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An Implementation of Direct Torque and Flux Control for Low Power Permanent Magnet Synchronous Motor Drives

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***Abstract*— This paper presents an implementation of Direct Torque and Flux Control (DTC) scheme for Permanent Magnet Synchronous Motor (PMSM) drive, focusing on the development of the digital control. The control was designed so that to allow accurate sampling related to control update time, considering in addition a measurement scheme common to the low power drives. Simulation and experimental results are presented to sustain the solution proposed for digital control implementation.**

***Index Terms*—direct torque and flux control, permanent magnet synchronous motor, digital control, implementation.**

I. Introduction

Direct Torque and Flux Control is a high dynamic drive control introduced back in mid-1980s. [1] Since then, this technique has been continuously in the researcher's attention [5], mainly due to its simple control scheme that remarkably uses voltage inverter switching states to achieve a very fast torque response and simultaneous flux control. The Swiss company ABB patented and introduced on the market the DTC as a high-performance alternative of Field Oriented Control (FOC).

The name DTC is derived from the fact that, on the basis of the errors between the reference and the estimated values of torque and flux, it is possible to directly control the inverter states in order to reduce the torque and flux errors within the prefixed band limits. Unlike Flux Oriented Control, DTC estimates and controls the torque and flux without needing coordinate transformations involving rotating coordinate frames or rotor position transducer.

However, DTC presents some disadvantages, such as high current and torque ripple, which were addressed in the literature [2], [8]-[11]. That is mainly due to the fact that the DTC exhibits large torque slopes. These are nevertheless advantageous from the dynamic response point of view, but care must be taken when implementing the digital controller, in order to limit the torque ripple as much as possible. The implementation presented in this paper proposes a solution which is meant to minimize the torque ripple by considering two important aspects: the control timing with respect to the stator currents sampling, and the derivation of the flux and torque discrete time equations, both aspects focusing on the idea of minimizing the unwanted delay between estimation and correction. The presented solution is based on a proper integration of the back-emf, which is strictly related to the mechanism of sampling, computation time and command update. In addition, there is emphasized the proper switching of transistors, in a widely used low power Permanent Magnet Synchronous Motor (PMSM) drive configuration, having current shunts placed on the lower side of the inverter.

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II. Principle of Direct Torque Control

The central idea of DTC is that the instantaneous values of torque and stator flux can be modified simultaneously by directly applying, for a short duration, one of the inverter voltage vectors, which has two effects on the stator flux vector: a fast rotation and a modification of the vector modulus. The rotation of the stator flux determines an angle increase between the stator and rotor flux, the latter being more sluggish (more filtered, due to the larger rotor electric time constant) [2]. Knowing that the torque is proportional to the angle between stator and rotor flux vectors, the resulted increase or decrease of the angle produces a torque variation, whose magnitude depends on the stator flux rotation speed. In the same time, as the effect of the same voltage vector, it is also modified the stator flux modulus.

The stator flux rotation and modification of its modulus are the result of a vector addition, as in the equation:

$$\underline{\Psi}_s^f = \underline{\Psi}_s^i + \Delta\underline{\psi}_s \tag{1}$$

where: $\underline{\Psi}_s^i$ and $\underline{\Psi}_s^f$ are respectively the initial and final stator flux vectors,

$\Delta\underline{\psi}_s \cong \underline{u}_s \Delta t$ is the stator flux variation, as a result of applying the voltage vector $\underline{u}_s$ for a duration Δt; this expression is the vector form of Faraday's Law, neglecting the ohmic drops on the winding resistance.

The control of the torque and flux is accomplished by choosing, every control loop, the inverter voltage vectors that will produce the desired variation direction (increase, decrease) of torque and flux, so that these quantities are kept within predefined bounds. For this purpose, there are utilized simple hysteresis comparators, which compare the desired (reference) torque and flux values with the calculated ones.

The selection of the proper voltage vector depends on the actual angle position of the flux vector, which is not determined by a precise angle in the stator orthogonal reference frame, but only by its location in one of six angular sectors, centered on the voltage vectors.

