r2} = 0.374H
Mutual inductance:	L_{m2} =0.2433H
Rated phase stator voltage:	V_{n2} = 220v
Pole pair number:	P_2 = 3
Rotor speed:	N_2 = 2830 tr/min
Friction coefficient:	Kf $_2$ = 0.001 N.m.s/rd
Moment of inertia:	J_2 = 0.023Kg.m²

Conclusion

This paper reported a new form of adaptive control based on fuzzy logic, named as the fuzzy behavior model control. The advantages of this control method lie in the fact that it exploits the difference between the response of the system and that of an imposed model in order to force the system to follow a defined behavior, by generating a control signal perfectly adapted to the functioning of the machine.

The results obtained, show that the superiority of the BMC controller for a variation of 400% of rotor variation inertia ($J_1 = 5J_{1n}$), which is really impressive. Despite this wide range of variation, excellent system performance is still achieved. Hence, the BMC controller is more robust and outperforms as the fuzzy controller (FLC) on the considered aspects and as powerful as a conventional controls (PI) because he has the capacity to keep these trajectories in spite of the external perturbations and the parameters variations. Therefore, the BMC is the most efficient.For a further work in this subject, we propose: faults diagnostic of the system.

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