# ROBUST DTC BASED ON ADAPTIVE FUZZY CONTROL OF DOUBLE STAR SYNCHRONOUS MACHINE DRIVE WITH FIXED SWITCHING FREQUENCY

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The double star synchronous machine (DSSM) is widely used for high power traction drives. It possesses several advantages over the conventional three phase machine. To reduce the torque ripple the DSSM are supplied with source voltage inverter (VSI). The model of the system DSSM-VSI is high order, multivariable and nonlinear. Further, big harmonic currents are generated. The aim of this paper is to develop a new direct torque adaptive fuzzy logic control in order to control DSSM and minimize the harmonics currents. Simulations results are given to show the effectiveness of our approach.

K e y w o r d s: double star synchronous machine (DSSM), direct torque control (DTC), tree level inverter, fixed switching frequency adaptive fuzzy control, harmonic currents

### **1 INTRODUCTION**

Synchronous machines are attractive for high powers. When they are supplied by thyristors current source inverter (CSI) torque oscillations of high amplitude at six times the supply frequency appear [1, 2]. This causes excess mechanical vibration and poor system performance. The use of two star windings shifted by  $30^{\circ}$  supplied by two thee phase voltage source inverters (VSIs) lowers considerably the torque ripple [3, 4]. However, the model of the system DSSM-VSIs in Park frame is high order, multivariable and non linear

Direct torque control is a very simple method to control AC machines. It shows a very high quality torque control without the need of tuning current controllers. The principle of conventional DTC consists to select stator voltage vectors, according to the differences between the stator flux linkage and torque and their references, to keep stator flux and torque within the limits of two hysteresis bands. However, the disadvantages of conventional DTC include notable torque, flux and current pulsations and non-constant switching frequency operation [5, 6].

The advantages allotted to traditional techniques DTC (dynamic, robustness, facilitated implementation ... ) nevertheless are counterbalanced by the use of a sampled hysteresis comparator; by principle, the comparator led functioning at variable frequency which increases the risks of mechanical excitation or acoustic resonances, and in addition, sampling at finished frequency results by an exceeding pseudo-random of the hysteresis band. These two factors contribute to make the contents harmonic of the various output signals not easily foreseeable.

Alternatively, space vector modulation (SVM) modulator is incorporated with direct torque control for induction motor drives and permanent magnet synchronous motors as shown in [7-9] to provide a constant inverter switching frequency. Noting that the torque ripple and noise issue for this SVM-based DTC approach are significantly improved for invoking zero inverter switching state within every switching period of inverter control.

Recently, there has been an increasing interest in adaptive fuzzy control (AFC) for non-linear dynamic systems. This interest has been motivated by the demands for high control performance in situations in which an accurate model of the controlled plant is not easily available, or when the plant undergoes unforeseen parameter variations. Fuzzy systems have the potential to play an important role in adaptive control, mainly thanks to their universal function approximation property and their amenability to (linguistic) interpretation of the inputoutput relationships.

An adaptive fuzzy controller is expected to comply with the following design requirements: (i) stability of the closed-loop system (all signals in the closed-loop must be bounded), (ii) asymptotic convergence of the tracking error to zero or to a neighbourhood of zero, (iii) robustness with respect to the approximation error, unavoidably introduced by any approximator with a finite number of parameters. The approximation error is usually treated as a disturbance acting on the system by means of standard modifications: an additional non-linear damping term, usually in a sliding mode framework [10–12].

The aim of this article is to develop an adaptive fuzzy control of DTC double star synchronous machine drive with fixed switching frequency. The fuzzy systems are used to approximate the model of controlled system. In order to compensate for the effects of the unavoidable reconstruction errors, we introduce a sliding term into the control law. The approximation theory and the Lyapunov method are used together to construct, in the first stage, the fuzzy adaptive control law and to establish, in the second stage, the convergence of the tracking error

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and the boundedness of the adaptive parameters and all plant signals. Furthermore, the space vector modulation technique is applied to inverter control in the proposed DTC-based double star synchronous motor drive system, thereby dramatically reducing the torque ripple, current harmonic and to impose the switching frequency.

The remainder of this paper is structured as follows: in Section 2 the used fuzzy logic system is briefly described and Section 3 is devoted to the problem statement. The machine model is developed in Section 4. In Section 5, the direct torque control of DSSM of double star synchronous machine drive is presented. The DTC adaptive fuzzy control is developed in Section 6. Finally, the conclusions are presented in Section 7.

## 2 DESCRIPTION OF THE USED FUZZY LOGIC SYSTEM

The fuzzy logic system performs a mapping from  $U_1 \times \cdots \times U_n \subset \mathbb{R}^n$  to  $\mathbb{R}$  where each  $U_i \subset \mathbb{R}$ ,  $i = 1, 2, \ldots, n$ . For the proposed fuzzy logic system, we use the implication and the reasoning method suggested by Takagi and Sugeno [13]. Consequently, the fuzzy IF-THEN rules are expressed as

$$R_k$$
: if  $x_1$  is  $A_1^{l_1}$  and  $\dots x_n$  is  $A_n^{l_n}$  then  $z_k = a^k$  (1)

where  $x = [x_1 \ x_2 \ \dots \ x_n]^\top \in \Re^n$  and  $z_k \in \Re$  are respectively the input of the fuzzy logic system and the consequent of the  $k^{\text{th}}$  rule. Here, the label  $A_i^{li}$  associated to input  $x_i$   $(i = 1, \dots, n)$  is a fuzzy set in  $U_i$  where the indices li takes a value in  $\{1, \dots, m_i\}$  and  $m_i$  is the number of fuzzy sets characterising the input  $x_i$ . The coefficient  $a^k$  ( for  $k = 1, 2, \dots, M$ ) is a constant coefficient of consequent part for the  $k^{\text{th}}$  fuzzy rule. The number of rules M is defined by the cartesian product as:  $M = m_1 \otimes m_2 \otimes \cdots \otimes m_n$ .

