Comparative Study Between the Field Oriented Control and Backstepping Control of Open-End Winding Five-Phase Induction Motor under Open Phase Fault Conditions

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***Abstract*—Open-End Winding Five-Phase Induction Motor (OEW-FPIM) are employed in applications where the reduction in the total power per phase and the highest level of overall system reliability is required, where a self-starting and operation under open phase(s) stand as the most salient features. This paper presents a comparative study of the most powerful control strategies for motor: field-oriented control and backstepping control. The awaited goal is to evaluate the proposed approaches which gives the best performance under open phase fault conditions. The performance of the two control approaches verified and compared by simulation. Provided tests analyze steady and transient states, including an open-phase fault condition. Obtained results prove the interest of the proposed approaches, which ensures the open-phase fault-tolerant capability of OEW-FPIM.**

***Keywords*—five-phase induction motor, field-oriented control , Backstepping control, open-phase fault, open-end winding.**

I. INTRODUCTION

Multi-phase induction motors have a lot of advantages in comparison to conventional three-phase induction motors and have attracted more and more interest for their use in applications where high reliability. This concerns especially the applications that, due to safety issues, cannot be suddenly shut down, such as in electric vehicles [1] and aerospace [2], electric ship propulsion or wind power generation systems [3]. Current interest in these machines is mostly the consequence of their advantages, when compared to a power equivalent three-phase machines such as: reducing the amplitude and increasing the frequency of torque pulsation, reducing the current per phase without increasing the voltage per phase, and providing higher flux density and more output torque [4]-[6]. Another distinguished advantage is improved reliability and continuous system operation even in the fault condition [7]. Among the available multiphase machines that have been proposed in the literature [8]-[10], the five-phase induction machines (FPIM) are now recognized as practical selections in safety-critical high power applications, because the FPIM is able to increase at least 10% torque compared to its three-phase counterpart of the same active volume [11] and they present a serious option that offers better machine performance under fault conditions [12]. On the other hand, multilevel inverters are used in single-sided supply mode for adjustable speed drive application. However, conventional multilevel inverters have some drawbacks such as dc link voltage balancing problems and use of high rating capacitors [13]. Alternative solution for obtaining multilevel output is by supplying an open-end winding (OEW) machine using dual inverter. The star point is opened and supply is given from both sides of the motor through inverters, which combines the advantages of high fault tolerance of multi-phase machines and high power quality of multilevel inverters and induction machines. This topology offers several advantages when compared with the traditional single-sided two-level inverter supplied motor [14, 15] such as higher redundancy, no clamping diodes or capacitors are needed and the possibility to reduction of common mode voltage. However, an unexpected fault may occur due to improper operation. According to statistics from [16, 17], among all types of faults, an open-phase or broken-phase of machines, short-circuiting of stator windings and damage to switching devices in the five-phase bridge inverter are the most common. In FPIM with one phase open, when open-circuit fault occurs, the unbalanced currents generate, which increases the induced torque ripple [18] and distorts the air gap flux distribution causing eventual breakdown of the motor [19].

Many control techniques concerning multiphase motor have been proposed in the literature, the scalar control method (V/f) is simpler and still widely used in industrial fields. In the literature, the control of multiphase induction motor using V/f is addressed in healthy [25] and faulty operation [26]. The Rotor-Field Oriented Control (RFOC) method, based on PI controller loop is the most widespread control technique in the multiphase case. This technique has been applied in FPIM in healthy [27] and faulty operation [28]. Another widely used control technique is the Direct torque control [29, 30]. Other well-known non-linear controller like Backstepping, has not been so deeply extended to the multiphase case, where only healthy operation of FPIM is considered [31]. The main objective of this paper was the synthesis of two control strategies for FPIM under phase open conditions via Backstepping control and flux oriented control. The basic goal of the FOC is decoupling the stator flux and electromagnetic torque by orientation of the rotor flux. RFOC gives good steady state as well as transient response. Moreover, the backstepping control offers numerous advantages such as, fast tracking, and acceptable torque ripple in addition to high dynamics and excellent stability proprieties. The basic idea of the backstepping control is to render the equivalent closed-loop system to stable subsystems with order equal one which are put in cascade by the Lyapunov theory, which gives them the qualities of strength and global asymptotic stability. The rest of this paper is organized as follows. Section II details the FPIM-OEW model is presented. The FOC and backstepping strategy are shown in the Section III and IV. The simulation results and a comparative study between the performances of the proposed control techniques, before and after the fault event are presented in the Section V. The conclusions are finally summarized in Section VI.

