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Chapter 6

Fuzzy Direct Torque-controlled Induction Motor Drives for Traction with Neural Compensation of Stator Resistance

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Additional information is available at the end of the chapter

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Abstract

In this chapter, a new method for stator resistance compensation in direct torque control (DTC) drives, based on neural networks, is presented. The estimation of electromagnetic torque and stator flux linkages using the measured stator voltages and currents is crucial to the success of DTC drives. The estimation is dependent only on one machine parameter, which is the stator resistance. Changes of the stator resistances cause errors in the estimated magnitude and position of the flux linkage and therefore in the estimated electromagnetic torque. Parameter compensation by means of stator current phasor error has been proposed in literature. The proposed approach in this chapter is based on a principle that states the error between the measured current magnitude of the stator feedback and the stator's command, verified with neural network, is proportional to the variation of the stator resistance and is mainly caused by the motor temperature and the varying stator frequency. Then the correction value of stator resistance is achieved by means of a fuzzy controller. For the first time, a combination of neural control and fuzzy control approach in stator resistance variations based on the stator current is presented. The presented approach efficiently estimates the correct value of stator resistance.

Keywords: Fuzzy direct torque control, neural compensation, induction motor drives

1. Introduction

The direct torque control is one of the excellent control strategies available for torque control of induction machine. It is considered as an alternative to field oriented control (FOC) technique [1]. In fact, among all control methods for induction motor drives, direct torque

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control (DTC) seems to be particularly interesting being independent of machine rotor parameters and requiring no speed or position sensors [2].

A basic concept of direct torque control of induction motor drives is simultaneous control of the stator flux and electromagnetic torque of a machine. Compared to the conventional vector-controlled drives, the torque and flux of a DTC-based drive are controlled in a closed-loop system that does not use the current loops.

In principle, DTC-based drives require only the knowledge of stator resistance and thereby decrease the associated sensitivity to parameter variations [3, 4]. Moreover, compared to the conventional vector-controlled drives, DTC-based drives do not require fulfilling the coordinate transformation between stationary and synchronous frames. Depending on how the switching sectors are selected, two different DTC schemes become possible [5].

Since a DTC-based drive selects the inverter switching states using a switching table, neither the current controllers nor the pulse-width modulation (PWM) modulator is required. As a result, the DTC-based drive provides a fast torque response [6]. The conventional direct torque control (CDTC) suffers from some drawbacks such as high current, flux and torque ripple, difficulties in torque, and flux control at very low speeds [7]. However, the switching-table-based DTC approach has some disadvantages. If the switching frequency of the inverter is not high, the torque and flux pulsation could be high; moreover, there would be a sluggish response during the start-up or change of the reference flux or reference torque [8]. Hence, to improve the performance of the DTC drive during the start-up or changes in the reference flux and torque, a fuzzy-logic-based switching-vector process is developed in this chapter [9–15]. In DTC drives, the feedback of the electromagnetic torque and stator flux linkage is used as the input of controller. Using the measured stator currents and voltages, the electromagnetic torque and also stator flux linkages are estimated in stator reference frames [16, 17]. "The machine model is only dependent on stator resistance" [18]. There are different forms of direct torque control induction motor based on how currents and voltages are measured or estimated [19–21]. The stator current might be obtained using only the DC-link current sensor, and the motor line voltages could be reconstructed inexpensively using gate signals [22]. Nevertheless, all the measured values suffer from precision and low-speed operational problems caused by errors induced by the varying stator resistance in the flux and its angle calculator [23, 24]. The stator resistance change has a wide range, varying from 0.75 to 1.7 times the stator's nominal value. The variation is largely due to temperature variations, and to a small extent, due to the stator frequency variations [21]. The variation deteriorates the drive performance by introducing errors in the estimated magnitude and position of the flux linkage and therefore in the electromagnetic torque estimation, particularly at low speeds [25]. Note that at low speeds, the voltage drops of the stator resistance constitute a significant portion of the applied voltages. Only a few control schemes have been proposed so far for overcoming the mentioned parameter sensitivity (which restricts the speed control range of the drives). The stator resistance has problems such as convergence and slowness of response. A partial operating-frequency-dependent hybrid-flux estimator has been proposed for tuning the stator resistance [10]. Adjustment of the stator resistance, based on the difference between the flux current and its command, has problems in identifying the actual flux current [26, 27]. Finding the stator

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resistance based on the steady state voltage has the shortcoming of using direct axis flux linkages that are affected by the stator resistance variations.

