A DSP IMPLEMENTATION OF AN ALTERNATIVE METHOD OF VECTOR CONTROL FOR INDUCTION MOTORS

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Abstract – **The paper presents a DSP implementation of an alternative vector control system for induction motors, which does not make use of the absolute position of any flux linkage (rotor, stator nor airgap flux linkage). The paper presents an overview of the alternative method of vector control in its hybrid (digital plus analogue) version and then the completely digital version that uses a DSP system. Some results from experimental tests are presented.**

KEYWORDS

Vector control, induction motor, DSP implementation, alternative method

I. INTRODUCTION

The vector control system theory was published at the beginning of the 70´s decade by K. Hasse [1] and F. Blaske [2], but it became widely known after the publication of the paper [3], in 1980, by R. Gabriel, W. Leonard and C. J. Nordby, which presents an implementation using the 8085 microprocessor launched in 1976.

The aim of a vector control system applied to an induction motor is to achieve a dynamic performance similar to that of the separately excited dc machine. Due to its physical characteristics the dc machine can be operated under a decoupled condition, where fast torque response is achieved by controlling the armature current, while keeping constant the field current.

Several vector control implementations have been presented in the last three decades and they can be divided into two groups: direct and indirect vector control methods. In the direct method the position of the flux linkage (rotor, airgap or stator flux linkage) is obtained from flux signal sensed by some devices inserted in the induction motor (for instance Hall effect sensor and search coils) or by estimation from terminal quantities. In the indirect method the flux position is obtained by integrating the stator frequency, which is obtained from the speed signal plus the estimated slip frequency. The slip frequency ω_{sl} is given by expression (1), as a result of the decoupled model of the induction motor [3]. The variables i_{ds} and i_{qs} are respectively the flux and the torque components of the stator current. τ_r is the rotor time constant.

$$\omega_{sl} = \frac{1}{\tau_r}\frac{i_{qs}}{i_{ds}} \tag{1}$$

The flux position is an essential peace of information to transform the variables i_{ds} and i_{qs} from the synchronously rotating reference frame into the variables i_a^*, i_b^* and i_c^* in the stationary reference frame. Expression (2) is used for this transformation.

$$\begin{bmatrix} i_d^s \\ i_q^s \end{bmatrix} = \begin{bmatrix} \cos\theta_s & -\operatorname{sen}\theta_s \\ \operatorname{sen}\theta_s & \cos\theta_s \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix}$$

$$\begin{bmatrix} i_a^* \\ i_b^* \\ i_c^* \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -1/2 & \sqrt{3}/2 \\ -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_d^s \\ i_q^s \end{bmatrix} \tag{2}$$

The variables i_a^*, i_b^* and i_c^* are the stator reference currents and are used for controlling the power converter that drives the induction motor.

In contrast to traditional systems, another vector control scheme is presented in [4] and [5], which does not need any flux position information. This method is called here by alternative vector control method.

II. THE ALTERNATIVE VECTOR CONTROL

It is shown in [4] that, even without the position of the rotor flux-linkage, the operating point of an induction motor can be changed in a jump from one steady state to another without any undesirable oscillations in the torque profile. In that approach two previously known operating points are used and the induction motor operates at zero speed. During the experimental tests the load condition is changed in a jump simultaneously with the magnitude (I_s), the frequency (ω_s) and the difference in phase (δ) of the stator current. Fig. 1 shows the unit that generates the stator reference currents from sinusoidal waves stored in three EPROMs.

The assembly for that experimental implementation consisted in load applied to the motor axis through a torque arm. Some weights were attached to the end of the torque arm and the induction motor was powered through an inverter drive so as to keep the torque arm in equilibrium in horizontal position. The rotor was kept at zero speed, but not locked. The experiment was performed reducing the load torque in a step by releasing some of the weights. At the same time the magnitude, frequency and phase for the new load condition were applied to the motor. The result was a transient-free change of the motor torque.