II. Open-End Winding Topology

A. *Modeling of five phase induction motor*

The multiphase system modeling is usually simplified using the vector space decomposition (VSD) approach [32]. For d-q-x-y reference, four independent variables divided in two orthogonal planes called d-q and x-y. However first components d-q are responsible for torque production, while x-y remaining components do not generate electrical torque and generates losses in the system. An additional axis called z is also defined related to the zero-sequence component of the system. The dynamic equation of five phase induction motor in the d-q-x-y reference frame after transformation can be written in the following form:

$$\begin{cases} \dfrac{di_{sd}}{dt} = \alpha_1 + \dfrac{1}{\sigma L_s} u_{sd} \\ \dfrac{di_{sq}}{dt} = \alpha_2 + \dfrac{1}{\sigma L_s} u_{sq} \\ \dfrac{di_{sx}}{dt} = \dfrac{R_s}{Ll_s} i_{sx} + \dfrac{1}{Ll_s} u_{sx} \\ \dfrac{di_{sy}}{dt} = \dfrac{R_s}{Ll_s} i_{sy} + \dfrac{1}{Ll_s} u_{sy} \end{cases} \tag{1}$$

$$\begin{cases} \dfrac{d\Psi_r}{dt} = \dfrac{L_m}{T_r} i_{sd} - \dfrac{\Psi_r}{T_r} \\ \dfrac{d\omega}{dt} = \dfrac{n_p^2 \Psi_r}{JL_r} - \dfrac{n_p}{J} T_L - \dfrac{F}{J}\omega \end{cases} \tag{2}$$

With:

$$\alpha_1 = -\frac{L_m^2 R_r + R_s L_r^2}{\sigma L_r^2 L_s} i_{ds} + \omega i_{sq} + \frac{L_m \Psi_r}{\sigma L_r T_r} + \frac{L_m i_{sq}^2}{T_r \Psi_r}$$

$$\alpha_2 = -\frac{L_m^2 R_r + R_s L_r^2}{\sigma L_r^2 L_s} i_{sq} - \omega\left(i_{sq} + \frac{L_m \Psi_r}{\sigma L_r^2}\right) - \frac{i_{sq} i_{sd} L_m}{T_r \Psi_r}$$

Where: R_s, R_r, L_m, L_s and L_r are the stator resistance, the rotor resistance, the magnetizing inductance, the stator cyclic inductance and the rotor cyclic inductance, respectively. $Ll_s = L_s - L_m$ is the stator leakage reactance, $Ll_r = L_r - L_m$ is the rotor leakage reactance, n_p is the number of pole pairs, T_r is the rotor time constant.

B. *Dual inverter open-end winding*

The proposed dual two-level inverter feeding the open-end stator winding of the five-phase induction motor is shown in Fig. 1, where only one DC source is used. The dual inverter are identified with indices 1 and 2. Inverter legs are denoted with capital letters A, B, C, D, E with their suffix 1 and 2 and O is assumed virtual neutral point. The phase voltages of the two inverters can be given as follows:

$$\begin{cases} u_{sa} = u_{a1O} - u_{a2O} = u_{A1O} - u_{A2O} \\ u_{sb} = u_{b1O} - u_{b2O} = u_{B1O} - u_{B2O} \\ u_{sc} = u_{c1O} - u_{c2O} = u_{C1O} - u_{C2O} \\ u_{sd} = u_{d1O} - u_{d2O} = u_{D1O} - u_{D2O} \\ u_{se} = u_{e1O} - u_{e2O} = u_{E1O} - u_{E2O} \end{cases} \tag{3}$$