In this chapter, a neural network estimator is developed to find the reference stator current values at each moment. Later, the error difference between the measured and the real stator current values is fed to a fuzzy logic controller, which then outputs the correct stator resistance value.

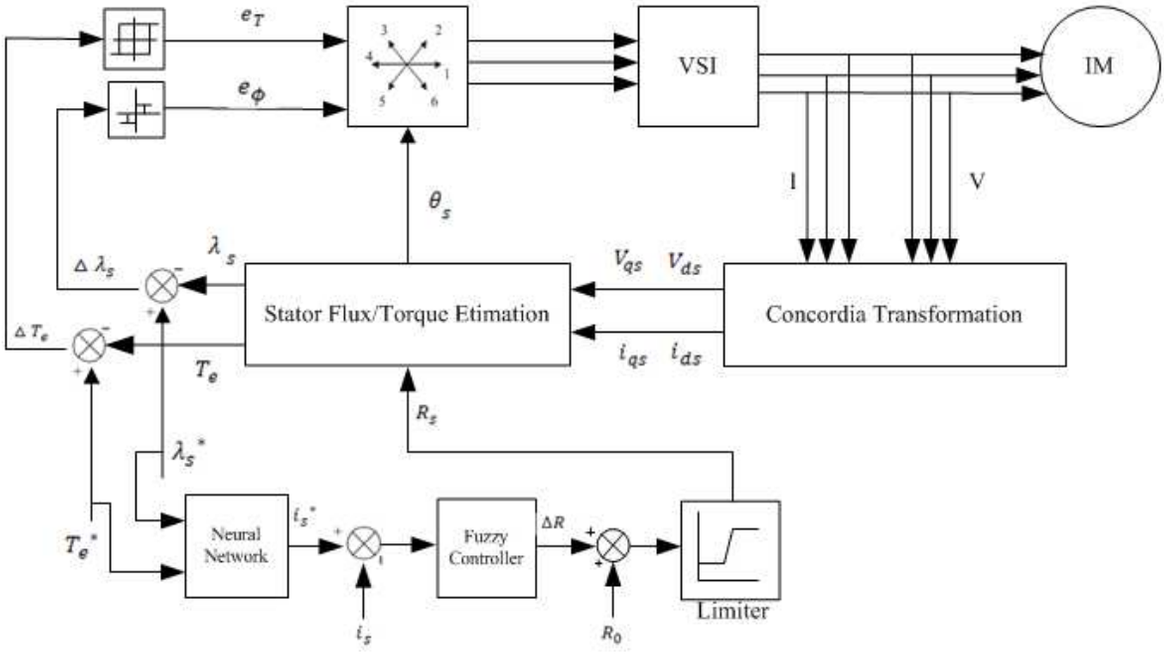


Figure 1. Block diagram of the fuzzy direct torque control of induction motor drives with a stator resistance estimator.

2. Fuzzy logic direct torque control

In this section, the concept and principle of direct torque control approach of an induction motor is briefly introduced. A schematic diagram of the proposed drive is shown in Fig. 1. The feedback control of torque and stator flux linkages, which are estimated from the measured voltages and currents of the motor, is used in the proposed drive scheme. In this approach, stator-reference frame model of the induction motor is used. To avoid the trigonometric operations faced in coordinate transformations of other reference frames, the same reference frame is used in the implementation [22]. This can be considered as one of the advantages of the control scheme. Through the integration of the difference between the phase voltage and the voltage drop in the stator resistance, Stator q and d and axis flux linkages λ_{qs}, λ_{ds} can be calculated as follows:

$$\lambda_{qs} = \int (V_{qs} - R_s i_{qs}) dt \tag{1}$$

$$\lambda_{ds} = \int (V_{ds} - R_s i_{ds}) dt \tag{2}$$

And the flux linkage phasor is as follows:

$$\lambda_s = \sqrt{\lambda_{qs}{}^2 + \lambda_{ds}{}^2)} \tag{3}$$

The stator flux linkage phasor position is:

$$\theta_s = \tan^{-1}(\frac{\lambda_{qs}}{\lambda_{ds}}) \tag{4}$$

And the electromagnetic torque is given by:

$$T_e = \frac{3}{2}\frac{P}{2}(i_{qs}\lambda_{ds} - i_{ds}\lambda_{qs}) \tag{5}$$