In this paper, we employ the product operation for fuzzy implication and T-norm. The definition of the product operation is the same as in [14]. Besides, the singleton fuzzifier and weighted average deffuzzification is used. The overall output value z(x) is

$$z(x) = \frac{\sum_{k=1}^{M} \alpha_k a^k}{\sum_{k=1}^{M} \alpha_k}$$
(2)

Where  $\alpha_k$  denotes the firing strength of the  $R_k$  rule, which is evaluated by using product inference and implication as

$$\alpha_k = \prod_{i=1}^n \mu A_i^{li}(x_i); \text{ with } l_i \in \{1, \dots, m_i\}$$
 (3)

where  $\mu A_i^{li}(x_i)$  is the membership function of  $x_i$  associated to fuzzy set  $A_i^{li}$ .

In (2), we fix the  $\mu A_i^{li}(x_i)$ s and  $a^k$ s form the adjustable parameters, z(x) can be rewritten as

$$z(x) = w^{\top}(x) \theta \tag{4}$$

where  $\theta$  is the parameter vector given by

$$\theta = \left[a^1 \, a^2 \dots a^M\right]^{\top} \tag{5}$$

and  $w^{\top}(x)$  is a regressive vector defined as

$$w^{\top}(x) = \begin{bmatrix} \frac{\alpha_1}{\sum_{k=1}^M \alpha_k} & \frac{\alpha_2}{\sum_{k=1}^M \alpha_k} & \cdots & \frac{\alpha_M}{\sum_{k=1}^M \alpha_k} \end{bmatrix}$$
(6)

In the sequel, the fuzzy logic system for multi-input single output system is represented by the mathematical model (4).

#### **3 PROBLEM FORMULATION**

The control of the DSSM supplied by voltage source inverters (VSIs) is very difficult since the model of the machine is high order, multivariable and non linear. Figure 1 presents the decoupled control scheme for DSSM supplied by two three level inverters. The decoupled control bloc is based on DTC control strategy witch the stator flux linkage and the torque would be decoupled.

#### **4 MACHINE MODEL**

In developing the equations which describe the behaviour of DSSM, the following assumptions are made: – The MMF in air-gap has a sinusoidal repartition.

- The saturation of the iron in machine is neglected.

The stator voltage equation for six-phase can be written as

$$\mathbf{v}_{s} = \mathbf{R}_{s} \mathbf{i}_{s} + \frac{\mathrm{d}}{\mathrm{d}t} \left( \mathbf{L}_{ss} \mathbf{i}_{s} + \mathbf{M}_{sr} i_{f} \right)$$
(7)

$$\mathbf{v}_{s} = \begin{bmatrix} v_{a1} & v_{a2} & v_{b1} & v_{b2} & v_{c1} & v_{c2} \end{bmatrix}^{+}, \qquad (8)$$

$$\mathbf{i}_{s} = \begin{bmatrix} i_{a1} & i_{a2} & i_{b1} & i_{b2} & i_{c1} & i_{c2} \end{bmatrix}^{\top}$$
. (9)

The original six dimensional system of the machine can be decomposed into three orthogonal subspaces  $(\alpha, \beta)$ ,  $(Z_1, Z_2)$  and  $(Z_3, Z_4)$  [15, 18]

$$\begin{bmatrix} F_{\alpha} & F_{\beta} & F_{z1} & F_{z2} & F_{z3} & F_{z4} \end{bmatrix}^{\top} = \mathbf{T}_{s} \mathbf{F}_{s}$$
(10)

with F: can be voltage, current or flux.

$$\mathbf{\Gamma}_{s} = \frac{1}{\sqrt{3}} \times \\
\begin{pmatrix}
\cos 0 & \cos \gamma & \cos \frac{2\pi}{3} & \cos \left(\frac{2\pi}{3} + \gamma\right) & \cos \frac{4\pi}{3} & \cos \left(\frac{4\pi}{3} + \gamma\right) \\
\sin 0 & \sin \gamma & \sin \frac{2\pi}{3} & \sin \left(\frac{2\pi}{3} + \gamma\right) & \sin \frac{4\pi}{3} & \sin \left(\frac{4\pi}{3} + \gamma\right) \\
\cos 0 & \cos(\pi - \gamma) & \cos \frac{4\pi}{3} & \cos(\frac{\pi}{3} - \gamma) & \cos \frac{2\pi}{3} & \cos \left(\frac{5\pi}{3} - \gamma\right) \\
\sin 0 & \sin(\pi - \gamma) & \sin \frac{4\pi}{3} & \sin \left(\frac{\pi}{3} - \gamma\right) & \sin \frac{2\pi}{3} & \sin \left(\frac{5\pi}{3} - \gamma\right) \\
1 & 0 & 1 & 0 & 1 & 0 \\
0 & 1 & 0 & 1 & 0 & 1
\end{pmatrix}$$
(11)



Fig. 1. Decoupled control scheme for DSSM



Fig. 2. Equivalent circuit of the DSSM model in: (a)  $-\alpha - \beta$  subspace, (b)  $-Z_1, Z_2$  and  $Z_3, Z_4$  subspaces

#### 4.1 Machine model in $\alpha - \beta$ subspace

The stator voltage equation is

$$\begin{pmatrix} v_{\alpha} \\ v_{\beta} \end{pmatrix} = R_s \begin{pmatrix} i_{\alpha} \\ i_{\beta} \end{pmatrix} + \frac{\mathrm{d}}{\mathrm{d}t} \begin{pmatrix} l_{fs} + 3M_{ss} & 0 \\ 0 & l_{fs} + 3M_{ss} \end{pmatrix} \begin{pmatrix} i_{\alpha} \\ i_{\beta} \end{pmatrix}$$
$$+ \sqrt{3}M_{sf} \frac{\mathrm{d}}{\mathrm{d}t} \begin{pmatrix} \cos(\theta) \\ \sin(\theta) \end{pmatrix} i_f$$
$$+ M_{sfm} \frac{\mathrm{d}}{\mathrm{d}t} \begin{pmatrix} 3\cos(2\theta) & 3\sin(2\theta) \\ 3\sin(2\theta) & -3\cos(2\theta) \end{pmatrix} \begin{pmatrix} i_{\alpha} \\ i_{\beta} \end{pmatrix}.$$
(12)

An equivalent circuit of DSSM in  $\alpha$ - $\beta$  subspace is shown in Fig. 2a.