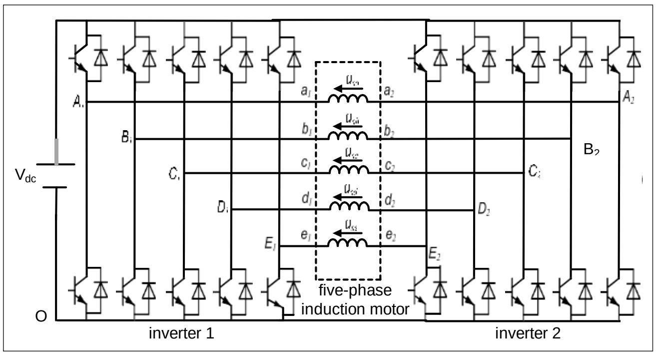

Fig. 1. The principle topology of a FPIM-OEW

Where: u_{A10}, u_{B10}, u_{C10}, u_{D10}, u_{E10} and u_{A20}, u_{B20}, u_{C20}, u_{D20}, u_{E20} are the pole voltages of inverter1 and inverter2, respectively.

III. Rotor Flux Oriented Control

The aim of the indirect rotor flux oriented control is to obtain the control of open-end winding five-phase induction motor similar to dc motor with separate excitation where there is a natural decoupling between the flux and the electromagnetic torque. The IFOC consists in making $\Psi_{rq} = 0$ while the rotor direct flux Ψ_{rd} converges to the reference Ψ_r^*:

$$\begin{cases} \Psi_{rd} = \Psi_r \\ \Psi_{rq} = 0 \end{cases} \tag{4}$$

The decoupled torque and the rotor flux can be controlled independently, by controlling the d and q components of stator current. In this work proposes the indirect rotor flux oriented control strategy. It is the most used because it eliminates the influence of stator and rotor leakage reactances and yield better results than the methods based on the orientation of the stator flux. The equation of the machine in reference related to the rotating frame can be simply described by the following equations:

$$\begin{cases} V_{sd} = \left[R_s + s\sigma(Ll_s + L_m)\right] i_{sd} + sL_m / (Ll_r + L_m)\Psi_r \\ \quad - \omega_s \sigma(Ll_r + L_m) i_{sq} \\ V_{sd} = \left[R_s + s\sigma(Ll_s + L_m)\right] i_{sq} + \omega_s L_m / (Ll_r + L_m)\Psi_r \\ \quad + \omega_s \sigma(Ll_r + L_m) i_{sd} \\ V_{sx} = \left(R_s + sLl_s\right) i_{xs} \\ V_{sy} = \left(R_s + sLl_s\right) i_{ys} \end{cases} \tag{5}$$

The expressions of the electromagnetic torque and slip speed are:

$$\begin{cases} T_e = \dfrac{n_p L_m}{(Ll_r + L_m)} i_{qs} \Psi_r \\ \omega_{sl} = \dfrac{n_p L_m}{\Psi_r T_r} i_{qs} \end{cases} \tag{6}$$

In this type of control the angle θ used for direct and inverse transformation is calculated using the following expression:

$$\theta = \int (\omega_s + \frac{L_m}{T_r} \Psi_r i_{qs}) dt \tag{7}$$

Where: ω_{sl} is the slip relation, ω_s is the synchronous speed, T_r is the rotor time constant.

IV. BACKSTEPPING CONTROL

The basic idea of the Backstepping design is the use of the so-called virtual control to systematically decompose a complex nonlinear control design problem into simpler, smaller ones. Backstepping design is divided into various design steps [33]. In each step we essentially deal with an easier, single-input-single-output design problem and each step will provide a reference for the next step. Stability and performance of our system will be studied using Lyapunov theory [34]. The synthesis of this control can be achieved in two steps.

1) Step 1: Computation of the reference stator currents: Since the rotor speed and the rotor flux module are our control variables, We define the errors e_ω and e_Ψ representing the error between the actual speed and the reference speed and the error between the rotor flux module and its reference, the tracking errors are defined by:

$$\begin{cases} e_\omega = \omega^* - \omega \\ e_\Psi = \Psi_r^* - \Psi_r \\ \dot{e}_\omega = \dot{\omega}^* - \dot{\hat{\omega}} \\ \dot{e}_\Psi = \dot{\Psi}_r^* - \dot{\Psi}_r \end{cases} \tag{8}$$