The model of the classical DTC PMSM motor scheme shown in figure 1, consists in torque and stator flux estimators, torque and flux hysteresis comparators, a switching table and a voltage source inverter (VSI) [6].

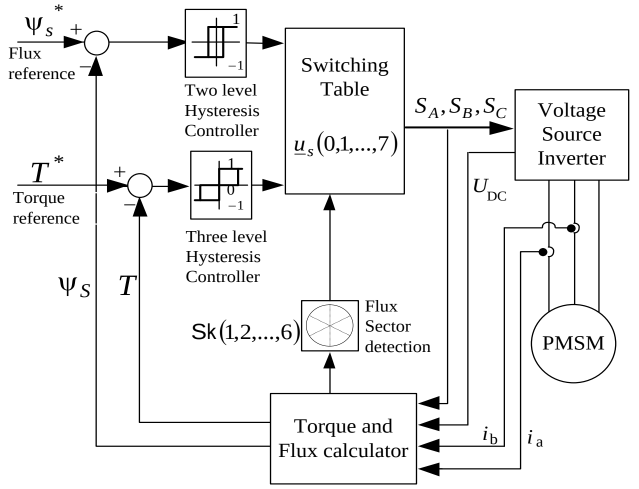

Fig. 1: Block diagram of the conventional DTC.

The outputs of flux and torque comparators indicate the variation sign (increase, decrease, keep unchanged) that must be induced to the flux and torque magnitudes, so that to keep them within the corresponding hysteresis bands.

The proper voltage vector selection is based on a voltage selection table (switching table), whose inputs are the desired torque and flux variations directions, and the sector in which the flux vector is situated [2].

III. Digital Implementation of Direct Torque and Flux Control for a Permanent Magnet Synchronous Motor

This paragraph presents aspects regarding the discrete-time implementation of DTC, in such a manner so that the computation time does not alter the promptness of the voltage command in response to the estimated magnetic flux and torque.

A. *Digitally implemented algorithm for the DTC*

The presented method is based on the calculation of the stator magnetic flux and torque using a linear approximation of the current variation over a switching period, in accordance with a proper synchronization of sampling and algorithm execution. The flux and torque are estimated in advance relative to the currents sampling moment, avoiding the delay which would be otherwise caused by the processing time.

Although the conventional DTC does not need Pulse Width Modulation (PWM), the present implementation uses this convenient feature in order to obtain hardware timing of sampling and command update. In the current PWM period, there are calculated the stator flux and the torque for the end of the same PWM period [5]. The stator flux is obtained by integrating the average value of the back-emf, using the currents sampled at the middle of the period, which approximate their average values. The torque is calculated using the stator flux and the extrapolated values of currents, to the end of the period. In this way, the voltage command for the next period will be prompt, and not altered by the execution time, like if the sampling and calculations were performed (almost) instantly, as in the case of a pure analogue controller.

The mechanism of coordinated sampling, processing and command update is illustrated in figure 3. The figure also contains modifications imposed by the two shunts current measurement configuration, very common in low power drives. The modification for this case is explained below, considering a VSI feeding three-phase Y-connected isolated neutral windings of a PMSM.

In the majority of low power drives, the motor currents are measured using two shunts placed on the low side of the VSI, as shown in figure 2. This simple and cost-effective measurement scheme suffices, since the 3rd current is obtained from the other two.

However, in the case DTC, which maintains the voltage phasor unchanged over the duration between two updates, there is a situation where current values cannot be measured. This is encountered when the current flows from a high-side transistor (dotted trace in figure 2), having no shunt in series, and returns through two low-side transistors, only one of which having a shunt in series. In this case, a single current is measured, insufficient for three-phase motor windings.

Therefore, in order to be able to measure two currents, the sampling has to occur when the line current passes the low side of the inverter leg (either through the transistor or the diode, depending on the direction of the current). The solution used to measure the current is a short insertion of a null voltage value to the motor terminal, by turning on all the low side transistors. During this short null voltage, the current flows through the freewheeling diode and the shunt (the continuous trace in figure 2.