According to Fig. 2, the inverter switching states are selected based on the errors of the torque and the flux (as indicated by ΔT_e and $\Delta\lambda_s$, respectively). Provided that

$$\begin{aligned} \Delta T_e &= T_e^* - T_e \\ \Delta\lambda_s &= \lambda^*{}_s - \lambda_s \end{aligned} \tag{6}$$

The optimum switching vector is selected to decrease the errors [23, 25, 26]. Using a fuzzy-logic-based switching-vector selection process, it would be possible to improve the performance of the DTC drive during start-up or changes in the reference flux and torque. For this, a Mamdani fuzzy-logic-based system is used. Using the flux and torque deviation from reference ones and the position of the stator flux linkage space vector, it is possible to select different voltages. Then a rule-base has to be formulated based on these states. Thus the aim of the approach is to use a fuzzy logic system to expand the system performance (i.e., gives faster torque and flux response), outputs the zero and non-zero voltage switching states (n), and uses three quantities as its inputs: e_Φ, the torque error (e_T), and the position of the stator flux space vector (θ_s). The stator flux linkage space vector can be located in any of the twelve sectors, each spanning a 60° wide region. These regions overlap each other as shown in Table 1.

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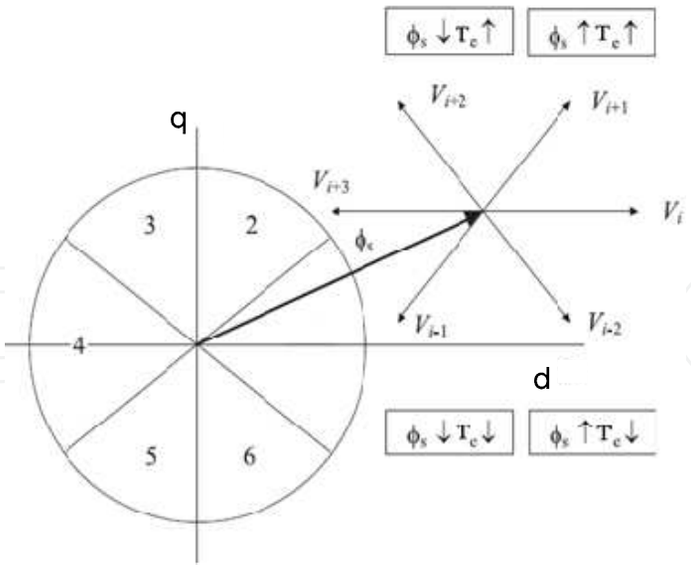


Figure 2. Changes of flux linkage space vectors due to the switching vectors.

S1	**S2**	**S3**	**S4**	**S5**	**S6**
315–75	345–45	15–75	45–105	75–135	105–165
S7	**S8**	**S9**	**S10**	**S11**	**S12**
135-195	165-225	195-255	225-285	255-315	285-345

Table 1. Overlaps between sectors

$e_\Phi e_t$	**P**	**ZE**	**N**
PL	1	2	2
PS	1	2	3
ZE	0	0	0
NS	6	0	4
NL	6	5	5

Table 2. Fuzzy vector selection in sector 1

Since it was assumed that there were three and five fuzzy sets for the flux error and the torque error, respectively, there will be 15 rules for every sector. Table 2 shows the various rules for sector 1. In particular, the stator flux error (e_Φ) can be positive (P), zero (ZE), or negative (N), corresponding to three overlapping fuzzy sets. The electromagnetic torque error can be positive large (PL), positive small (PS), zero (ZE), negative small (NS), or negative large (NL). This is because the intention is to make the torque variations smaller. Therefore, the universe

of the torque is divided into five overlapping fuzzy sets. The various membership functions are shown in Fig. 3. Since there are 12 sectors, the total number of rules becomes 180. Each one of the rules can be described by the input variables and the control variable, which is the switching state (n). For example, Table 2 shows various rules for sector 1 as below:

Rule 1: If e_Φ is positive (P), e_T is positive large (PL) and θ_s is S1, then n is 1.

Rule 2: If e_Φ is positive (P), e_T is positive small (PS) and θ_s is S2, then n is 1.

Rule 3: If e_Φ is positive (P), e_T is ZE and θ_s is S3, then n is 0.