Accounting for Eq. (2), Eq. (8) turns to be:

$$\begin{cases} \dot{e}_\omega = \dot{\omega}^* - \dfrac{n_p^2 L_m}{J L_r} \hat{\Psi}_r i_{sq} + \dfrac{F}{J}\omega + \dfrac{T_L}{J} n_p \\ \dot{e}_\Psi = \dot{\Psi}_r^* + \dfrac{\Psi_r}{T_r} - \dfrac{L_m}{T_r} i_{sd} \end{cases} \tag{9}$$

The first Lyapunov function associated with the speed and rotor flux errors can be defined as following:

$$\begin{aligned} V_1 &= 0.5\left(e_\Psi^2 + e_\omega^2\right) \\ \dot{V}_1 &= \left(-K_\Psi . e_\Psi^2 - K_\omega . e_\omega^2\right) \end{aligned} \tag{10}$$

The pursuit of goals are achieved by choosing the references of the current components representing the stabilizing functions as:

$$\begin{cases} i_{sd}^* = \dfrac{T_r}{L_m}\left(K_\Psi e_\Psi + \dot{\Psi}^* + \dfrac{\Psi_r}{T_r}\right) \\ i_{sq}^* = \dfrac{L_r}{n_p^2 L_m \Psi_r}\left(\dot{\omega}^* + K_\omega e_\omega + \dfrac{F\omega}{J} + \dfrac{T_L n_p}{J}\right) \\ i_{sx}^* = 0 \\ i_{sy}^* = 0 \end{cases} \tag{11}$$

Where K_Ψ and K_ω are positive constants that determine the closed loop dynamics. To satisfy equation (11), we must choose the dynamic errors as the following form:

$$\begin{cases} \dot{e}_\Psi = -K_\Psi . e_\Psi^2 \\ \dot{e}_\omega = -K_\omega . e_\omega^2 \end{cases} \tag{12}$$

Thus, the tracking objectives will be satisfied if we choose. So, the control i_{sd}^* and i_{sq}^* are asymptotically stabilizing and are considered as references for the next step.

2) Step 2: Computation of the reference stator voltages: To calculate the control law u_{sd}^*, u_{sq}^*, u_{sx}^* and u_{sy}^* of the complete system, we will define in the second step the new errors $e_{i_{sd}}$, $e_{i_{sq}}$, $e_{i_{sx}}$ and $e_{i_{sy}}$, which represent respectively the errors signals between stator the current and and their references as follows:

$$\begin{cases} e_{i_{sd}} = i_{sd}^* - i_{sd} \\ e_{i_{sq}} = i_{sq}^* - i_{sq} \\ e_{i_{sx}} = i_{sx}^* - i_{sx} \\ e_{i_{sy}} = i_{sy}^* - i_{sy} \end{cases} \tag{13}$$

The derivative of stator current errors gives:

$$\begin{cases} \dot{e}_{i_{sd}} = \dot{i}_{sd}^* - \dot{i}_{sd} \\ \dot{e}_{i_{sq}} = \dot{i}_{sq}^* - \dot{i}_{sq} \\ \dot{e}_{i_{sx}} = \dot{i}_{sx}^* - \dot{i}_{sx} \\ \dot{e}_{i_{sy}} = \dot{i}_{sy}^* - \dot{i}_{sy} \end{cases} \tag{14}$$

Accounting for Eq. (8), Eq. (14) turns to be:

$$\begin{cases} \dot{e}_{i_{sd}} = \dfrac{di_{sd}^*}{dt} - \alpha_1 - \dfrac{1}{\sigma L_r} u_{sd} \\ \dot{e}_{i_{sq}} = \dfrac{di_{sq}^*}{dt} - \alpha_2 - \dfrac{1}{\sigma L_r} u_{sq} \\ \dot{e}_{i_{sx}} = \dfrac{di_{sx}^*}{dt} - \dfrac{R_s}{Ll_s} i_{sx} - \dfrac{1}{Ll_s} u_{sx} \\ \dot{e}_{i_{sy}} = \dfrac{di_{sy}^*}{dt} - \dfrac{R_s}{Ll_s} i_{sy} - \dfrac{1}{Ll_s} u_{sy} \end{cases} \tag{15}$$

