Single-Phase Induction Motor Drives with Direct Torque Control

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Abstract – **Although many techniques for high performance torque and flux control in three-phase induction motors are already available, most of them were not generalized for application in single-phase induction motor drives. In this paper some high performance single-phase induction motor drive systems are investigated. A stator fixed reference frame dq model is used to develop a direct torque control scheme for single-phase machines having main and auxiliary windings. A stator flux oriented reference frame control scheme is also proposed. In both strategies presented no speed or position signal for the torque control is necessary. Simulation results are provided to show the techniques effectiveness.**

I. INTRODUCTION

Three-phase induction motor drives have been thoroughly discussed during the last decades. Many field oriented control (FOC) and direct torque control (DTC) schemes have been proposed.

Single-phase induction motors with main and auxiliary stator windings can be viewed as two-phase machines, since these windings are displaced 90 degrees from each other. A four-switch inverter topology as shown in Fig. 1 has been proposed as a low cost solution for single-phase drives [1].

Some single-phase high performance induction motor drives were proposed recently [2], [3]. In these schemes, rotor field oriented control principles are adapted to the single-phase induction motor model and then, the speed signal is necessary for the torque control. Since single-phase induction motors are used in low power - low cost applications, the need for an encoder is a great disadvantage of these drives.

In this paper, some DTC strategies for single-phase induction motor drives are investigated. In the schemes proposed, as in most DTC three-phase drives, neither the speed nor the position is necessary for the torque and flux control. Furthermore, no current controllers are necessary. This may significantly reduce the drive cost for systems where the speed does note have to be controlled.

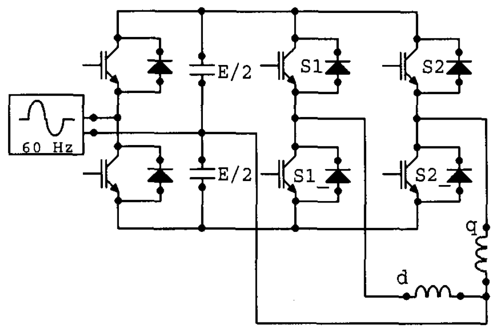

Fig. 1. Four switch inverter for single-phase drives.

In the first DTC method presented, hysteresis control is used in a way similar to the proposed in [4] for three-phase drives. The other strategy uses the stator flux oriented dq model to define PI controllers for torque and flux control. Simulation results are presented to show the effectiveness of the proposed control schemes.

II. SINGLE-PHASE INDUCTION MOTOR MODEL

B. Stationary Reference Frame

The single-phase induction motor model equations are more complex than that of the three-phase induction machines due to the fact that the main and auxiliary stator windings have different resistances and inductances.

Referring all the windings as if they had N_{sq} turns (the number of turns of the stator q axis winding), the single-phase induction motor model in a stationary reference frame can be described by the following equations [5]:

$$v_{sd} = R_{sd} i_{sd} + \frac{d\lambda_{sd}}{dt} \quad (1)$$

$$v_{sq} = R_{sq} i_{sq} + \frac{d\lambda_{sq}}{dt} \quad (2)$$

$$0 = R_r i_{rd} + \frac{d\lambda_{rd}}{dt} + \omega_r \lambda_{rq} \quad (3)$$

$$0 = R_r i_{rq} + \frac{d\lambda_{rq}}{dt} - \omega_r \lambda_{rd} \quad (4)$$

$$\lambda_{sd} = L_{sd} i_{sd} + L_{mq} i_{rd} \quad (5)$$

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$$\lambda_{sq} = L_{sq} i_{sq} + L_{mq} i_{rq} \tag{6}$$

$$\lambda_{rd} = L_{mq} i_{sd} + L_r i_{rd} \tag{7}$$

$$\lambda_{rq} = L_{mq} i_{sq} + L_r i_{rq} \tag{8}$$

$$T_e = \frac{P}{2}\frac{L_{mq}}{L_r}\left(\lambda_{rd} i_{sq} - \lambda_{rq} i_{sd}\right) \tag{9}$$

It should be noted that the rotor quantities and the d axis stator quantities are referred to the stator q axis winding. Thus, in order to use the model in any control scheme, the measured values of v_{sd} and i_{sd} must be multiplied by N_{sq}/N_{sd} and N_{sd}/N_{sq}, respectively.