The goal of the fuzzy system is to obtain a crisp value (as the appropriate switching state) on its output. A general "*i*th rule" has the following form:

Rule *i*: If e_Φ is A_i, e_T is B_i and θ_s is C_i, then n is N_i

Thus, by using the minimum operation for the fuzzy *AND* operation and the firing strength of the *i*th rule, α_i can be obtained from

$$\alpha_i = \min[\mu_{Ai}(e_\Phi), \mu_{Bi}(e_T), \mu_{ci}(\theta_s)] \tag{7}$$

where $\mu_{Ai}(e_\Phi)$, $\mu_{Bi}(e_T)$, $\mu_{ci}(\theta_s)$ are membership functions of fuzzy sets A_i, B_i, and C_i of the variables flux error, the torque error, and the flux position, respectively. The output form of the *i*th rule is obtained from

$$\mu_{Ni}(n) = \min[\alpha_i, \mu_{Ni}(n)] \tag{8}$$

where $\mu_{Ni}(n)$ is the membership function of fuzzy set N_i of variable n. Therefore, the overall (or the combined) membership function of output n is gained by using the max operator as follows:

$$\mu_N(n) = \max[\mu_{Ni}(n)] \tag{9}$$

In this case, the outputs include crisp numbers, switching states, and for defuzzification, the maximum used criteria.

3. Stator resistance compensation

3.1. Scheme

A mismatch between the controller-set stator resistance and its actual value in the machine can create the instability shown in Fig. 6a. This figure shows the simulations for the changes of the

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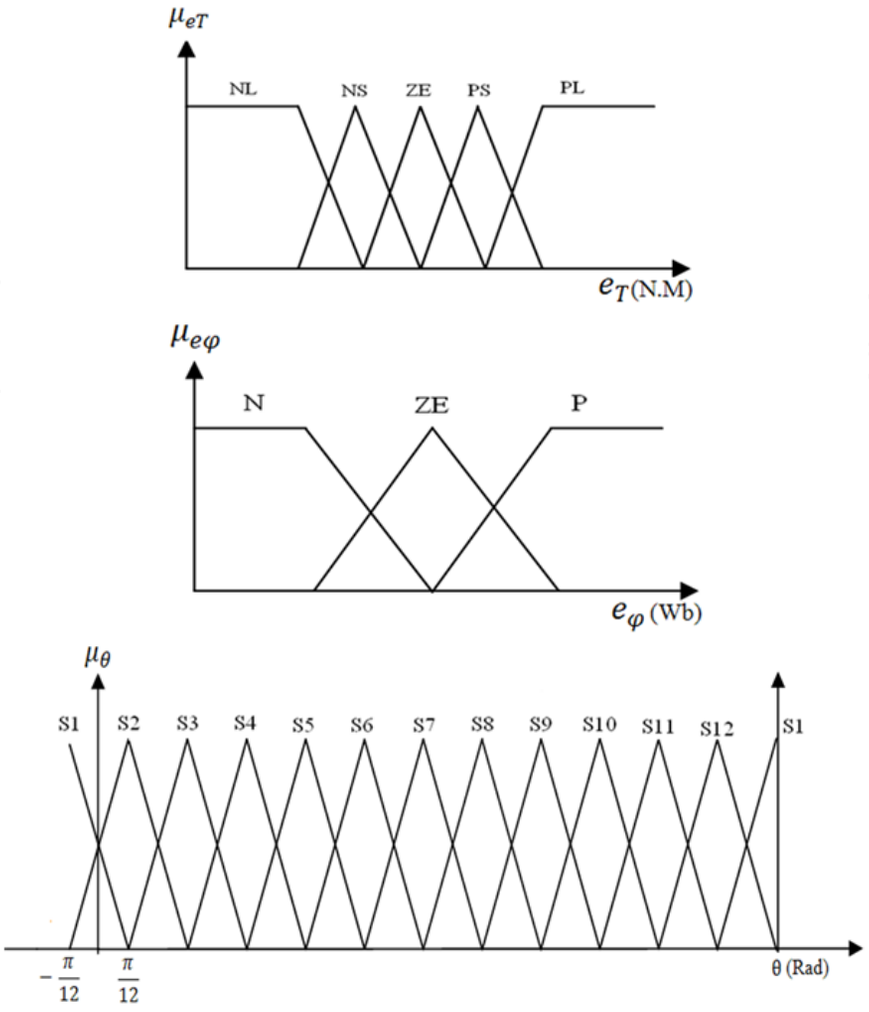