### **4.2** Machine model in $Z_1, Z_2$ subspace

The stator voltage equation is

$$\begin{pmatrix} v_{z1} \\ v_{z2} \end{pmatrix} = R_s \begin{pmatrix} i_{z1} \\ i_{z2} \end{pmatrix} + \frac{\mathrm{d}}{\mathrm{d}t} \begin{pmatrix} l_{fs} & 0 \\ 0 & l_{fs} \end{pmatrix} \begin{pmatrix} i_{z1} \\ i_{z2} \end{pmatrix}. \quad (13)$$

The currents harmonics  $i_{z1}$  and  $i_{z2}$  must be as low as possible to reduce the extra losses in the DSSM. These currents are only limited by stator resistance and leakage inductance. An equivalent circuit of DSSM in  $Z_1, Z_2$ subspace is shown in Fig. 2b.

# **4.3** Machine model in $Z_3, Z_4$ subspace

The stator voltage equation is

$$\begin{pmatrix} v_{z3} \\ v_{z4} \end{pmatrix} = R_s \begin{pmatrix} i_{z3} \\ i_{z4} \end{pmatrix} + \frac{\mathrm{d}}{\mathrm{d}t} \begin{pmatrix} l_{fs} & 0 \\ 0 & l_{fs} \end{pmatrix} \begin{pmatrix} i_{z3} \\ i_{z4} \end{pmatrix}.$$
(14)

The currents  $i_{z3}$  and  $i_{z4}$  are equal to zero because the two three-phase windings are connected with isolated neutrals.

# 4.4 Machine model in rotor reference frame d-q

To express the stator and rotor equations in the same stationary reference frame, the following rotation transformation is appropriate

$$\mathbf{P} = \begin{pmatrix} \cos\theta & -\sin\theta\\ \sin\theta & \cos\theta \end{pmatrix}.$$
 (15)

With this transformation, the components of the  $\alpha$ - $\beta$  plane can be expressed in the d-q plane as

$$\begin{pmatrix} v_d \\ v_q \end{pmatrix} = \begin{pmatrix} R_s + pL_d & -\omega L_d \\ \omega L_d & R_s + pL_d \end{pmatrix} \begin{pmatrix} i_d \\ i_q \end{pmatrix} + M_d \omega \begin{pmatrix} 0 \\ 1 \end{pmatrix} i_f.$$
(16)

The electromagnetic torque of DSSM is expressed as

$$F_e = P(\phi_d i_q - \phi_q i_d).$$
(17)

4.5 Machine model in stator flux reference frame x-y

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Fig. 3. The stator and rotor flux linkages in different reference  $$\rm frames$$ 

In order to control directly and independently the flux and the torque the model of the DSSM is expressed in the stator flux reference frame by using the following transformation, witch transforms variable in the rotor flux reference frame (d-q) to the stator flux reference frame x-y (Fig. 3)

$$\begin{pmatrix} \cos\delta & \sin\delta\\ -\sin\delta & \cos\delta \end{pmatrix}.$$
 (18)

The stator flux linkage and electromagnetic torque equations in x-y reference frame are as follows

$$v_x = R_s i_x - w_s L_d i_y + L_d \frac{\mathrm{d}i_x}{\mathrm{d}t} + w \phi_r \sin \delta ,$$
  

$$v_y = R_s i_y + w_s L_d i_x + L_d \frac{\mathrm{d}i_y}{\mathrm{d}t} + w \phi_r \cos \delta$$
(19)



Fig. 4. Direct torque control scheme for DSSM



Fig. 5. Voltage vectors of three-level

 $w_s$ : rotating speed of stator flux linkage

 $\delta~$  : Angle between rotor and stator flux linkage.

The fluxes and the torque are given by

 $\phi_x = \phi_s = L_d i_x + \phi_r \cos \delta \,, \quad \phi_y = 0 = L_d i_y - \phi_r \sin \delta \,,$ 

$$T_e = P\phi_s i_y \tag{20}$$

From (13) and (14) the model for the DSSM can be written like:

$$\frac{d\phi_s}{dt} = f_1 + b_1 v_x , \ \frac{dT_e}{dt} = f_2 + b_2 v_y , \ \frac{d\Omega}{dt} = f_3 + b_3 T_e \ (21)$$

$$f_{1} = -\frac{R_{s}}{L_{d}}(\phi_{s} - \phi_{r}\cos\delta),$$

$$f_{2} = -\frac{P\phi_{s}}{L_{d}}(R_{s}i_{y} + w_{s}\phi_{s} + (w_{s} - w)\phi_{r}\cos\delta) - Pi_{y}\dot{\phi}_{s},$$

$$f_{3} = -\frac{1}{j}(T_{r} + f_{c}\Omega),$$

$$b_{1} = 1, \qquad b_{2} = \frac{P\phi_{s}}{L_{d}}, \qquad b_{3} = \frac{1}{j}.$$
(22)



Fig. 6. The chosen inverter voltage vectors projected on  $\alpha - \beta$  plane

# **5 DIRECT TORQUE CONTROL OF DSSM**

The basic principle of the DTC is to select proper voltage vectors using a pre-defined switching table. The selection is based on the hysteresis control of the stator flux linkage and the torque [16] [17]. The principle of DTC control is based on the space vector notation. The space vectors of the voltage, current and flux of the DSSM are defined as follows

$$\overline{v}_{s} = \sqrt{\frac{2}{3}} \Big( v_{a1} + v_{b1} e^{j\frac{2\pi}{3}} + v_{c1} e^{j\frac{4\pi}{3}} + v_{a2} e^{j\gamma} + v_{b2} e^{j(\frac{2\pi}{3}+\gamma)} + v_{c2} e^{j(\frac{4\pi}{3}+\gamma)} \Big),$$

$$\overline{i}_{s} = \sqrt{\frac{2}{3}} \Big( i_{a1} + i_{b1} e^{j\frac{2\pi}{3}} + i_{c1} e^{j\frac{4\pi}{3}} + i_{a2} e^{j\gamma} + i_{b2} e^{j(\frac{2\pi}{3}+\gamma)} + i_{c2} e^{j(\frac{4\pi}{3}+\gamma)} \Big),$$

$$\overline{\phi}_{s} = \sqrt{\frac{2}{3}} \Big( \phi_{a1} + \phi_{b1} e^{j\frac{2\pi}{3}} + \phi_{c1} e^{j\frac{4\pi}{3}} + \phi_{a2} e^{j\gamma} + \phi_{b2} e^{j(\frac{2\pi}{3}+\gamma)} + \phi_{c2} e^{j(\frac{4\pi}{3}+\gamma)} \Big).$$
(23)