B. Arbitrary reference frame

The stationary reference frame model can be written in an arbitrarily rotating reference frame as:

$$v_{sd} = R'_{sd} i_{sd} + R'_{sm} i_{sq} + \frac{d\lambda_{sd}}{dt} - \omega_e \lambda_{sq} \tag{10}$$

$$v_{sq} = R'_{sm} i_{sd} + R'_{sq} i_{sq} + \frac{d\lambda_{sq}}{dt} + \omega_e \lambda_{sd} \tag{11}$$

$$0 = R_r i_{rd} + \frac{d\lambda_{rd}}{dt} - (\omega_e - \omega_r)\lambda_{rq} \tag{12}$$

$$0 = R_r i_{rq} + \frac{d\lambda_{rq}}{dt} + (\omega_e - \omega_r)\lambda_{rd} \tag{13}$$

$$\lambda_{sd} = L_{sd} i_{sd} + L_{mq} i_{rd} \tag{14}$$

$$\lambda_{sq} = L_{sq} i_{sq} + L_{mq} i_{rq} \tag{15}$$

$$\lambda_{rd} = L_{mq} i_{sd} + L_r i_{rd} \tag{16}$$

$$\lambda_{rq} = L_{mq} i_{sq} + L_r i_{rq} \tag{17}$$

$$T_e = \frac{P}{2}\frac{L_{mq}}{L_r}\left(\lambda_{rd} i_{sq} - \lambda_{rq} i_{sd}\right) \tag{18}$$

where

$$R'_{sd} = R_{sd}\cos^2\theta_e + R_{sq}\sin^2\theta_e \tag{19}$$

$$R'_{sq} = R_{sd}\sin^2\theta_e + R_{sq}\cos^2\theta_e \tag{20}$$

$$R'_{sm} = \left(R_{sq} - R_{sd}\right)\sin\theta_e\cos\theta_e \tag{21}$$

$$L'_{sd} = L_{sd}\cos^2\theta_e + L_{sq}\sin^2\theta_e \tag{22}$$

$$L'_{sq} = L_{sd}\sin^2\theta_e + L_{sq}\cos^2\theta_e \tag{23}$$

$$L'_{sm} = \left(L_{sq} - L_{sd}\right)\sin\theta_e\cos\theta_e \tag{24}$$

The rotor equations are very similar to the three-phase machine equations, and the techniques for rotor flux oriented control may be easily adapted. However, the single-phase machine stator equations are more complex than those for the three-phase machine, due to the existing asymmetry. Even so, it is possible to obtain simple high performance control schemes based on the stator equations, as it will be shown in the following section.

III. CONTROL SCHEMES

A. Indirect field oriented control

A rotor flux oriented indirect control scheme was already proposed [2]. In a rotor flux reference frame, (12), (13) and (18) lead to:

$$\tau_r \dot{\lambda}_r + \lambda_r = L_m i_{sd} \tag{25}$$

$$T_e = \frac{P}{2}\frac{L_m}{L_r}\lambda_r i_{sq} \tag{26}$$

$$\theta_e = \theta_r + \int \frac{L_m}{\tau_r}\frac{i_{sq}}{\lambda_r} dt \tag{27}$$

Then, in a rotor flux reference frame, i_{sd} can be used to control the rotor flux magnitude (independently from the electromagnetic torque) and i_{sq} can be used to control the torque. However, it should be observed that the technique requires the rotor angular position (or speed) for the flux and torque control.

B. Hysteresis Direct Torque Control

Conventional direct torque control strategies for three-phase induction machines are based on the idea of accelerating stator flux vector to increase torque [4], [6] and [7]. Considering that the squirrel-cage rotor of a single-phase machine is similar to that of a three-phase machine, accelerating the stator flux vector in a single-phase machine would have the same effect of increasing the torque.