Figure 3. Membership functions.

step stator resistance from 100% to 50% of its nominal value at second 0.5. The drive system becomes unstable if the controller-instrumented stator resistance is higher than its actual value in the motor [11]. An explanation for this could be as follows: when motor resistance decrease in machine and the applied voltage is the same, the current increases, resulting in increased flux and electromagnetic torque [28]. The opposite effect occurs in controller. In fact, by current increments, which are inputs of the system, the stator resistance voltage drops will increase in the calculator. Therefore, lower flux linkages and electromagnetic torque estimations will present. Compared with their command values, they give large torque and flux linkages deviations, which result in commanding larger voltages and currents and leading to a run off condition as shown in Fig. 6a. "The parameter mismatch between the controller and machine will result in a nonlinear relation between the torque and the torque's reference, making it a non-ideal torque amplifier" [29]. This will have undesirable effect in a torque drive and speed-controlled drive systems. Therefore, it will be reasonable to design a motor resistance adaption law to overcome instability and to guarantee a linear torque amplifier in the DTC drive. A new approach is presented in the next section for stator resistance parameter adaption.

3.2. Stator current phasor command

A diagram of the applied stator resistance compensation is shown in Fig. 4. The presented technique is based on the principle that the error between the measured stator feedback current-phasor magnitude i_s and the stator's command i^*_s is proportional to the stator resistance variation, which is mainly caused by the motor temperature and the varying stator frequency. The correction value is obtained by means of a fuzzy controller. The final estimated value of $\hat{R}_s$ is obtained as the output of the limiter. The above algorithm requires the stator current phasor command, which is a function of the commanded torque and the commanded stator flux linkages.

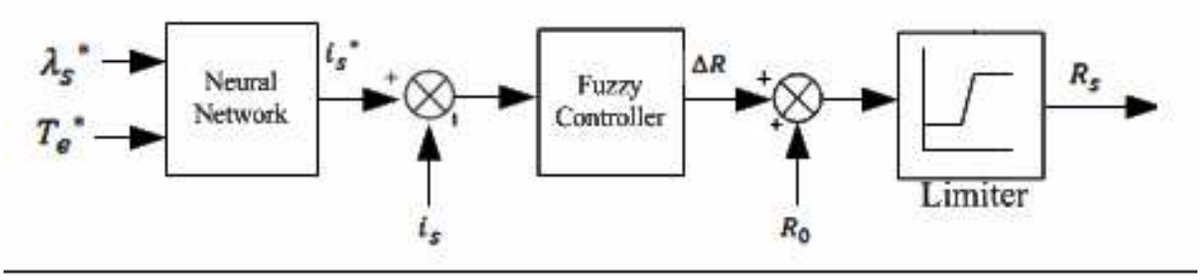


Figure 4. Block diagram of the adaptive stator resistance compensator.

A neural network estimator, presented in the following, is designed to evaluate the stator current command from the torque and stator flux linkage commands.

The stator feedback current phasor magnitude i_s is obtained from the q and d axis measured currents as

$$i_s = \sqrt{(i_{qs}^{\ 2} + i_{ds}^{\ 2})} \tag{10}$$

The stator command current phasor magnitude i^*_s is derived from the dynamic equations of the induction motor in the synchronous-rotating reference frame, using the torque command T^*_e and the stator flux linkage command λ^*_s and aligning the d axis with the stator flux linkage phasor as

$$\lambda^e_{qs} = 0, p\lambda^e_{qs} = 0, \lambda^e_{ds} = \lambda^*_s \tag{11}$$

where p is the number of poles. Substituting these equations in flux linkages and torque equations results in

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$$T_e = \frac{3}{2}\frac{2}{p} i_{qs}^{\ e} \lambda_s \tag{12}$$

Then the q axis current command is directly obtained by using the torque command T^*_e and the stator flux linkage command λ^*_s as

$$i^{e*}_{qs} = \frac{2}{3}\frac{2}{p}\frac{T^*_e}{\lambda^*_s} \tag{13}$$

It can be shown that i^{e*}_{ds} is given by

$$\begin{aligned} &L_s(i_{ds}^{\ e*})^2 - \lambda_s^{\ *}(1 - \frac{L_s L_r}{L^2_m - L_s L_r}) i_{ds}^{\ e*} + \\ &L_s(i_{qs}^{\ e*})^2 - \frac{(\lambda_s^{\ *})^2 L_r}{L^2_m - L_s L_r} = 0 \end{aligned} \tag{14}$$