The final Lyapunov function V_2 based on the errors of speed, rotor flux and of the stator currents, which is given by the following expression:

$$V_2 = \frac{1}{2}\left(e_\Psi^2 + e_\omega^2 + e_{i_{sd}}^2 + e_{i_{sq}}^2 + e_{i_{sx}}^2 + e_{i_{sy}}^2\right) \tag{16}$$
$$\dot{V}_2 = \left(e_\Psi \dot{e}_\Psi + e_\omega \dot{e}_\omega + e_{i_{sd}} \dot{e}_{i_{sd}} + e_{i_{sq}} \dot{e}_{i_{sq}} + e_{i_{sz1}} \dot{e}_{i_{sz1}} + e_{i_{sz2}} \dot{e}_{i_{sz2}}\right)$$

By replacing Eq (15) in Eq (16), one can obtains:

$$\begin{aligned}
\dot{V}_2 = &\left(-K_\Psi e_\Psi^2 - K_\omega e_\omega^2 - K_{i_{sd}} e_{i_{sd}}^2 - K_{i_{sq}} e_{i_{sq}}^2 - K_{i_{sx}} e_{i_{sx}}^2 - K_{i_{sy}} e_{i_{sy}}^2\right) \\
&+ e_{i_{sd}}\left(\frac{L_m}{T_r} e_\Psi + \frac{di_{sd}^*}{dt} - \alpha_1 - \frac{u_{ds}}{\sigma L_r} + e_{i_{sd}} K_{i_{sd}}\right) \\
&+ e_{i_{sq}}\left(\eta \Psi_r e_\omega + \frac{di_{sq}^*}{dt} - \alpha_2 - \frac{u_{qs}}{\sigma L_r} + e_{i_{sq}} K_{i_{sq}}\right) \\
&+ e_{i_{sx}}\left(\frac{di_{sx}^*}{dt} - \frac{R_s}{Ll_s} i_{sx} - \frac{1}{Ll_s} u_{sx} + e_{i_{sx}} K_{i_{sx}}\right) \\
&+ e_{i_{sy}}\left(\frac{di_{sy}^*}{dt} - \frac{R_s}{Ll_s} i_{sy} - \frac{1}{Ll_s} u_{sy} + e_{i_{sy}} K_{i_{sy}}\right)
\end{aligned} \tag{17}$$

To ensure that the derivative of the V_2 must be negative:

$$\dot{V}_2 = \begin{cases}
\left(-K_\Psi e_\Psi^2 - K_\omega e_\omega^2 - K_{i_{sd}} e_{i_{sd}}^2 - K_{i_{sq}} e_{i_{sq}}^2 - K_{i_{sx}} e_{i_{sx}}^2 - K_{i_{sy}} e_{i_{sy}}^2\right) = 0 \\
\left(\frac{L_m}{T_r} e_\Psi + \frac{di_{sd}^*}{dt} - \alpha_1 - \frac{u_{sd}}{\sigma L_r} + e_{i_{sd}} K_{i_{sd}}\right) = 0 \\
\left(\eta \Psi_r e_\omega + \frac{di_{sq}^*}{dt} - \alpha_2 - \frac{u_{sq}}{\sigma L_r} + e_{i_{sq}} K_{i_{sq}}\right) = 0 \\
\left(\frac{di_{sx}^*}{dt} - \frac{R_s}{Ll_s} i_{sx} - \frac{1}{Ll_s} u_{sx} + e_{i_{sx}} K_{i_{sx}}\right) = 0 \\
\left(\frac{di_{sy}^*}{dt} - \frac{R_s}{Ll_s} i_{sy} - \frac{1}{Ll_s} u_{sy} + e_{i_{sy}} K_{i_{sy}}\right) = 0
\end{cases} \tag{18}$$

The control voltages based on currents errors are given:

$$\begin{cases}
u_{sd}^* = \sigma L_s \left(K_{i_{sd}} e_{i_{sd}} - \alpha_1 + \frac{L_m}{T_r} e_\Psi + di_{sd}^*/dt\right) \\
u_{sq}^* = \sigma L_s \left(K_{i_{sq}} e_{i_{sq}} - \alpha_2 + \eta \Psi_r e_\omega + di_{sq}^*/dt\right) \\
u_{sx}^* = \sigma (Ll_s + L_m)\left(K_{i_{sx}} e_{i_{sx}} + di_{sx}^*/dt - \frac{R_s}{Ll_s} i_{sx}\right) \\
u_{sy}^* = \sigma (Ll_s + L_m)\left(K_{i_{sy}} e_{i_{sy}} + di_{sy}^*/dt - \frac{R_s}{Ll_s} i_{sy}\right)
\end{cases} \tag{19}$$