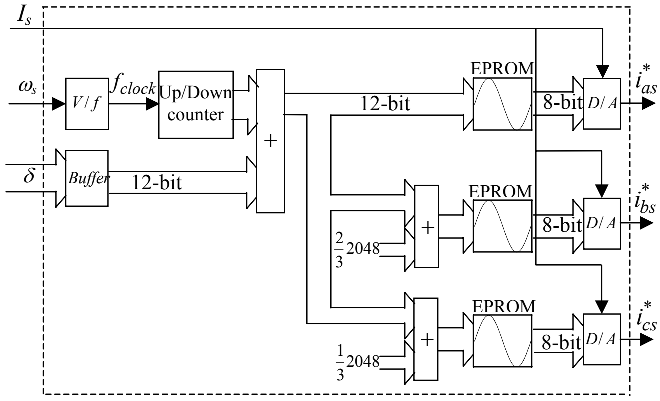

Fig. 1 – Reference Current Generating Unit

The difference in phase is given by (3) [4]:

$$\delta = \xi_2 - \xi_1 = arc\tan(\omega_{sl2}\tau_r) - arc\tan(\omega_{sl1}\tau_r) \quad (3)$$

The variables ξ_1 and ξ_2 are respectively the phases for the operating points 1 and 2 (Fig. 2). The phase represents the relative position between the stator current vector ($\vec{i}_s$) and the vector of the rotor flux-linkage ($\vec{\psi}_r$). The rotor flux-linkage triangle depicted in Fig. 2 is used in [4] to explain the vector control method. It is shown that the orthogonality between the rotor flux-linkage vector and the rotor current vector is kept not only under a steady state condition but also throughout the operating condition change.

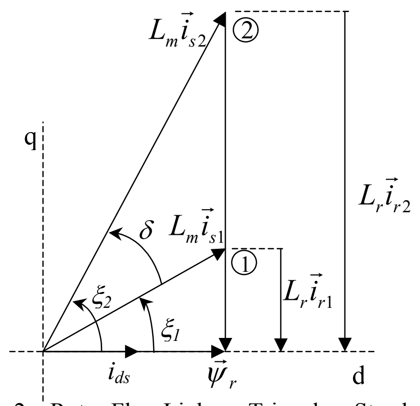

Fig. 2 – Rotor Flux-Linkage Triangle – Steady State

The result from the experimental implementation shows that the difference in phase is the key feature for this implementation, instead of the absolute position of the rotor flux-linkage. It is seen that the motor is operating under steady state condition and the absolute position is not measured or estimated at any time. The phases ξ_1 and ξ_2 for the operating points 1 and 2 are used just to calculate the difference in phase δ, which is the variable that is effectively used for the transient-free jump. It is important to stress that a jump in the phase difference means a jump in the reading address of the EPROM. Step change to a higher value of load torque results in a forward jump in the reading address of the EPROM, while a step change to a lower value results in a backward jump.

III. THE HYBRID VERSION

The system described in the last section was implemented for operating at zero speed and it needed a previous knowledge of the load conditions for both operating points. It consisted in an open loop system. Fig. 3 illustrates the system implemented in [5] for closed-loop operation.

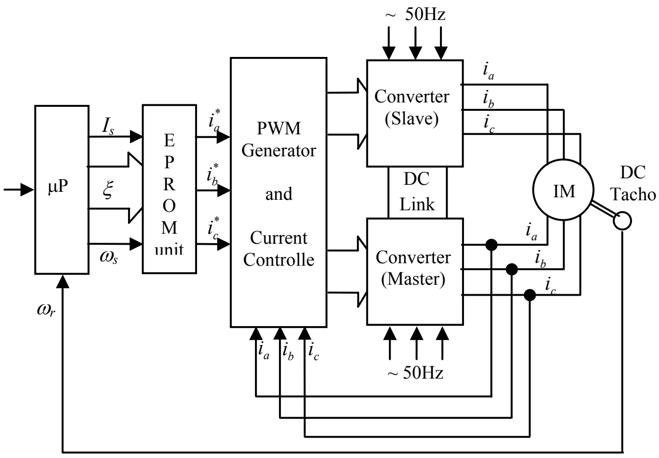

Fig. 3 – Hybrid System

The unit for generating the reference for the stator currents is the same as before, but a microcomputer is used for calculating the magnitude, frequency and phase (instead of phase difference) of the stator current, in accordance with expressions (1), (4) and (5).