Neglecting the stator resistance, the relation between the stator flux linkage vector and the voltage vector of DSSM



Fig. 7. Direct torque control scheme for DSSM when  $\Omega < \Omega_{nom}/2$ 

**Table 1.** The switching states for inverters when  $\Omega \ge \Omega_{nom}/2$ 

| _      |    |            |                        |            |            |            |            |            |            |            |               |               |               |                         |            |            |            |            |            |            |            |            |               |               |               |
|--------|----|------------|------------------------|------------|------------|------------|------------|------------|------------|------------|---------------|---------------|---------------|-------------------------|------------|------------|------------|------------|------------|------------|------------|------------|---------------|---------------|---------------|
|        |    |            | (a) the first inverter |            |            |            |            |            |            |            |               |               |               | (b) the second inverter |            |            |            |            |            |            |            |            |               |               |               |
| $\Phi$ | au | $\Theta_1$ | $\Theta_2$             | $\Theta_3$ | $\Theta_4$ | $\Theta_5$ | $\Theta_6$ | $\Theta_7$ | $\Theta_8$ | $\Theta_9$ | $\Theta_{10}$ | $\Theta_{11}$ | $\Theta_{12}$ | $\Theta_1$              | $\Theta_2$ | $\Theta_3$ | $\Theta_4$ | $\Theta_5$ | $\Theta_6$ | $\Theta_7$ | $\Theta_8$ | $\Theta_9$ | $\Theta_{10}$ | $\Theta_{11}$ | $\Theta_{12}$ |
|        | 1  | $V_{3L}$   | $V_{3M}$               | $V_{4L}$   | $V_{4M}$   | $V_{5L}$   | $V_{5M}$   | $V_{6L}$   | $V_{6M}$   | $V_{1L}$   | $V_{1M}$      | $V_{2L}$      | $V_{2M}$      | $V_{1M}$                | $V_{2L}$   | $V_{2M}$   | $V_{3L}$   | $V_{3M}$   | $V_{4L}$   | $V_{4M}$   | $V_{5L}$   | $V_{5M}$   | $V_{6L}$      | $V_{6M}$      | $V_{1L}$      |
| 1      | 0  | $V_{5s}$   | $V_{6s}$               | $V_{6s}$   | $V_{1s}$   | $V_{1s}$   | $V_{2s}$   | $V_{2s}$   | $V_{3s}$   | $V_{3s}$   | $V_{4s}$      | $V_{4s}$      | $V_{5s}$      | $V_{6s}$                | $V_{6s}$   | $V_{1s}$   | $V_{1s}$   | $V_{2s}$   | $V_{2s}$   | $V_{3s}$   | $V_{3s}$   | $V_{4s}$   | $V_{4s}$      | $V_{5s}$      | $V_{5s}$      |
|        | -1 | $V_{5L}$   | $V_{5M}$               | $V_{6L}$   | $V_{6M}$   | $V_{1L}$   | $V_{1M}$   | $V_{2L}$   | $V_{2M}$   | $V_{3L}$   | $V_{3M}$      | $V_{4L}$      | $V_{4M}$      | $V_{5M}$                | $V_{6L}$   | $V_{6M}$   | $V_{1L}$   | $V_{1M}$   | $V_{2L}$   | $V_{2M}$   | $V_{3L}$   | $V_{3M}$   | $V_{4L}$      | $V_{4M}$      | $V_{5L}$      |
|        | 1  | $V_{2L}$   | $V_{2M}$               | $V_{3L}$   | $V_{3M}$   | $V_{4L}$   | $V_{4M}$   | $V_{5L}$   | $V_{5M}$   | $V_{6L}$   | $V_{6M}$      | $V_{1L}$      | $V_{1M}$      | $V_{2M}$                | $V_{3L}$   | $V_{3M}$   | $V_{4L}$   | $V_{4M}$   | $V_{5L}$   | $V_{5M}$   | $V_{6L}$   | $V_{6M}$   | $V_{1L}$      | $V_{1M}$      | $V_{2L}$      |
| 0      | 0  | $V_{1s}$   | $V_{1s}$               | $V_{2s}$   | $V_{2s}$   | $V_{3s}$   | $V_{3s}$   | $V_{4s}$   | $V_{4s}$   | $V_{5s}$   | $V_{5s}$      | $V_{6s}$      | $V_{6s}$      | $V_{4s}$                | $V_{5s}$   | $V_{5s}$   | $V_{6s}$   | $V_{6s}$   | $V_{1s}$   | $V_{1s}$   | $V_{2s}$   | $V_{2s}$   | $V_{3s}$      | $V_{3s}$      | $V_{4s}$      |
|        | -1 | $V_{6L}$   | $V_{6M}$               | $V_{1L}$   | $V_{1M}$   | $V_{2L}$   | $V_{2M}$   | $V_{3L}$   | $V_{3M}$   | $V_{4L}$   | $V_{4M}$      | $V_{5L}$      | $V_{5M}$      | $V_{4M}$                | $V_{5L}$   | $V_{5M}$   | $V_{6L}$   | $V_{6M}$   | $V_{1L}$   | $V_{1M}$   | $V_{2L}$   | $V_{2M}$   | $V_{3L}$      | $V_{3M}$      | $V_{4L}$      |

can be approximated by

$$\overline{\phi}_s(T_s) \approx \overline{v}_s T_s + \overline{\phi}_s(0) \,. \tag{24}$$

Replacing the voltage vector (17) in (18) the flux linkage vector can be written as

$$\overline{\phi}_s(T_s) \approx \overline{v}_{s1}T_s + \overline{v}_{s2}T_s + \overline{\phi}_s(0) \tag{25}$$

$$\overline{v}_{s1} = \sqrt{\frac{2}{3}} \Big( v_{a1} + v_{b1} e^{j\frac{2\pi}{3}} + v_{c1} e^{j\frac{4\pi}{3}} \Big),$$

$$\overline{v}_{s2} = \sqrt{\frac{2}{3}} \Big( v_{a2} e^{j\gamma} + v_{b2} e^{j\left(\frac{2\pi}{3} + \gamma\right)} + v_{c2} e^{j\left(\frac{4\pi}{3} + \gamma\right)} \Big).$$
(26)

It can be seen from equation (25) that the stator flux linkage can be controlled by application of the output voltages vectors of each inverter independently. Figure 4 shows the proposed DTC scheme for DSSM.