In a stationary reference frame, the stator voltage equation is:

$$\frac{d}{dt}\vec{\lambda}_s = \vec{v}_s - R_s \vec{i}_s \tag{28}$$

If the resistive voltage drop $R_s \vec{i}_s$ is small, as compared to one of the active voltage vectors available in the inverter output, then the stator flux variation is imposed by the voltage vector. Considering the four-switch inverter presented in Fig. 1, there are four possible voltage vectors, as shown in Fig. 2. Extending the idea used in [4], if the dq plane is divided into four sectors as shown in Fig.2, then the choice of the voltage vector to be applied for flux and torque correction would be done according to Table I, where i is the sector in which the stator flux vector is located.

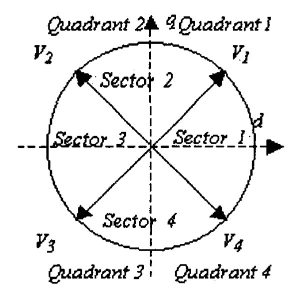

Fig. 2. Voltage vectors in single-phase inverter output.

TABLE I

VOLTAGE VECTOR SELECTION FOR A DTC SCHEME SIMILAR TO THE PROPOSED IN [4]

Flux error ($\lambda_s^* - \lambda_s$)	Torque error ($T_e^* - T_e$)	Voltage vector chosen
> 0	> 0	V_i
> 0	< 0	V_{i-1}, if i > 1 V_4, if i = 1
< 0	> 0	V_{i+1}, if i < 4 V_1, if i = 4
< 0	< 0	V_{i+2}, if i < 3 V_{i-2}, if i > 2

However, when the stator flux vector is near the border of a sector, the voltage vector component in quadrature with the flux vector may be too small to give the desired result of reducing the torque error and the technique may fail. This problem does not exist in the three-phase DTC method. There are six active voltage vectors available from a three-phase inverter and the dq plane is divided into six sectors. This makes possible to choose a voltage vector for flux and torque correction that is at least 30 degrees apart from the stator flux vector.

A DTC scheme for single-phase induction motors without the described problem can be obtained if the dq plane sectors are defined as the four quadrants and some rules are imposed to give priority for the flux or torque control. These rules can be: a) if the magnitude of the flux error is bigger then some limit, one voltage vector is chosen which guarantees the flux error to be reduced; b) When the flux error is within the hysteresis band, the voltage vector choice must guarantee flux acceleration for torque error reduction. The voltage vector choice for this technique is presented in Table II, where i is the quadrant in which the stator flux vector is located.

Another way to overcome the problem is to define the sectors as in Fig. 2 and use switching Table I, except when the torque error magnitude increases during two consecutive sampling intervals. If this occur, the voltage vector selection is made from Table II.

Both alternatives were simulated and very similar results obtained. For this reason, only the results obtained by using the first alternative are presented.

If the scheme uses analog circuitry, a variable switching frequency will occur, which may be very high during some periods of torque control, since the time constant for torque variation is usually very small. In a microprocessor control with a constant sampling rate, the chosen voltage vector will be applied during one entire sampling interval. A high sampling rate may be necessary to avoid high torque and flux oscillations.

In the method proposed in the next section, a reference voltage vector is obtained from the torque and flux controllers and a PWM scheme is used to synthesize the reference voltages. Thus, mean voltage vectors with magnitudes smaller than the available from the inverter output can be produced and the flux and torque oscillations are expected to be minimized.

C. Field Oriented Direct Torque Control

In a stator flux oriented reference frame, the stator voltage equations become:

$$v_{sd} = R_{sd}' i_{sd} + R_{sm}' i_{sq} + \dot{\lambda}_s \tag{29}$$

$$v_{sq} = R_{sq}' i_{sq} + R_{sm}' i_{sd} + \omega_e \lambda_s \tag{30}$$

It is clear from the equations above that the d axis voltage component can be used to control the flux magnitude while, as the flux magnitude is maintained approximately constant, the q axis voltage component is to accelerate the flux and then control the torque. Fig. 3 illustrates the scheme. Feed-forward terms are added to compensate the coupling current terms present in (29) and (30).