$$I_s = \sqrt{i_{ds}^2 + i_{qs}^2} \quad (4)$$

$$\xi = \arctan\left(\frac{i_{qs}}{i_{ds}}\right) \quad (5)$$

The speed signal (ω_r) is obtained through a dc tachometer and it is fed back to the microcomputer unit. Added to the slip frequency it results in the stator frequency (ω_s). The calculated values of the stator current magnitude, frequency and phase are used to generate the references for the stator currents through the reference current generating unit (Fig. 1). The result is the same as if the expression (6) were used, instead. $\theta_s = \int \omega_s dt$.

$$\begin{aligned} i^*_{as} &= I_s \cos(\theta_s + \xi) \\ i^*_{bs} &= I_s \cos(\theta_s + \xi - 2\pi/3) \\ i^*_{cs} &= I_s \cos(\theta_s + \xi + 2\pi/3) \end{aligned} \quad (6)$$

The hybrid system uses two PWM converters so as to allow independent control over each phase current. The experimental results [6] show a high dynamic response to speed variation (such as that depicted in Fig. 4) and load change. The experimental test with load change show some results for operation at different speeds, including zero speed.

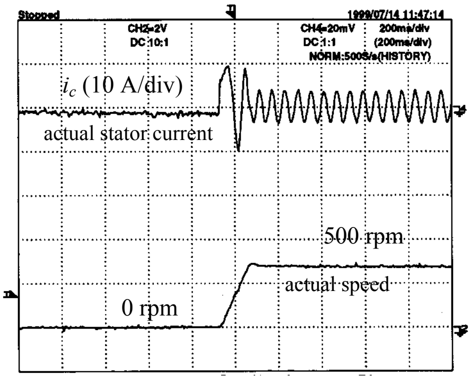

Fig. 4 – Accelerating process from 0 to 500 rpm – hybrid system

The torque response follows the expression (7), where the rotor flux linkage is kept constant, while the torque component of the stator current is changed.

$$T_e = k \cdot \psi_r \cdot i_{qs} \tag{7}$$

IV. DSP IMPLEMENTATION

Fig. 5 illustrates the DSP implementation of the alternative vector control system. It shows that only one PWM converter is used and that the DSP unit replaces the microcomputer, the EPROM unit, the current controller and the PWM generating unit of the hybrid system.

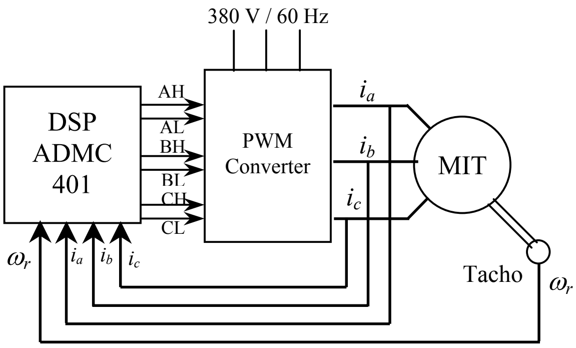

Fig. 5 – DSP Implemented System

The DSP ADMC401 [7] by Analog Devices is used in this work. Some of its features are: 26 MIPS (35 ns instruction cycle), PWM outputs, 12-bit analogue to digital converters and 12-bit digital to analogue converters. The algorithm used in this vector control system was developed using the DSP assembly language. It also makes use of some application notes related to mathematical and trigonometric routines.

Fig. 6 shows the program sequence, which was divided into 5 parts: data acquisition unit, magnitude, frequency and phase processing unit, reference current generating unit, current controller unit and PWM generating unit.