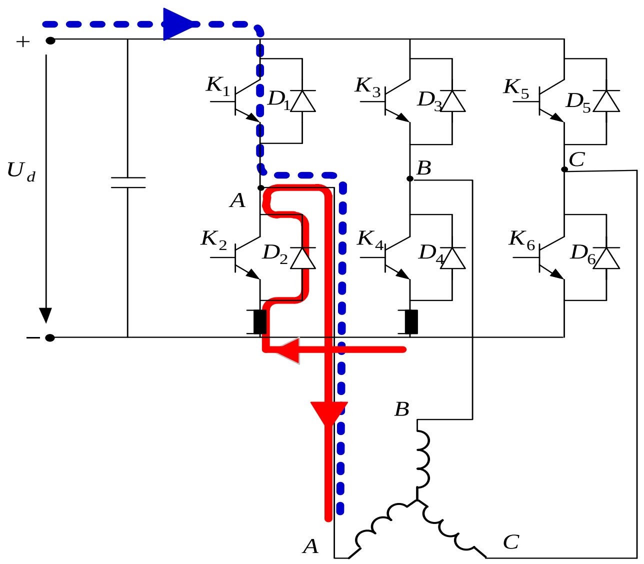

Fig. 2: Current path through the high-side of the inverter leg, as imposed by the voltage command (dotted trace) and the path forced through the low side with the shunt, according to the switching modification.

In order to insert the null voltage, it was used PWM, with a duty cycle less than 1. The correspondence between the switching functions (S_A , S_B , S_C) that define the voltage phasors and the width of the command impulses for the transistors is as follows:

- For $S_X = 1$ the duty cycle is $\delta < 1$;
- For $S_X = 0$, the duty cycle equals 0.

where X = A, B or C.

The duration for which the null voltage is applied has been selected as the minimum conduction time required for the inverter's transistor, which is 4 μs is this case.

The components of the voltage phasor during a PWM period are determined from the equations:

$$\begin{aligned} u_{s\alpha}(k) &= \frac{2}{3}\delta U_d(k)\left(S_A(k) - \frac{S_B(k)}{2} - \frac{S_C(k)}{2}\right) \\ u_{s\beta}(k) &= \delta U_d(k)\left(\frac{S_B(k) - S_C(k)}{\sqrt{3}}\right) \end{aligned} \qquad (2)$$

where:

- $U_d(k)$ is the DC bus voltage,
- δ is the duty cycle,
- $S_A(k)$, $S_B(k)$, $S_C(k)$ are the switching functions for the current PWM period.

Considering the null voltage insertion, the current sampling takes place at the beginning of the PWM period, immediately followed by the control loop execution, and the command update is done at the half of the period, as presented in figure 3.