With $K_{i_{sd}}$, $K_{i_{sq}}$, $K_{i_{sx}}$ and $K_{i_{sy}}$ are positive parameters selected to guarantee a faster dynamic of the stator currents, the rotor flux and the rotor speed, the equation can be expressed as:

$$\begin{cases}
\dot{e}_{i_{sd}} = -K_{i_{sd}} e_{i_{sd}} - \frac{L_m}{T_r} e_\Psi \\
\dot{e}_{i_{sq}} = -K_{i_{sq}} e_{i_{sq}} - \eta \Psi_r e_\omega \\
\dot{e}_{i_{sx}} = -K_{i_{sx}} e_{i_{sx}} \\
\dot{e}_{i_{sy}} = -K_{i_{sy}} e_{i_{sy}}
\end{cases} \tag{20}$$

This equation guarantees the asymptotically convergence of the speed, currents and flux errors. The block diagram of the Backstepping control of FPIM-OEW is shown in Fig.2.

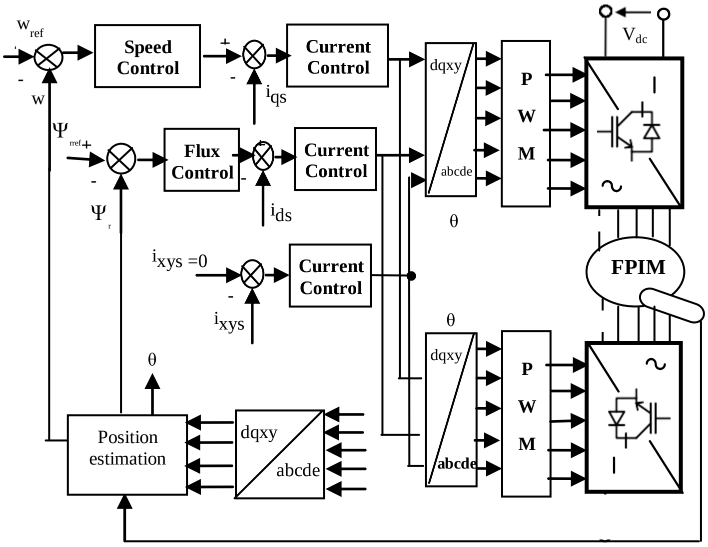

Fig. 2. The block diagram of the Backstepping control

V. SIMULATIONS RESULTS AND DISCUSSIONS

Computer simulations have been carried out to validate the effectiveness of the indirect rotor flux oriented control compared to backstepping control. The Matlab/Simulink software package has been used for this purpose to test the motor under different operation conditions, including the open-phase fault. The performed tests consider that the open-phase fault condition occurs in phase ‘a1a2’ and the fault occurrence is emulated by opening a power relay connected in series with the faulty phase, and with the following electrical and mechanical specifications of the used FPIM-OEW: R_s = 1.2Ohms Rr = 1.8Ohms, L_{ls} = 5.4 mH, L_{lr} = 5.4 mH, L_m = 150 mH, 4-pole, J=0.07 kgm^2 and F= 00.1. It is important to mention that during the following simulation tests the rotor flux reference was set to 1 Wb. The resulting waveforms for backstepping and RFOC are shown in Figures 4and 5, side by side, speed, torque, d-q stator current and rotor flux are presented.

The behavior of the proposed indirect rotor flux oriented control during healthy and open-phase operation of the FPIM-OEW is first analyzed assuming a no load as starting up condition and the load is changed to 2 N.m at t=3s and a step in the speed reference from 0 to 100 rad/s that is applied at t=2 s. The fault occurs at time t=4 s. Fig. 4 shows the obtained results: the phase voltage (a1a2), speed, torque and d-q phase current and rotor flux responses before and after the fault appearance, respectively. The waveforms in zoom 1 and zoom 2 are the magnified speed and torque, while the waveforms in zoom 3 are the magnified rotor flux, both in healthy and fault conditions. It can be observed that the indirect rotor flux oriented control scheme provides an accurate tracking of the reference speed and it can be observed that 2.18 Nm’s average torque with 0.55 ripple is produced under single phase open circuit fault. It must be mentioned that the maximum stator current reference in the q-axis, related to the maximum achievable electrical torque. In addition, the response of the flux is good because the real rotor flux tracks the reference value adequately, decoupling between flux and torque is maintained.