The data acquisition unit reads the actual values of the three currents and of the speed signal, it makes the necessary signal conditioning and calculates the torque component of the stator current i_{qs} through a PI controller. In the next stage the default value of i_{ds} is used together with i_{qs} to calculate the magnitude I_s, the phase ξ and frequency ω_s ($\omega_r + \omega_{sl}$). These two previous stages are performed by the microcomputer in the hybrid version. The reference current generating unit and the current controller are then used to generate the signal that is necessary to update the PWM output.

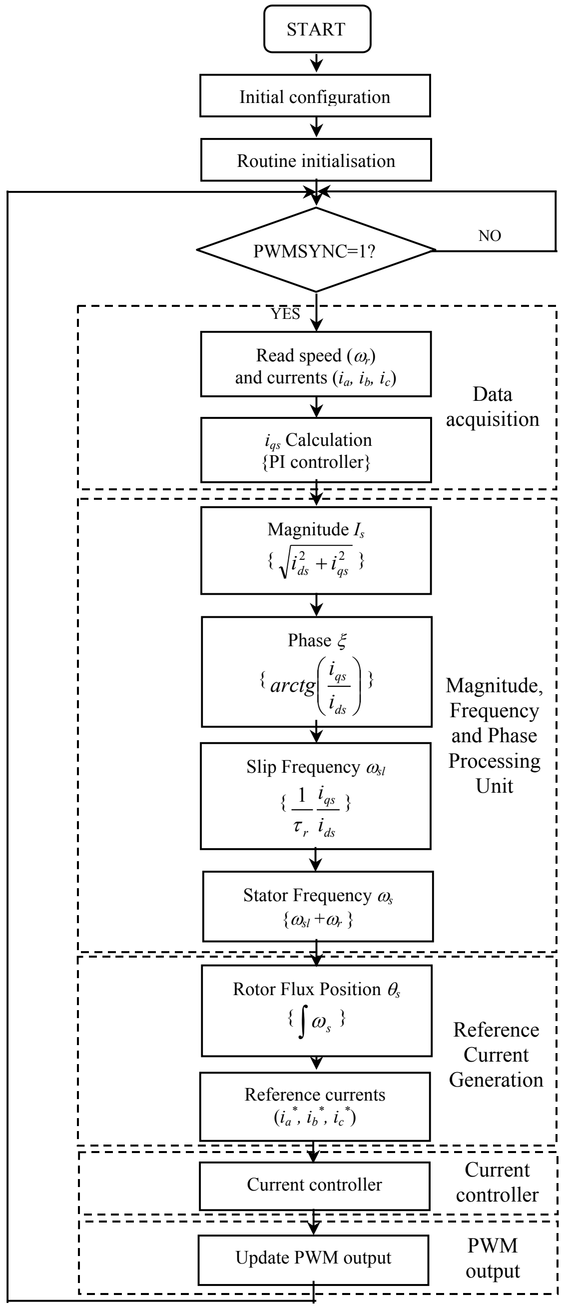

Fig. 6 – Processing sequence

On the hybrid version the current generation makes use of sinusoidal waves stored in EPROMs, as shown in Fig. 1. The DSP version could use a loop-up table for this purpose however it is used a trigonometric function in this first DSP implementation. A look-up table is to be used in the next stage of the research, which might result in a shorter processing time. The total processing time for the present version is shown in Fig. 7. It is 23.6 μs, against 47 μs of the hybrid version. The chosen PWM frequency is 20 kHz.