Equation 14 gives two solutions for i^{e*}_{ds}, and the appropriate solution is the one that outputs a smaller value. Finally, the stator current command is calculated from

$$i^*_s = \sqrt{(i^{e*}_{qs}{}^2 + i^{e*}_{ds}{}^2)} \tag{15}$$

It is shown here that evaluation of the stator current command is a complicated and time-consuming process. Instead of using the numerical solution for the system, it is possible to perform the stator current command by using an artificial neural network (ANN) since it is known that ANN is a general nonlinear function estimator. As a result, a multilayer feed-forward back-propagation ANN, whose inputs are the torque and flux reference values, is trained to estimate the stator current command. A 2-8-8-1 structure, which has two hidden layers with 8 hidden nodes, is obtained by trial and error. The activation functions of the hidden layers are tan-sigmoid functions. Fig. 5 shows the structure of the ANN estimator. The neural estimator evaluates the reference stator current with less than 0.01% error. Furthermore, it is shown that more complicated ANN structures result in higher error rates.

4. Results

Dynamic simulations are performed to validate the performance of the proposed technique. The induction motor details, used in the simulation, are given in the appendix. Fig. 6a and 6b show the simulations for a step change in the stator resistance parameter-uncompensated and compensated torque drive system respectively. The system controller has the nominal value

of the stator resistance, and after half a second, the stator resistance is changed to ½ of its nominal value. Then the corresponding effects are studied. In the compensated system, it is observed that the estimation of stator resistance has experienced an initial transient state, and after a short time, it converges gradually to its final actual value in a steady state. The similar transitions are observed in other variables. However, all variables reach to their steady state situation. A step variation in the stator resistance is rather an extreme test and not a significant case encountered in practice. In real operating conditions, the temperature change rate is very slow and so is the stator resistance.