The stator flux linkage and the torque are estimated with

$$\begin{aligned} \phi_{\alpha}(t) &= \int_{0}^{t} (v_{\alpha} - R_{s}i_{\alpha}) \mathrm{d}\tau + \phi_{\alpha}(0) \,, \\ \phi_{\beta}(t) &= \int_{0}^{t} (v_{\beta} - R_{s}i_{\beta}) \mathrm{d}\tau + \phi_{\beta}(0) \,, \\ \phi_{s} &= \frac{1}{\sqrt{3}} \sqrt{\phi_{\alpha}^{2} + \phi_{\beta}^{2}} \,, \end{aligned}$$

$$\begin{aligned} \mathrm{tg} \,\theta_{s} &= \frac{\phi_{\beta}(t)}{\phi_{\alpha}(t)} \,, \\ T_{e} &= P(\phi_{\alpha}i_{\beta} - \phi_{\beta}i_{\alpha}) \,. \end{aligned}$$

$$(27)$$

### 5.1 Selection of voltage vectors

The DSSM is supplied by two three-level VSIs. Each inverter can be controlled independently. Three-level inverter has 27 vectors, which has 8 redundant vectors in it. Figure 5 shows the voltage vectors of the first three-level inverter in  $\alpha - \beta$  frame, in which subscripts Z, S, M, and L denote zero, small, medium, and large voltages, respectively.

In order to minimize the torque and flux ripple in low speed, only the small voltage vector is applied when speed  $\Omega < \Omega_{\rm nom}/2$ . In this case the three-level inverter is similar to the two-level inverter. However if we consider the two inverters as a six-phase voltage source inverter we obtain a total of 64 switching modes. By using the transformation matrix (5) the 64 voltage vectors corresponding to the switching modes are projected on three planes. From 64 vectors there are only 12 voltage vectors that offer a maximum voltage on the  $\alpha-\beta$  plane and keep the harmonics on the  $Z_1 Z_2$  plane at a minimum [18].

The voltage vector plane is divided into twelve sectors so that each voltage vector divides each region into two equal parts as shown in Fig. 6. In each sector, four of the twelve voltage vectors may be used [19, 20]. The used control system is depicted in Fig. 7.

The switching table used is indicated in Table 2. The output of the flux hysteresis comparator is denoted as  $\Phi$ , the output of the torque hysteresis comparator is denoted as  $\tau$ . The flux hysteresis comparator is a two valued comparator and the torque hysteresis comparator is a three valued comparator.



Fig. 8. (a) – Performance of conventional DSSM DTC when  $\Omega < \Omega_{\text{nom}}/2$ , (b) – Permanents currents at  $T_r = 10 \text{ Nm}$  and  $\Omega = 100 \text{ rps} (\Omega < \Omega_{\text{nom}}/2)$ 

Table 2. The switching states for inverters when  $\Omega < \Omega_{nom}/2$ 

| Φ | au      | $\Theta_1$ | $\Theta_2$ | $\Theta_3$ | $\Theta_4$ | $\Theta_5$ | $\Theta_6$ | $\Theta_7$ | $\Theta_8$ | $\Theta_9$ | $\Theta_{10}$ | $\Theta_{11}$ | $\Theta_{12}$ |
|---|---------|------------|------------|------------|------------|------------|------------|------------|------------|------------|---------------|---------------|---------------|
|   | 1       | $u_3$      | $u_4$      | $u_5$      | $u_6$      | $u_7$      | $u_8$      | $u_9$      | $u_{10}$   | $u_{11}$   | $u_{12}$      | $u_1$         | $u_2$         |
| 1 | 0       | $u_1$      | $u_2$      | $u_3$      | $u_4$      | $u_5$      | $u_6$      | $u_7$      | $u_8$      | $u_9$      | $u_{10}$      | $u_{11}$      | $u_{12}$      |
|   | -1      | $u_{11}$   | $u_{12}$   | $u_1$      | $u_2$      | $u_3$      | $u_4$      | $u_5$      | $u_6$      | $u_7$      | $u_8$         | $u_9$         | $u_{10}$      |
|   | 1       | $u_5$      | $u_6$      | $u_7$      | $u_8$      | $u_9$      | $u_{10}$   | $u_{11}$   | $u_{12}$   | $u_1$      | $u_2$         | $u_3$         | $u_4$         |
| 0 | 0       | $u_7$      | $u_8$      | $u_9$      | $u_{10}$   | $u_{11}$   | $u_{12}$   | $u_1$      | $u_2$      | $u_3$      | $u_4$         | $u_5$         | $v_6$         |
|   | $^{-1}$ | $u_9$      | $u_{10}$   | $u_{11}$   | $u_{12}$   | $u_1$      | $u_2$      | $u_3$      | $u_4$      | $u_5$      | $u_6$         | $u_7$         | $u_8$         |

When the speed  $\Omega \geq \Omega_{\text{nom}}/2$  we use Table 1a to select the proper voltage vector for the first star and the Table 1b for the second star.

#### 5.2. Simulations results

The simulations results in Figs. 8 and 9 shows that basic DTC regulates the speed and stator flux well. We can see that, this control approach ensure good decoupling between stator flux linkage and torque. However, in this approach the phase currents are distorted as consequence a very high currents harmonics  $iz_1$  and  $iz_2$  especially when  $\Omega \ge \Omega_{\text{nom}}/2$  where we use medium and large voltages. Figure 10 gives the response of the torque and flux when the stator resistance increases. The obtained result shows that the torque and flux are oscillating and the global system is unstable. To overcome these problems, we propose in the next section a new DSSM-DTC method based on adaptive fuzzy control.

### 6 DTC BASED ON ADAPTIVE FUZZY CONTROL

In this section, our purpose is to synthesis a control laws in order to obtain almost sinusoidal phase currents and keep at a minimum currents harmonics  $iz_1$  and  $iz_2$ by using appropriate PWM modulation strategies. The proposed DTC adaptive fuzzy control problem is to force the system (21) to follow a given bounded reference signal, Fig. 11.