The scheme is field oriented since it uses the dq model equations written in a stator flux oriented reference frame. But it is based on the principles of accelerating the flux vector to increase torque and does not need speed or position information for the torque control nor the use of current controllers. Thus, it can also be considered a DTC strategy.

TABLE II

ALTERNATIVE VOLTAGE VECTOR SELECTION SCHEME

Flux error ($\lambda_s^* - \lambda_s$)	Torque error ($T_e^* - T_e$)	Voltage vector Chosen
$\varepsilon_\lambda > h$	Don't care	V_i
$\varepsilon_\lambda < -h$	Don't care	V_{i+2}, if i < 3 V_{i-2}, if i > 2
$-h < \varepsilon_\lambda < h$	> 0	V_{i+1}, if i < 4 V_1, if i = 4
$-h < \varepsilon_\lambda < h$	< 0	V_{i-1}, if i > 1 V_4, if i = 1

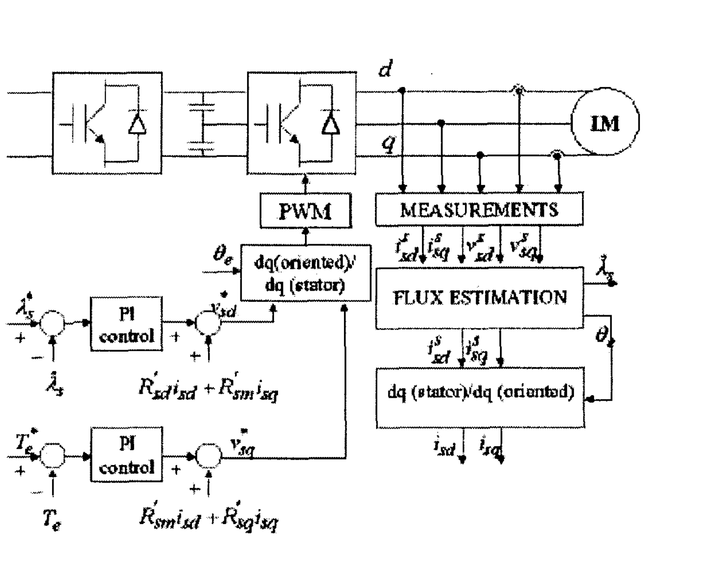

Fig. 3. Field oriented DTC scheme.

IV. FLUX ESTIMATION

As in conventional three-phase DTC strategies, stator flux can be calculated from stator voltage equation in the stationary reference frame:

$$\lambda_{sd} = \int (v_{sd} - R_s i_{sd}) dt \qquad (31)$$

$$\lambda_{sq} = \int (v_{sq} - R_s i_{sq}) dt \qquad (32)$$

The main advantage of this flux estimator is that the speed signal is not necessary. However, its use has some drawbacks due to instability problems caused by integrating offsets and drifts. Its performance can be improved, specially in low speeds if the pure integrator is substituted by a low-pass filter. This type of estimator was investigated in many papers for flux estimation in three-phase induction motors. Further investigations were not carried out in this paper.

V. INVERTER PWM METHOD

In the field oriented DTC strategy presented, some PWM strategy must be used in order to synthesize the required voltage vector. A scalar PWM strategy was used [2], due to its simplicity.

With the inverter topology presented in Fig. 1, the d axis stator voltage depends only on the state of switches S1 and S1_ of leg 1. Considering that the states of these switches are complementary, then the mean d axis stator voltage during one switching period is

$$\overline{V}_{sd} = \frac{E}{2}(2\tau_1 - 1), \qquad (33)$$

where τ_1 is the duty cycle of switch S1. Similarly, the q axis stator voltage mean value is

$$\overline{V}_{sq} = \frac{E}{2}(2\tau_2 - 1) \qquad (34)$$

Thus, in order to produce mean stator voltage components v_{sd}^* and v_{sq}^* the switches duty cycles are determined simply by

$$\tau_1 = \frac{1}{2} + \frac{v_{sd}^*}{E} \qquad (35)$$

$$\tau_2 = \frac{1}{2} + \frac{v_{sq}^*}{E} \qquad (36)$$

VI. SIMULATION RESULTS

In this section, some simulation results are presented in order to verify the proposed drives effectiveness.