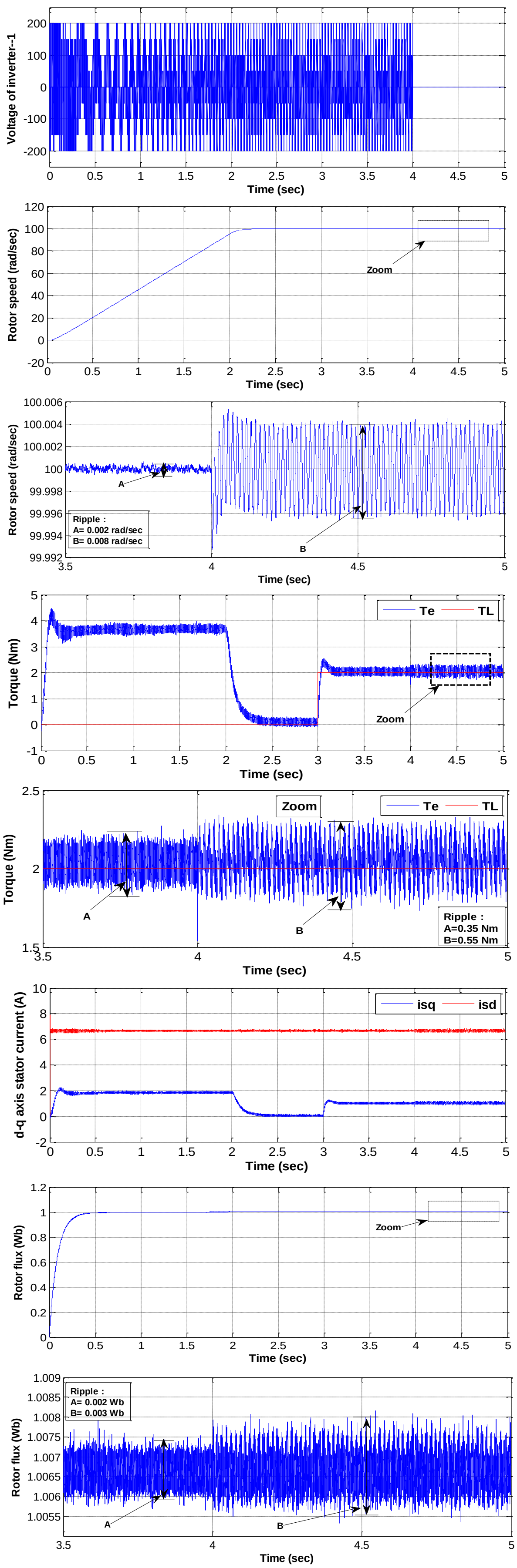

Fig. 3. The simulation results for backstepping control of FPIM-OEW, both in healthy and fault conditions