PROPOSITION 1. Consider the DSSM plant model (15). Let the ideal control laws be

$$v_x^* = \frac{1}{b_1} \left( -f_1 + \frac{d\phi_s^*}{dt} + k_1 e_1 \right),$$
  

$$v_y^* = \frac{1}{b_2} \left( -f_2 + \frac{dT_e^*}{dt} + k_2 e_2 \right),$$
  

$$T_e^* = \frac{1}{b_3} \left( -f_3 + \frac{d\Omega^*}{dt} + k_3 e_3 \right)$$
(28)

where

$$e_1 = \phi_s^* - \phi_s , \quad e_2 = T_e^* - T_e , \quad e_3 = \Omega^* - \Omega ,$$
  
 $k_1, k_2, k_3 > 0 .$  (29)

Then, the regulated errors  $e_i$  are asymptotically stable.

Proof. Consider the following Lyapunov function

$$V = \frac{1}{2} \left( e_1^2 + e_2^2 + e_3^2 \right). \tag{30}$$

Its time derivative is expressed by

$$\dot{V} = \dot{e}_1 e_1 + \dot{e}_2 e_2 + \dot{e}_3 e_3 \,. \tag{31}$$

The derivative of the errors (29) becomes:

$$\dot{e}_{1} = \frac{\mathrm{d}\phi_{s}^{*}}{\mathrm{d}t} - f_{1} - b_{1}v_{x},$$
  

$$\dot{e}_{2} = \frac{\mathrm{d}T_{e}^{*}}{\mathrm{d}t} - f_{2} - b_{2}v_{y},$$
  

$$\dot{e}_{3} = \frac{\mathrm{d}\Omega^{*}}{\mathrm{d}t} - f_{3} - b_{3}T_{e}.$$
(32)



Fig. 9. (a) – Performance of conventional DSSM DTC when  $\Omega \ge \Omega_{\text{nom}}/2$ , (b) – Permanents currents at  $T_r = 10$  Nm and  $\Omega = 200 \operatorname{rps}(\Omega \ge \Omega_{\text{nom}}/2)$ 



Fig. 10. Performance of conventional DSSM DTC with stator resistance variation



Fig. 11. DTC Adaptive Fuzzy Control scheme for DSSM

Replacing the control low (28) in (32), equation (31) becomes

Relation (33), means that  $e_1$ ,  $e_2$  and  $e_3$  converge at last asymptotically to zero. Therefore the control laws (28) can drive the plant system in the ideal case.

$$\dot{V} = -k_1 e_1^2 - k_2 e_2^2 - k_3 e_3^2 < 0.$$
(33)

In practice the nonlinear functions  $f_i$  (i = 1, 2, 3) involved in the state space model (21) are badly known. The implementation of the control law (28) is inoperative since it requires a precise model. To overcome this problem we use the fuzzy logic system to estimate on-line the nonlinear function given in (28). The ideal control law (28) is ideally rebuilt by a fuzzy logic system such that:

$$v_x^* = W_1^{\top}(\phi_s)\theta_1 + \varepsilon_1(\phi_s),$$
  

$$v_y^* = W_2^{\top}(T_e)\theta_2 + \varepsilon_2(T_e),$$
  

$$T_e^* = W_3^{\top}(w)\theta_3 + \varepsilon_3(w)$$
(34)

where  $W_i(\cdot)$  is a basis function [21] and  $\theta_i$  is vector of optimal parameters, while  $\varepsilon_i(\cdot)$  is the unavoidable reconstruction error satisfying the condition [22, 23]

$$|\varepsilon_1(\phi_s)| \le \bar{\varepsilon}_1, \quad |\varepsilon_2(T_e)| \le \bar{\varepsilon}_2, \quad |\varepsilon_3(w)| \le \bar{\varepsilon}_3.$$
 (35)

The ideal control law (34) is estimated by:

$$\hat{v}_x = W_1^{\top}(\phi_s)\hat{\theta}_1, \quad \hat{v}_y = W_2^{\top}(T_e)\hat{\theta}_2, \quad \hat{T}_e = W_3^{\top}(\Omega)\hat{\theta}_3$$
(36)

Where  $\hat{\theta}_i$ , the estimate of the best parameters  $\theta_i$ .

Subtracting (36) from (34), we obtain the estimation errors

$$v_x^* - \hat{v}_x = W_1^\top (\phi_s) \theta_1 + \varepsilon_1(\phi_s),$$
  

$$v_y^* - \hat{v}_y = W_2^\top (T_e) \tilde{\theta}_2 + \varepsilon_2(T_e),$$
  

$$T_e^* - \hat{T}_e = W_3^\top (w) \tilde{\theta}_3 + \varepsilon_3(w)$$
(37)

where  $\tilde{\theta}_i = \theta_i - \hat{\theta}_i$  is vector parametric error.

**PROPOSITION 2.** Consider the following control laws:

$$\hat{v}_x^* = \hat{v}_x + v_{xa}, \quad \hat{v}_y^* = \hat{v}_y + v_{ya}, \quad \hat{T}_e^* = \hat{T}_e + T_{ea} \quad (38)$$

with

$$v_{xa} = \hat{\bar{\varepsilon}}_1 \operatorname{sign}(e_1), \ v_{ya} = \hat{\bar{\varepsilon}}_2 \operatorname{sign}(e_2), \ T_{ea} = \hat{\bar{\varepsilon}}_3 \operatorname{sign}(e_3)$$
(39)

where  $\hat{\bar{\varepsilon}}_i$  and  $\hat{\theta}_i$  are updated under the laws

$$\dot{\hat{\varepsilon}}_1 = \eta_1 |e_1|, \quad \dot{\hat{\varepsilon}}_2 = \eta_2 |e_2|, \quad \dot{\hat{\varepsilon}}_3 = \eta_3 |e_3|, \quad (40)$$

$$\dot{\hat{\theta}}_1 = \gamma_1 W_1(\phi_s) e_1 , \ \dot{\hat{\theta}}_2 = \gamma_2 W_2^\top(T_e) e_2 , \ \dot{\hat{\theta}}_3 = \gamma_3 W_3^\top(\Omega) e_3$$
(41)

where  $\eta_i$  and  $\gamma_i$  are positives constants.

Then, the tracking errors  $e_i$  converge asymptotic toward zero and the parameters are bounded.