Initially, responses to steps in the commanded stator flux magnitude and torque signals were obtained. Rated reference flux is maintained as commanded torques of 0N.m, 1N.m, -1N.m and 0.5 N.m at t = 0s, 0.2s, 0.4s and 0.6s, respectively, are applied.

Fig. 4 shows the commanded and actual values of speed, stator flux magnitude and torque for the hysteresis DTC scheme. A digital control scheme with a constant sampling rate was considered and the chosen voltage vector was then applied during the entire sampling interval. Flux and torque follow the commanded values fast and accurately but a high sampling rate (25kHz) was necessary to reduce torque oscillations. It can be seen that torque variations practically do not affect flux control. *d* and *q* axis current components (in a stator fixed reference frame) are shown in Fig. 5. High current values are initially imposed in order to produce the magnetic flux. After the flux reaches the reference value such high currents disappear and the current variations are associated to torque changes.

The same test was made with the field oriented DTC scheme. Fig. 6 shows the commanded and actual values of speed, stator flux magnitude and torque. As it can be seen, torque and flux oscillations are very reduced, despite the lower sampling rate used for the digital control (5kHz). The described scalar PWM method was used to produce the desired d and q axes voltages. The switching frequency was also 5kHz. In order to reduce the initial peak current, a saturation limit is imposed to the field oriented d axis voltage. As it can be seen in Fig. 7, the initial peak current is about one half of that obtained with the hysteresis DTC scheme. As it could be expected, the stator flux response is correspondingly slower. Fig. 8 shows that, as expected, d and q axis current components are strongly related to the flux and torque production, respectively.

Responses to steps in the commanded speed (considering that a speed or position sensor is available) are presented in Figs. 9 and 10. Fast and precise flux and torque responses show once again the scheme effectiveness.

Fig. 11 shows the responses of the field oriented DTC drive to steps in the commanded flux and torque without the use of the feedforward compensation terms of (29) and (30). indicating that the control performance would be almost the same.

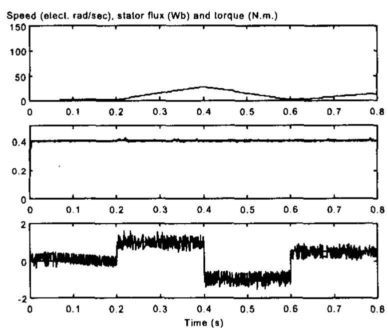

Fig. 4. Hysteresis DTC scheme. Speed (elect. rad/s), commanded and estimated stator flux magnitude (Wb) and commanded and actual electromagnetic torque (N.m)

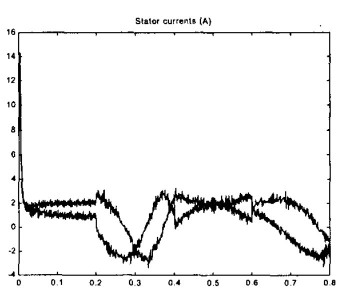

Fig. 5. Hysteresis DTC scheme. Stator currents i_{sd} and i_{sq} in a stator fixed reference frame.

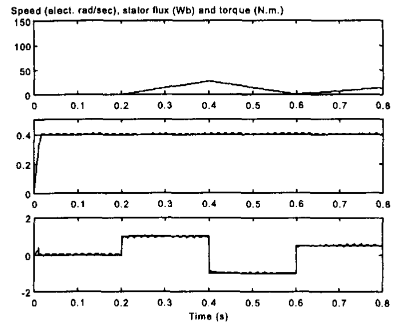

Fig. 6. Field oriented DTC scheme. Speed (elect. rad/s), commanded and estimated stator flux magnitude (Wb) and commanded and actual electromagnetic torque (N.m)

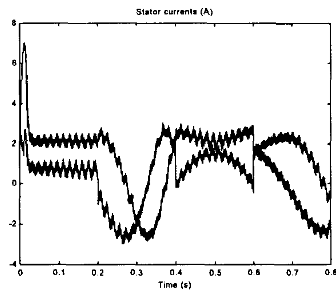

Fig. 7. Field oriented DTC scheme. Stator currents i_{sd} and i_{sq} in a stator fixed reference frame.