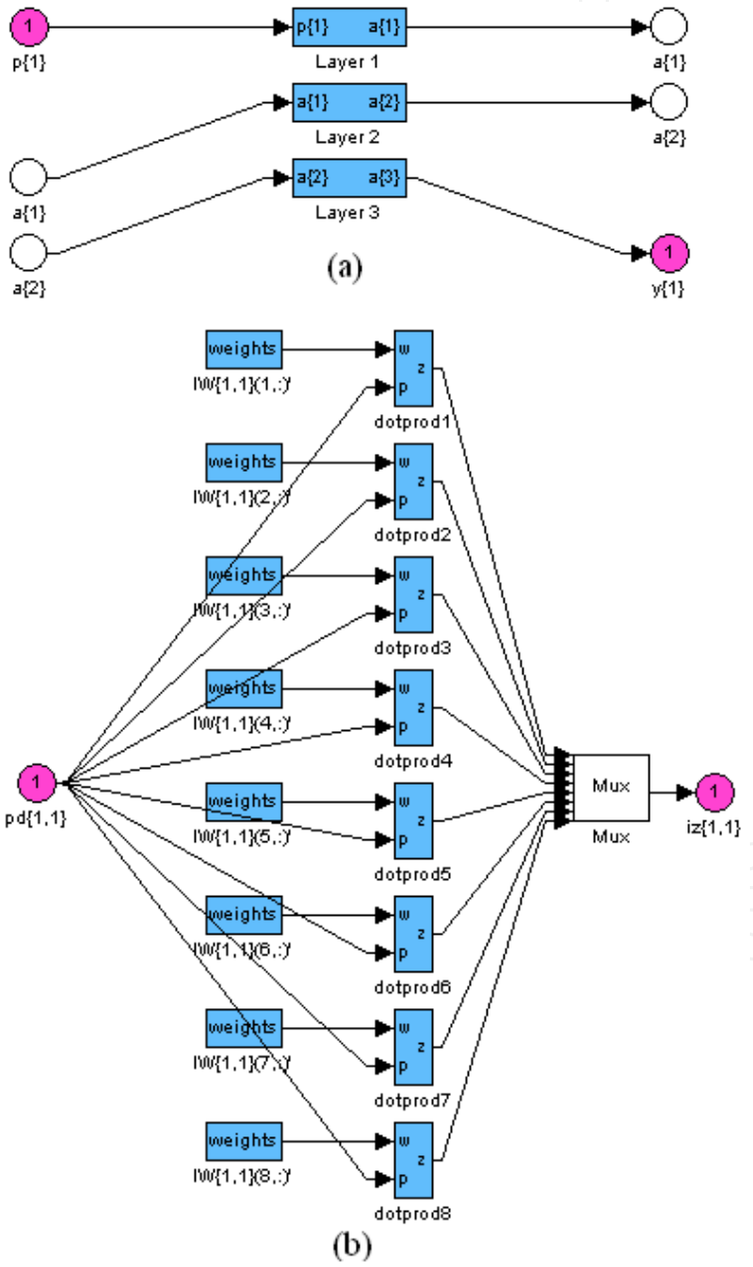


Figure 5. Neural network structure: (a) structure of layers and (b) structure of first hidden layer.

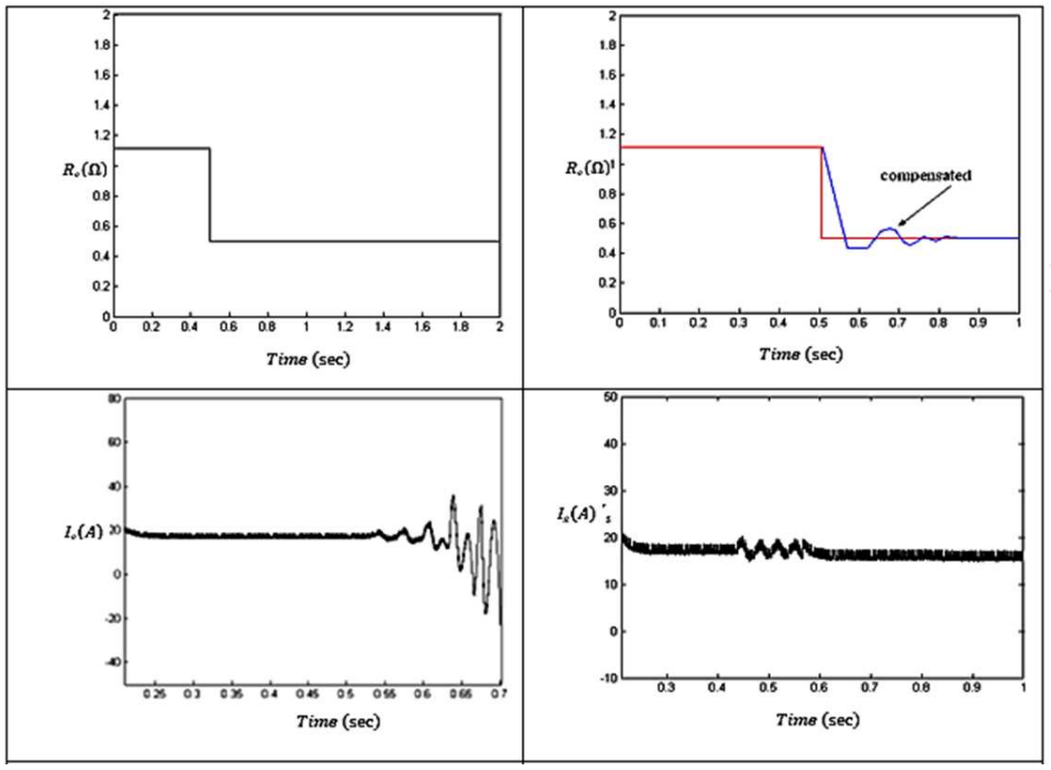
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Stator flux linkages and the torque command are proportionally decreased and increased linearly from/to their original reference values. The tracking of motor variables and stator resistance is achieved, thus proving the effectiveness of the adaptive controller in the flux-weakening region. It also perfectly operates in stator resistance incremental case and in gradually stator resistance changes due to temperature changes. In these cases, there is not any oscillation even at the initial moments of resistance variations.