Proof . Let us choose the Lyapunov function

$$V = V_1 + V_2 + V_3 \tag{42}$$

with

$$V_{1} = \frac{1}{2b_{1}}e_{1}^{2} + \frac{1}{2\gamma_{1}}\tilde{\theta}_{1}^{\top}\tilde{\theta}_{1} + \frac{1}{2\eta_{1}}\tilde{\varepsilon}_{1}^{2},$$

$$V_{2} = \frac{1}{2b_{2}}e_{2}^{2} + \frac{1}{2\gamma_{2}}\tilde{\theta}_{2}^{\top}\tilde{\theta}_{2} + \frac{1}{2\eta_{2}}\tilde{\varepsilon}_{2}^{2},$$

$$V_{3} = \frac{1}{2b_{3}}e_{3}^{2} + \frac{1}{2\gamma_{3}}\tilde{\theta}_{3}^{\top}\tilde{\theta}_{3} + \frac{1}{2\eta_{3}}\tilde{\varepsilon}_{3}^{2}$$
(43)

where  $\tilde{\tilde{\varepsilon}}_i = \bar{\varepsilon}_i - \hat{\bar{\varepsilon}}_i$ . These functions are globally positive. Its time derivates are given by

$$\dot{V}_{1} = \frac{1}{b_{1}}e_{1}\dot{e}_{1} - \frac{1}{\gamma_{1}}\tilde{\theta}_{1}^{\top}\dot{\theta}_{1} - \frac{1}{\eta_{1}}\tilde{\varepsilon}_{1}\dot{\tilde{\varepsilon}}_{1},$$

$$\dot{V}_{2} = \frac{1}{b_{2}}e_{2}\dot{e}_{2} - \frac{1}{\gamma_{2}}\tilde{\theta}_{2}^{\top}\dot{\theta}_{2} - \frac{1}{\eta_{2}}\tilde{\varepsilon}_{2}\dot{\tilde{\varepsilon}}_{2},$$

$$\dot{V}_{3} = \frac{1}{b_{3}}e_{3}\dot{e}_{3} - \frac{1}{\gamma_{3}}\tilde{\theta}_{3}^{\top}\dot{\theta}_{3} - \frac{1}{\eta_{3}}\tilde{\varepsilon}_{3}\dot{\tilde{\varepsilon}}_{3}$$
(44)

where the time derivates of the tracking error dynamic are

$$\dot{e}_{1} = -k_{1}e_{1} + b_{1} \left( W_{1}^{\top}(\phi_{s})\tilde{\theta}_{1} + \varepsilon_{1}(\phi_{s}) - v_{xa} \right), \\ \dot{e}_{2} = -k_{2}e_{2} + b_{2} \left( W_{2}^{\top}(T_{e})\tilde{\theta}_{2} + \varepsilon_{2}(T_{e}) - v_{ya} \right), \qquad (45)$$
$$\dot{e}_{3} = -k_{3}e_{3} + b_{3} \left( W_{3}^{\top}(w)\tilde{\theta}_{3} + \varepsilon_{3}(w) - T_{ea} \right).$$

Substituting (35) and (45) in (44), it results

$$\dot{V}_{1} = -\frac{k_{1}}{b_{1}}e_{1}^{2} + \varepsilon_{1}(\phi_{s})e_{1} - v_{xa}e_{1} - \frac{1}{\eta_{1}}\tilde{\varepsilon}_{1}\dot{\tilde{\varepsilon}}_{1},$$
  
$$\dot{V}_{2} = -\frac{k_{2}}{b_{2}}e_{2}^{2} + \varepsilon_{2}(T_{e})e_{2} - v_{ya}e_{2} - \frac{1}{\eta_{2}}\tilde{\varepsilon}_{2}\dot{\tilde{\varepsilon}}_{2}, \qquad (46)$$
  
$$\dot{V}_{3} = -\frac{k_{3}}{b_{3}}e_{3}^{2} + \varepsilon_{3}(w)e_{3} - T_{ea}e_{3} - \frac{1}{\eta_{3}}\tilde{\varepsilon}_{3}\dot{\tilde{\varepsilon}}_{3}.$$

Introducing (40) into (46), we obtain

$$\dot{V}_{1} = -\frac{k_{1}}{b_{1}}e_{1}^{2} + \varepsilon_{1}(\phi_{s})e_{1} - \hat{\varepsilon}_{1}|e_{1}| - \frac{1}{\eta_{1}}\tilde{\varepsilon}_{1}\dot{\tilde{\varepsilon}}_{1},$$
  
$$\dot{V}_{2} = -\frac{k_{2}}{b_{2}}e_{2}^{2} + \varepsilon_{2}(T_{e})e_{2} - \hat{\varepsilon}_{2}|e_{2}| - \frac{1}{\eta_{2}}\tilde{\varepsilon}_{2}\dot{\tilde{\varepsilon}}_{2}, \qquad (47)$$
  
$$\dot{V}_{3} = -\frac{k_{3}}{b_{3}}e_{3}^{2} + \varepsilon_{3}(w)e_{3} - \hat{\varepsilon}_{3}|e_{3}| - \frac{1}{\eta_{3}}\tilde{\varepsilon}_{3}\dot{\tilde{\varepsilon}}_{3}.$$

Use the condition (35) yields

$$\dot{V}_{1} \leq -\frac{k_{1}}{b_{1}}e_{1}^{2} + \tilde{\varepsilon}_{1}|e_{1}| - \frac{1}{\eta_{1}}\tilde{\varepsilon}_{1}\dot{\tilde{\varepsilon}}_{1},$$
  
$$\dot{V}_{2} \leq -\frac{k_{2}}{b_{2}}e_{2}^{2} + \tilde{\varepsilon}_{2}|e_{2}| - \frac{1}{\eta_{2}}\tilde{\varepsilon}_{2}\dot{\tilde{\varepsilon}}_{2},$$
  
$$\dot{V}_{3} \leq -\frac{k_{3}}{b_{3}}e_{3}^{2} + \tilde{\varepsilon}_{3}|e_{3}| - \frac{1}{\eta_{3}}\tilde{\varepsilon}_{3}\dot{\tilde{\varepsilon}}_{3}.$$
  
$$(48)$$

By exploiting the update law (40), relation (48) becomes:

$$\dot{V}_1 \le -\frac{k_1}{b_1}e_1^2, \quad \dot{V}_2 \le -\frac{k_2}{b_2}e_2^2, \quad \dot{V}_3 \le -\frac{k_3}{b_3}e_3^2.$$
 (49)

Finally, we can write

$$V_1 \le -\frac{k_1}{b_1}e_1^2 - \frac{k_2}{b_2}e_2^2 - \frac{k_3}{b_3}e_3^2 \le 0.$$
 (50)

Hence using the Lyapunov theorem, we conclude that the control law (38) and the parameters adaptation law (40) and (41) ensure the asymptotic convergence of the tracking error  $e_i$  toward zero and the boundedness of the parameters.