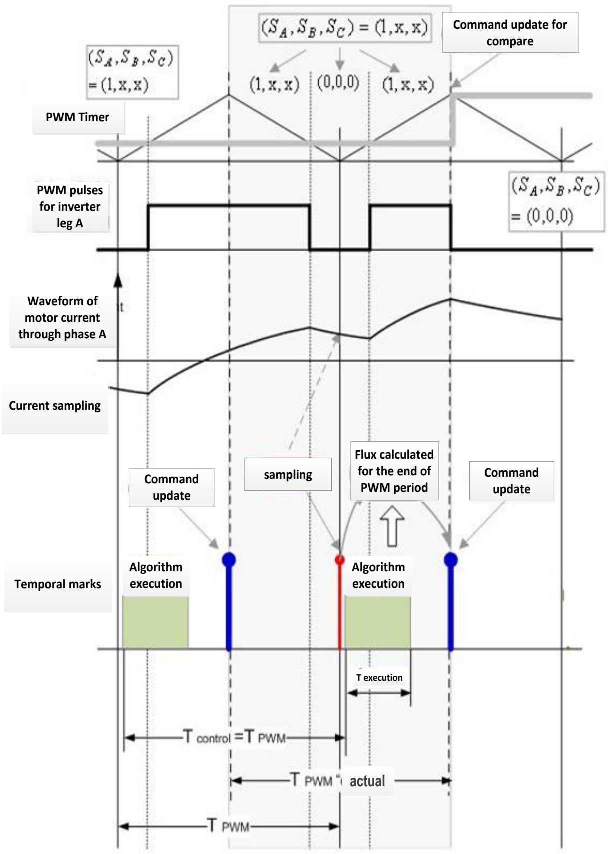

Fig. 3: Correlated sampling, processing and voltage command update.

The instantaneous torque equation is written with the flux and current components:

$$m(k) = \frac{3}{2}p\left(\psi_{s\alpha}(k)\cdot i_{s\beta}(k) - \psi_{s\beta}(k)\cdot i_{s\alpha}(k)\right) \qquad (3)$$

In practice, it is desirable to use a low pass filter instead of an integrator, in order to avoid the build-up due to the inherent offset of the measured currents. A low pass filter having the transfer function, described by equation 4:

$$H(s) = \frac{1}{s + 2\pi \cdot f_c}, \qquad (4)$$

approximates the integrator at frequencies beyond cut-off frequency [3]. By using the z-transform of the above filter instead of integrator, the difference equations of the stator flux components become:

$$\begin{cases} \psi_{s\alpha}(k) = \dfrac{1}{1 + T\cdot 2\pi \cdot f_c}\left[\psi_{s\alpha}(k-1) + T\cdot\left(u_{s\alpha}(k) - R_s \cdot i_{s\alpha}(k)\right)\right] \\ \psi_{s\beta}(k) = \dfrac{1}{1 + T\cdot 2\pi \cdot f_c}\left[\psi_{s\beta}(k-1) + T\cdot\left(u_{s\beta}(k) - R_s \cdot i_{s\beta}(k)\right)\right] \end{cases} \qquad (5)$$

The stator frame α axis overlaps the A axis and the β axis is in quadrature, counterclockwise.

B. Simulation results of the DTC command in Matlab-Simulink for the Permanent Magnets Synchronous Motor

The presented control algorithm was firstly tested by simulation, in Matlab-Simulink. It was considered for the simulation the parameters of the Pittman3441 PMSM, which was used for experiments, having the following data:

- Rated voltage 19.1 V,
- Rated current 1.16 A,
- Rated torque 29 mNm,
- Stator phase resistance 2.625 Ω ,
- Stator phase inductance 0.23 mH,
- Amplitude of the line e.m.f. at 1000 rpm (flux constant): 2.63 V/1000 rpm,
- Torque constant 24.9 mNm/A,
- Two pairs of poles (p=2).

The magnetic flux was determined using the following formula:

$$\Psi_m = \frac{60}{2\pi \cdot p \cdot 1000 \cdot \sqrt{3}} E = 7.2 \cdot 10^{-3} \text{ Wb} \qquad (6)$$

where $E = 2.63$ V/1000 rpm represents the peak value of the e.m.f., which is a constant in the motor datasheet, and p is the number of pairs of poles of the stator windings.

Initially the rotor aligns with the $\alpha \equiv A$ axis, thus the initial flux components are:

$$\begin{aligned} \Psi_{s\alpha}(0) &= \Psi_m \\ \Psi_{s\beta}(0) &= 0 \end{aligned} \qquad (7)$$

The initial values of the stator flux components were determined based on the flux induced by a stator phase and the initial position of the rotor.

The control parameters were the following:

- Control frequency of flux and torque of the DTC algorithm:

$f_{ctrl} = 20KHz$ (Time period $= 50\mu s$),

- The motor speed sampling time = 1 KHz.