5. Conclusion

A fuzzy direct torque-controlled drive was introduced, and an adaptive stator resistance compensation scheme was applied to a typical three-phase induction motor. With this approach, the elimination of parameter sensitivity of the stator resistance by using only the existing stator current feedback occurred. The scheme was simple to implement, and its realization was indirectly dependent on stator inductances. Since the flux was controlled in the machine, the inductances used in the computation of stator phasor current command were constants. A procedure for finding the phasor command of the stator current from the torque and stator flux linkage commands was derived to realize the complication of this method. The ANN estimator was designed to effectively evaluate the reference stator current value. The scheme was verified via dynamic simulation for various operating conditions, including the flux-weakening mode. The scheme was successful despite rapid changes in the stator resistance, such as step changes. It was observed that the scheme adapted very well without transients even for simultaneous variations of the torque and flux linkages command while the stator resistance was varying. Finally, a simple fuzzy controller was used to generate the exact stator resistance value.

Variable	Symbol	Variable	Symbol
Stator q and d flux linkages	λ_{qs} , λ_{ds}	Stator command current phasor magnitude	i^*_s
Stator q and d and axis voltages	V_{qs} , V_{ds}	Torque command	T^*_e
Stator resistance	R_s	Stator flux linkage command	λ^*_s
Stator q and d and axis currents	i_{qs} , i_{ds}	Stator self-inductance	L_s
Stator flux linkage phasor position	θ_s	Mutual inductance	M
Motor torque	T_e	Rotor self-inductance	L_r
Number of poles	p		



(a)

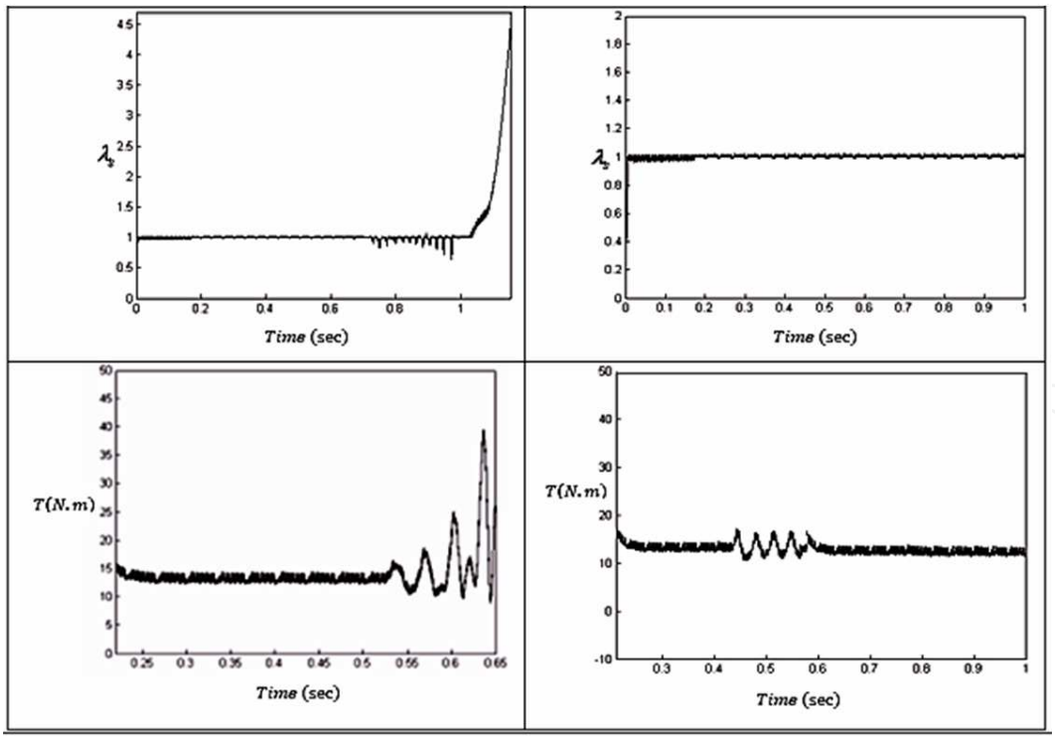


Figure 6. (a) The step response for a parameter uncompensated system. (b)The step response for a parameter compensated system.

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6. Appendix

Induction motor parameters

Nominal operation condition		
Rated power	4	kW
Rated voltage	480	V
Rated speed	3000	rpm
Rated frequency	50	Hz

Parameters of simulated induction motor		
Stator resistance	1.115	Ω
Stator self-inductance	0.005974	H
Mutual inductance	0.2037	H
Friction factor	0.00051	N·m/s
Rotor resistance	1.083	Ω
Rotor self-inductance	0.005974	H
Rotor inertia	0.0333	kg/m^2
Number of poles	2	

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