Fig. 12. (a) – Performance of DTC adaptive fuzzy control of DSSM when  $\Omega < \Omega_{\text{nom}}/2$ , (b) – Permanents currents at  $T_r = 10$  Nm and  $\Omega = 100 \text{ rps} (\Omega < \Omega_{\text{nom}}/2)$ 



Fig. 13. (a) – Performance of DTC adaptive fuzzy control of DSSM when  $\Omega \ge \Omega_{\text{nom}}/2$ , (b) – Permanents currents at  $T_r = 10$  Nm and  $\Omega = 200 \text{ rps} (\Omega \ge \Omega_{\text{nom}}/2)$ 

### 6.1 Simulations results

# Figures 12 and 13 give the performance of DTC adaptive fuzzy control of DSSM fed by two tree level inverter PWM respectively when $\Omega < \Omega_{\text{nom}}/2$ and $\Omega \ge \Omega_{\text{nom}}/2$ . It appears that, the speed and stator flux follows their reference, the torque and the flux are decoupled and the harmonic currents are greatly reduced. Figure 14 gives the response of the torque and flux when the stator resistance increases. The obtained result shows that the torque and flux follows their reference and the system is stable.

In this paper, a new direct torque control method of DSSM is presented and its merits over the conventional DTC approach are confirmed by the simulation results. This new direct torque control uses the adaptive controller based on fuzzy systems. The proposed method uses the Sugeno-Takagi fuzzy system as a universal approximator of continuous nonlinear functions. The adaptive control law is established based on the Lyapunov method. So, the output convergence, the boundedness of

7 CONCLUSION



Fig. 14. Performance of DTC adaptive fuzzy control of DSSM with stator resistance variation

au

 $\delta$ 

 $\theta_r, \theta_s$ 

the parameters and the states are derived. Moreover, the fuzzy adaptive law incorporates a compensatory sliding term, which compensates for effects of the unavoidable reconstruction errors. The features of the proposed direct torque control method include 1) providing constant inverter switching frequency, 2) significantly reducing the torque ripple and harmonic currents on the  $z_1-z_2$  plane, 3) good robustness to the variation of the stator resistance.

# Appendix 1. List of principal symbols

| пррения                               | 1. List of principal symbols                              |
|---------------------------------------|---|
| $i_{a1}, i_{b1}, i_{c1}$ :            | stator current $a, b, c$ phase of the first wind-         |
|                                       | ing,  |
| $i_{a2}, i_{b2}, i_{c2}$ :            | stator current $a, b, c$ phase of the second              |
|                                       | winding,  |
| $i_{\alpha}, i_{\beta}$ :             | stator current $\alpha - \beta$ axis,                     |
| $i_x, i_y$ :                          | stator current $x$ - $y$ axis,                            |
| $i_s$ :                               | stator current vector,                                    |
| $v_{a1}, v_{b1}, v_{c1}$ :            | simple voltage of stator three phase first                |
|                                       | simple voltage of stater three phase second               |
| $v_{a2}, v_{b2}, v_{c2}.$             | winding   |
| a, .                                  | stator voltage vector                                     |
| <i>U<sub>S</sub></i> .                | stator voltage vector,                                    |
| $v_d, v_q$ .                          | stator voltages $a-q$ axis,                               |
| $v_{\alpha}, v_{\beta}$ :             | stator voltages $\alpha - \beta$ axis,                    |
| $v_x, v_y$ :                          | stator voltages $x-y$ axis,                               |
| $L_d, L_q$ :                          | d-q inductances,  |
| $L_{ss}$ :                            | stator inductance matrix,                                 |
| $M_{sr}$ :                            | stator-rotor mutual inductance matrix,                    |
| $\mathbf{R}_s$ :                      | $\operatorname{diag}(R_s \ R_s \ R_s \ R_s \ R_s \ R_s),$ |
| $R_s$ :                               | stator resistance,  |
| $R_f$ :                               | rotor resistance,   |
| $T_{e}, T_{e}^{*}$ :                  | electromagnetic torque, reference torque,                 |
| $T_r$ :                               | load torque,  |
| $\varphi_s, \varphi_s^*$ :            | stator flux vector, reference flux vector,                |
| $\varphi_d, \varphi_q$ :              | stator flux $d-q$ axis,                                   |
| $\varphi_{\alpha}, \varphi_{\beta}$ : | stator flux $\alpha - \beta$ axis,                        |
| $\varphi_x, \varphi_y$ :              | stator flux $x-y$ axis,                                   |
| Ω :                                   | rotating speed of rotor flux linkage,                     |
| w :                                   | rotating speed of rotor flux linkage,                     |
| $w_s$ :                               | rotating speed of stator flux linkage,                    |
| $\Phi$ :                              | output of the flux hysteresis comparator.                 |
|                                       | - • • • • • •   |

: output of the torque hysteresis comparator,

: angle between rotor and stator flux linkage,

: angle of rotor flux linkage, angle of stator flux linkage,

 $\Theta_i$  : the region numbers for the stator linkage positions,

 $E_T, E_{\varphi}$  : torque error, flux error,

J : moment of inertia,

 $f_r$  : friction coefficient,

P : number of pole pairs.

### Appendix 2. DSSM parameters

 $\begin{array}{ll} P_n = 5 \ {\rm kW}, & U_c = 232 \ {\rm V}, & i_f = 1 \ {\rm A}, & R_s = 2.35 \ {\Omega}, \\ R_f = 30.3 \ {\Omega}, \ L_d = 0.3811 \ {\rm H}, \ L_f = 15 \ {\rm H}, \ M_d = 2.146 \ {\rm H}, \\ J = 0.05 \ {\rm Nms}^2/{\rm rd}, & f_r = 0.001 \ {\rm Nms}/{\rm rd}, \ P = 1. \end{array}$ 

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#### Received 10 July 2010

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