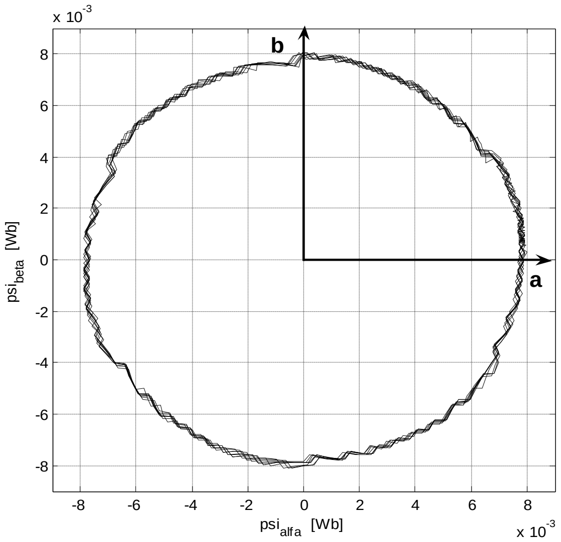

Fig. 4: The locus of the peak stator flux phasor in the stator coordinate system.

The simulation results in figure 4 show the stator flux locus in the stator frame, and figures 5 and 6 represent the variations in time of the following parameters:

1. Stator flux components in the stator reference system (α, β), Psi_alfa and Psi_beta $[Wb]$,
2. Phase currents, i_a and i_b [A],
3. Rotor speed [rpm].

Figure 5 presents the variation of above quantities during motor start-up and steady state regimes, whereas figure 6 includes in addition a motor reversal.

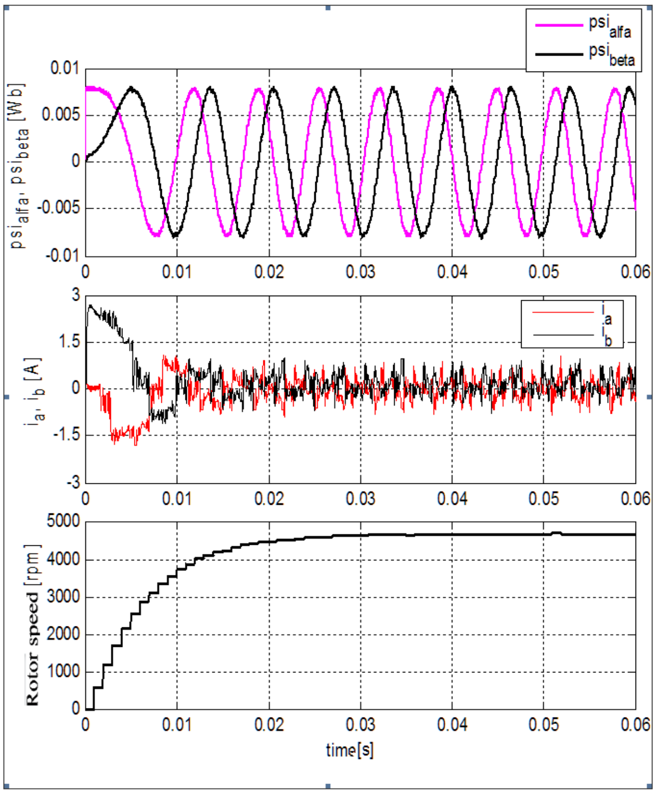

Fig. 5: Stator flux components, phase current and the rotor speed during start up and stabilized regime.