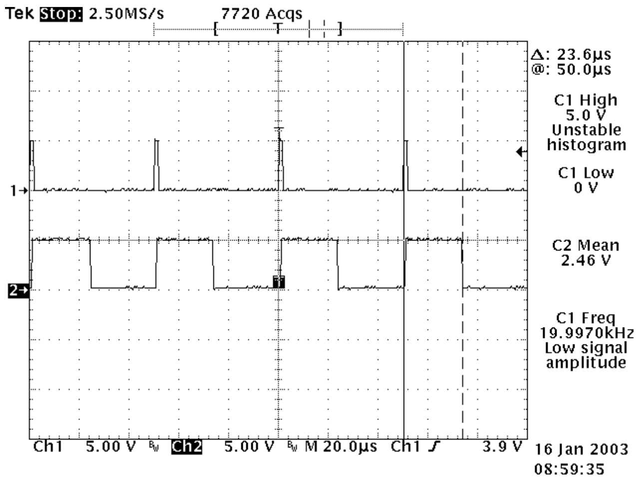

Fig. 7 – Processing time: PWMSYNC signal (upper trace) and processing time (lower trace)

Some of the first experimental results are presented here to illustrate the dynamic performance of the implemented vector control system. Fig. 8 shows the accelerating process from zero to 600 rpm. The middle trace is the actual speed and the lower trace is the torque component of the stator current. The speed profile shows a fast response when a change in speed is called for and it also shows no oscillation in the end of the accelerating process.

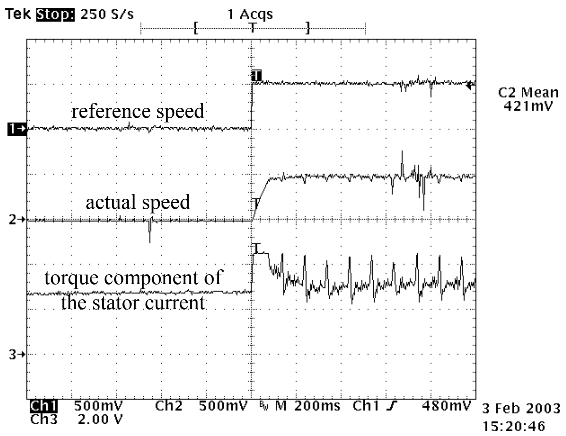

Fig. 8 – Accelerating process from zero to 600 rpm

The torque component profile shows a rapid change at the beginning of the accelerating process and a constant value throughout the accelerating period, as expected from a vector control system. However, at 600 rpm the torque component profile presents some peak values at a constant rate, which were not expected. They are credited to some noise signal coming from the tachometer and the tuning of the current controller.

Fig. 9 shows the decelerating process from 600 rpm to zero. The actual speed (middle trace) shows a linear decrease in speed (constant torque) and no oscillations at the end of the stopping process, as expected from a vector control system.

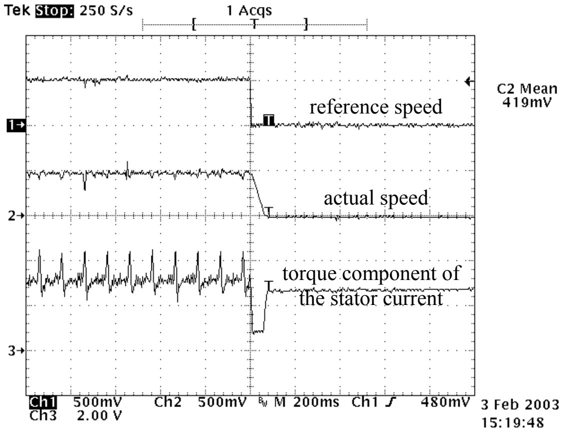

Fig. 9 – Decelerating process from 600 rpm to zero.

The torque component of the stator current shows the same problem as before, at 600 rpm, however it shows a rapid response when the change in speed is called for, it stays constant during the decelerating period and it presents no oscillations when the speed reaches zero.

V. CONCLUSION

The DSP version of the alternative vector control system presented here has not reached yet its ideal adjusts however the first results indicate a dynamic response as expected from a vector control system. The noise signal is expected to be minimised with a better tuning of the current and speed controllers. Despite the experimental problems faced by the presented DSP implementation, there is no doubt about it´s advantages over the hybrid version, such as versatility and dimensions. A sensorless implementation of the vector control system presented here has been under study..

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