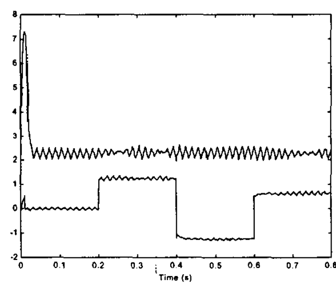

Fig. 8. Field oriented DTC scheme. Stator currents i_{sd} and i_{sq} in a stator flux oriented reference frame.

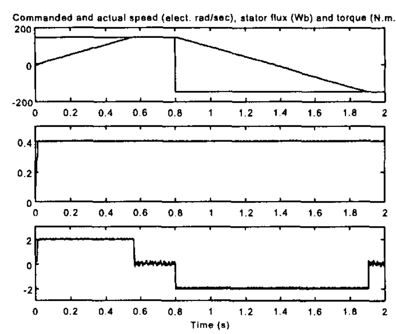

Fig. 9. Field oriented DTC scheme. Commanded and actual speed (elect. rad/s), commanded and estimated stator flux magnitude (Wb) and commanded and actual electromagnetic torque (N.m)

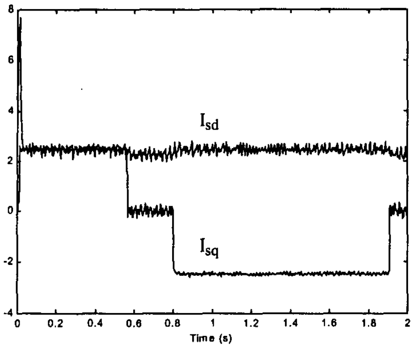

Fig. 10. Field oriented DTC scheme. Stator currents i_{sd} and i_{sq} in the stator flux oriented reference frame.

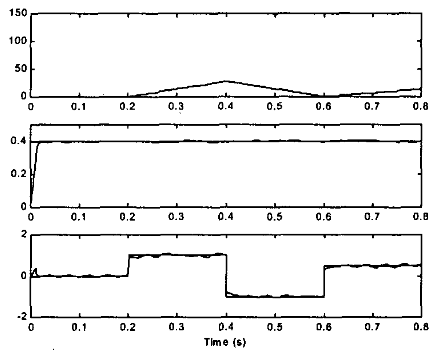

Fig. 11. Field oriented DTC scheme without the feedforward compensation for the coupling current terms in (29) and (30). Speed (elect. rad/s), commanded and estimated stator flux magnitude (Wb) and commanded and actual electromagnetic torque (N.m)

VII. CONCLUSION

In this paper two low-cost single-phase induction motor drive schemes are proposed and analyzed. The first method is based on the conventional DTC schemes for three-phase induction machines. The *dq* plane is divided into four sectors and a switching table is used for the choice of the voltage vector to be applied leading to flux and torque correction. However, there are some regions in the *dq* plane where the use of the switching table does not guarantee torque error reduction. The problem can be overcome by using an alternative division of the *dq* plane in order to allow an appropriate voltage vector selection that ensures torque error reduction whenever the flux error magnitude is smaller than some hysteresis limit. The simulation results obtained with this method show that fast and accurate torque and flux responses are obtained, but for its digital implementation without excessive torque oscillations, a high sampling rate is necessary.

A second DTC scheme is then presented, in which stator flux field oriented equations and PI controllers are used to determine the *d* and *q* axis voltage components for the flux magnitude control and its acceleration (torque control). A PWM method is used to synthesize the required voltage vector. The simulation results show that flux and torque follow their commanded values fast and precisely, despite the lower sampling rate used.

Both strategies proposed allow torque and flux control without needing speed or position measurement.

VIII. ACKNOWLEDGEMENT

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