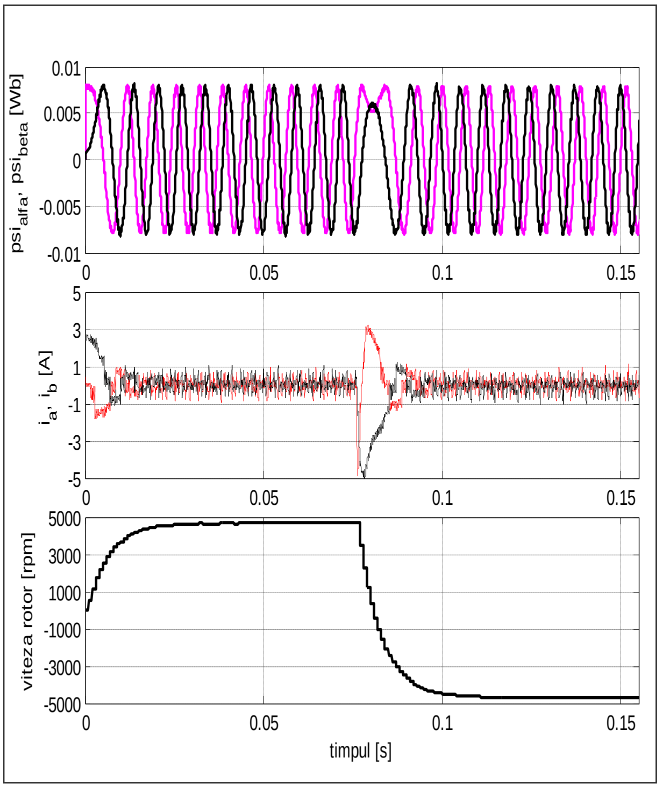

Fig. 6: Stator flux components, phase current and the speed at start up and during reversal.

C. Experimental results

The presented algorithm was experimentally tested using the MCK2812 development kit from Technosoft Company, shown in figure 7, which includes:

- The Digital Motion Control Development Pro v3.0 development environment, which also allows the acquisition and graphical representation of variables,
- The control board, based on the Texas Instruments TMS320F2812 fixed point processor, clocked at 150 MHz,
- The PM50 power module. It is a tri-phase inverter IGBTs bridge, with 100 KHz maximum switching frequency.
- Tri-phase PMSM, Pittman3441,
- Incremental encoder with 500 lines per rotation.

Fig. 7: The Technosoft MCK2812 kit.

The control parameters are:

- Torque and flux control frequency of the *DTC* algorithm, $f_{ctrl} = 20KHz$;

- PWM frequency: $f_{PWM} = 20KHz = f_{ctrl}$.

The torque reference was applied directly, without a speed regulator. Motor speed was sampled with 1 KHz frequency.

The geometric locus of the stator flux phasor amplitude is illustrated in figure 8.

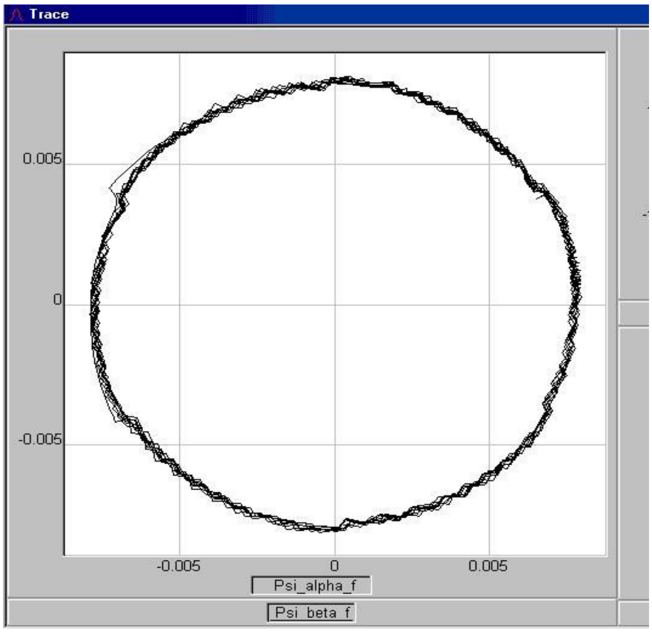

Fig. 8: Experimental results. The geometric locus of the stator flux phasor amplitude within the stator reference frame.

Figures 9 and 10 show the variations in time of the following parameters:

- Stator flux components in the stator reference system (α, β), Psi_alfa and Psi_beta $[Wb]$,
- Phase currents, i_a and i_b [A],
- Rotor speed [rpm].

Figure 9 presents the variation of above quantities during motor start-up and steady state regimes, whereas figure 10 includes in addition a motor reversal.