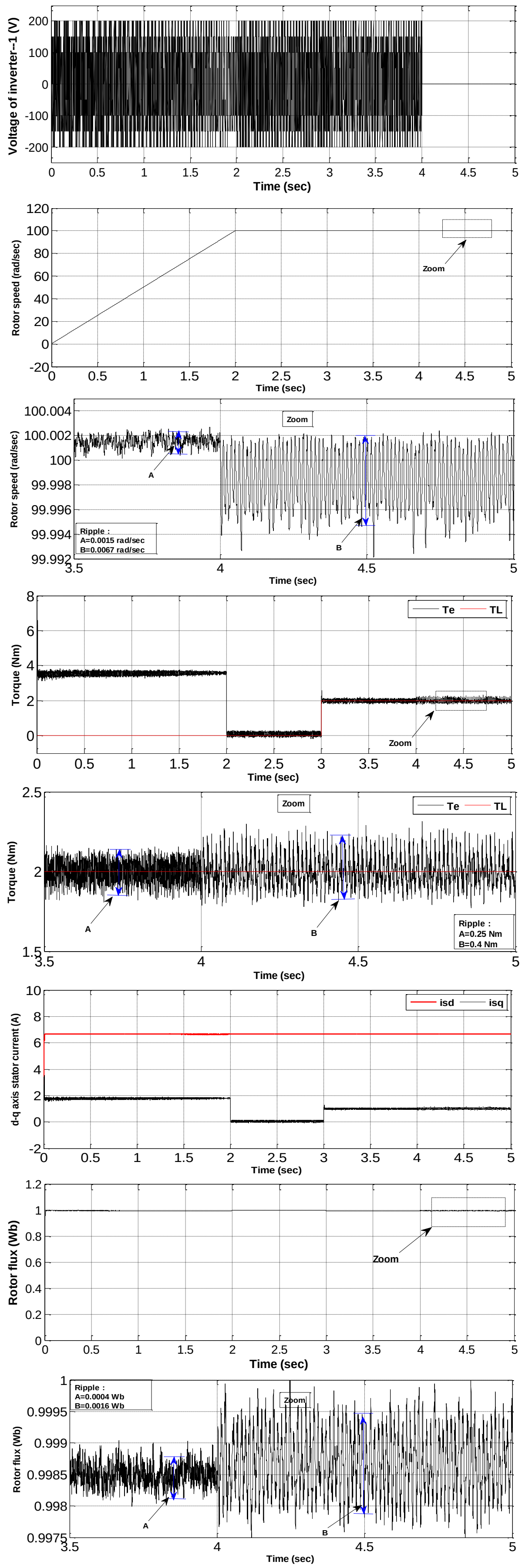

Fig. 4. The simulation results for DFOC of FPIM-OEW, both in healthy and fault conditions

Next, the performance of the backstepping control when a fault appears is studied. Fig. 5 shows simulation results of the proposed backstepping control strategy for FPIM-OEW during healthy and open-phase fault at the reference speed of 100 rad/s at 2 s. The open phase fault is generated in phase “a1a2” at t = 4 s. Fig. 5 shows the phase voltage (a1a2), speed, torque and d-q phase current and rotor flux responses both in healthy and fault conditions, respectively. The waveforms in zoom 4 and zoom 5 are the magnified speed and torque, while the waveforms in zoom 6 are the magnified rotor flux, both in healthy and fault conditions. It is found that the fault occurrence does not affect the rotor speed and the speed reference is accurately followed without any overshoot and steady-state error. Quantitatively, ω is 100.001rad/s with 0.0015 rad/s speed ripple in healthy condition, but w is 99.999 rad/s with 0.0067 rad/s speed ripple in fault condition, respectively. It can be observed that the average torque is 2.25 Nm ripple under normal conditions, whereas 2.4Nm average torque is produced under open phase fault conditions. Then, from d-q phase current waveform. It can be seen that the transition from the healthy to the fault situation is almost unnoticeable. So , the backstepping strategy is effective during the loss of one phase, offering a similar performance as in healthy condition. This is an interesting characteristic in favor of the backstepping strategy in applications where the highest reliability must be preserved and the drive must be operated in safety conditions until a corrective maintenance is carried out.

When comparing Figs. 4 and 5, these results show that using the backstepping strategy a better torque response can be achieved in terms of settling time and maximum overshoot. The different dynamic behavior is due to the presence of PI regulators in DFOC scheme, which delay the torque response. In addition, the simulation results are almost similar except in the DFOC technique, the ripple of torque in steady state is reduced remarkably compared with DFOC technique (Fig. 4).

VI. BACKSTEPPING CONTROL

In this paper, the principle and a several characteristics of RFOC and Backstepping Control schemes are studied by simulation in order to determinate the main advantages and drawbacks of each control and to make a comparison between them for FPIM-OEW under open phase fault conditions. The aim of the paper was to give a fair comparison between RFOC and backstepping techniques, to allow the users to identify the more suitable solution for any application that requires control under open phase fault. In order to achieve this goal, a control strategy using field oriented first and backstepping secondly is shown. Backstepping control offers better response compared to other nonlinear control of motor. The backstepping control offers high performance in both healthy and open phase conditions. Indeed, backstepping control gives high dynamic and good tracking.

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