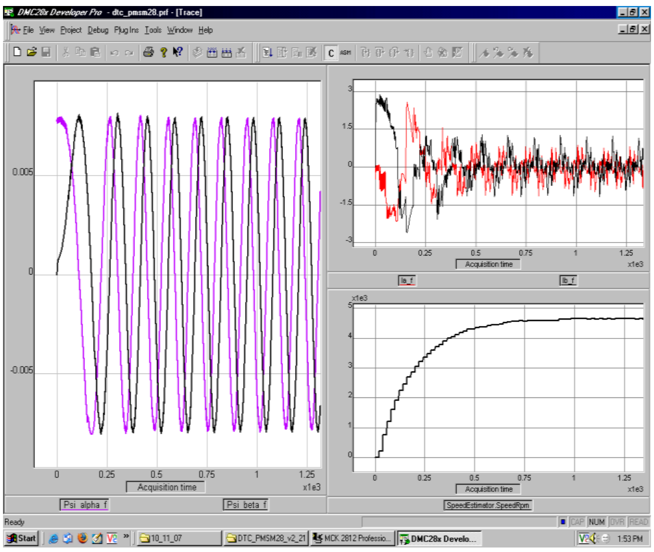

Fig. 9: Experimental results: time variation of stator flux components, phase currents and speed in start-up and steady state regimes.

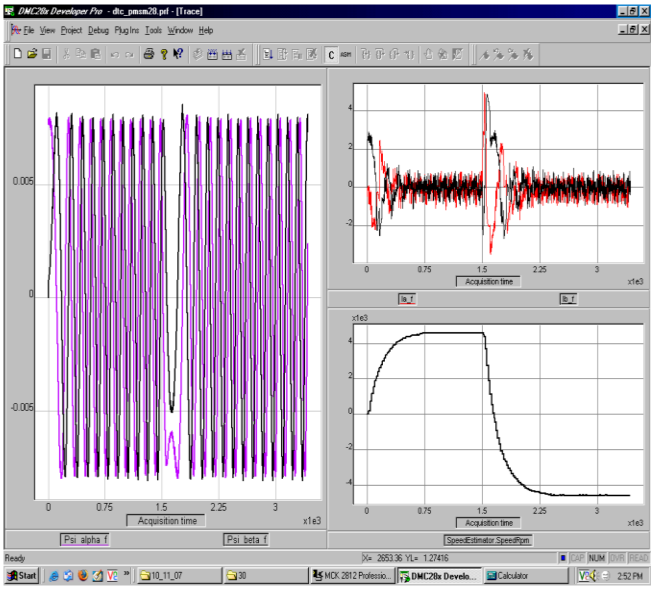

Fig. 10: Experimental results: time variation of stator flux components, phase currents and speed during start-up, steady state and motor reversal.

Time in the graphs is expressed as samples, which represent the number of DTC control periods (1000 samples represent 50 ms).

Even with large current ripple test results show that flow is correctly estimated, with no oscillations of the spatial phasor amplitude.

The large current ripple observed is due to a very small stator time constant when associating the DTC voltage control method with a low power motor. Specifically, DTC control uses only one voltage phasor on a control period comparable with the electric time constant of the motor. Even more so, current spikes are amplified by sector transitions, also specific to DTC control.

An improvement in this area could be done by increasing the control frequency, possible in this implementation because of the low execution time (5.9 µs. maximum) of the DTC control loop.

By comparing the simulation and experimental results it can be noticed a fair degree of similarity. The experimental results are close to the simulations results, showing the accuracy of the control implementation solution.

IV. Conclusion

This paper presented a method of digital implementation of Direct Torque Control applied for PMSM, whose aim is to eliminate the negative influence of the computation time on the torque and flux responses. By this method, the torque and flux are estimated in advance. Since it involves strict synchronization of sampling, execution and command update, it was built a control model that allows thorough analysis of the waveforms, as well as the test of the algorithm.

The simulation and experimental results proved the efficacy of the proposed solution. Results illustrate that the developed algorithm is correct, even in the control of a low power PMSM, having a very low stator time constant.

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