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Position Sensorless Synchronous Reluctance Motor Drive using the Stator Phase Voltage Third Harmonic

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Abstract A direct field orientation controller for synchronous reluctance motors (SynRMs) based on determination of the spatial position of the air gap flux from the third harmonic component of the stator phase voltages is presented. The control utilizes the spatial saturation harmonic components rotating at synchronous frequency that are generated in the air gap flux when the machine operates in a saturated condition. When the machine is wye connected, the sum of the three phase voltages results in a signal dominated by the third harmonic component. Using this signal, a completely sensorless drive is implemented which can successfully synchronously start the reluctance motor from zero speed. Extensive experimental results showing the effect of different estimation schemes and practical problems of detecting the flux position from the harmonic voltage signal are presented.

1. INTRODUCTION

The potential of the synchronous reluctance motor (SynRM) has been studied intensely over the past several years, and an increased interest in its capabilities and performance has been clearly demonstrated. Since the converter-fed SynRM does not need a starting cage, an optimized rotor for synchronous performance must be designed. The motor is made auto synchronous by the use of electronic control, and optimal operation can be implemented for all the functioning regimes. Motors with high ratios of X_d/X_q have been designed, thus significantly improving the power density of the machine. The simple and robust construction, reduced rotor losses, etc., are other significant advantages of the SynRM. A drawback of these drives resides in the need of synchronization of the power supply with the rotor position. Thus, a rotor position transducer is usually used, resulting in an increase of the cost and a decrease of the reliability of the drive.

Recently, the important benefits of measuring the air gap flux linkage via the third harmonic voltage component of the stator phase voltages have been demonstrated. A discussion of ac machines operation under saturation and the consequent generation of harmonic components in the air gap flux has been presented in [1-2] and has been utilized to correct the time constant model in an indirect field oriented control scheme for the induction machine [3-4]. Control principles that use the third harmonic voltage to sense the air gap flux position and thus realize direct field orientation, are also proposed in [5-6] for the induction machine. These controllers require only access to the neutral point of the stator windings to realize measurement of the phase voltage. Hence, a low cost medium performance controller can be realized without the need for explicit field position measurement in the torque loop. A similar approach is used in the present paper for the SynRM, eliminating the need for the speed and position transducers in order to implement the control of this machine.

2. CONTROL OF SYNCHRONOUS RELUCTANCE MOTOR

2.1. The SynRM model

The SynRM has stator windings similar to those of polyphase ac machines, and a rotor designed with appreciably different values of the reluctance on the two axes, perpendicular on each other (d and q axes). High performance SynRM uses axially laminated rotor, thus high ratios of X_d/X_q are obtained. A model of the SynRM that includes core losses may be found in [7]. A resistor R_m, coupled to the total flux, is added to incorporate the iron losses of the machine. The vector diagram for the motor is presented in Figure 1. The effect of considering the iron losses is the occurrence of a phase difference ($\delta\theta$) between the stator current (i_{qds}) and MMF (i_{qdm}).

As one can see from the Figure 1, three different reference frames may be considered in order to implement the machine control:

- a fixed stator reference frame (ds,qs axes);
- a rotor reference frame (dr, qr axes);
- a field reference frame (f, t axes).

This fact has several consequences on the overall control performances of the machine. The proposed model yields useful analysis of motor performances, and offer control strategies to compensate the effect of current vector displacement due to iron losses, as well as the fact that the current component responsible for torque production (i_{dqm}) is not directly measurable in this case [7].

2.2. Control strategies for the SynRM

A comparative study of different control methods of the SynRM is done in [8]. The performances of the motor (ideal model) are analyzed, for four different control techniques. The ratio $\zeta = L_d/L_q$ has important influence on the machine behavior. An important study parameter is also the current angle with respect to the rotor axis, γ. The control strategies that are compared are:

(a) maximum torque control (maximum torque/ampere), obtained by setting $\gamma = \pi/4$;

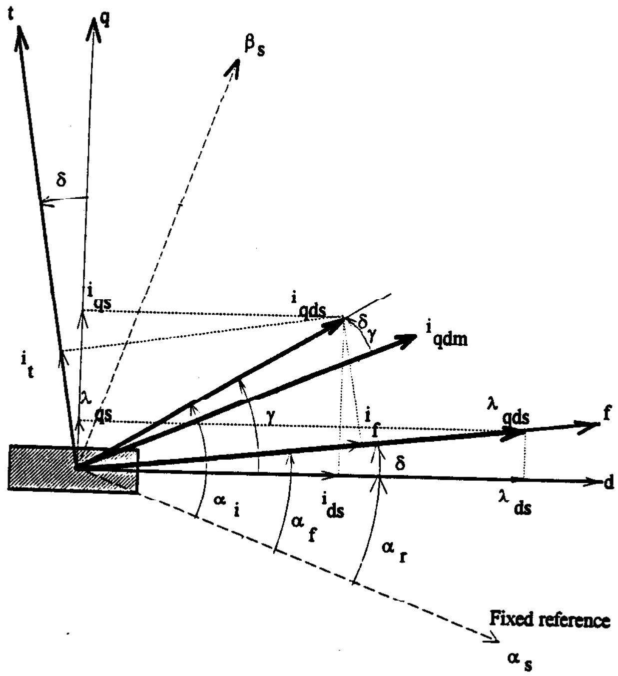

Figure 1. Vector diagram of the SynRM

(b) maximum rate of change of the torque control, obtained when $\gamma = \tan^{-1}\zeta$;
(c) maximum power factor control, obtained when $\gamma = \tan^{-1}\sqrt{\zeta}$;
(d) constant current in inductive axis, when the i_{ds} component of the current is maintained constant, and the torque is controlled by the q axis current component.

Each of the proposed control methods has its advantages and its drawback. As a general conclusion, the first three methods (which are constant angle control methods), give better rate of change of the torque at high speeds and torque, while the fourth method is better at lower speeds and torque. For all control strategies, bigger values of ζ improve the performances of the system. In [9], a flux oriented reference frame is used for the machine control, thus a rapid torque response control is implemented. The method was further extended in [10], where a flux observer is used, based on machine measured voltages and currents.

All the mentioned papers use position transducers. Several attempts were also made in order to implement sensorless SynRM drives. Thus, in [11], several approaches for nonlinear state observers schemes are studied, based on the voltages and currents of the motor. High steady-state accuracy is obtained for the position estimates, but the estimate converges only after a certain transient time (approximate 2π electrical radians) and requires a sophisticated computational algorithm. In [9], the motor synchronization is performed by a closed loop control of the torque angle γ, calculated on the basis of an approximated steady-state model of the motor. A different approach is presented in [10], based on measured terminal quantities of the motor.

2.3. The implemented control scheme

The implemented approach in this study uses a modified constant field current component control, in a field reference frame, with several position estimation schemes based on the third harmonic voltage of the machine. Taking into account that the choice of the flux oriented reference frame offers a fast torque response control ([9], [10]), and that the position of the air-gap flux is obtained from v_3 signal, this choice proves to be very convenient for the SynRM control implementation. Thus, the quadrature axes (f, t) are chosen as reference frame, where f is the field axis and t perpendicular to f (see Figure 1). In this reference frame, the motor equations (neglecting the iron losses), are (considering that λ_t=0 and $d\lambda_t/dt$=0):

$$v_{ts} = i_{ts}r_s + \frac{d\lambda_t}{\delta t} - \omega\lambda_f$$

$$v_{fs} = i_{fs}r_s + \frac{d\lambda_f}{dt} + \omega\lambda_t \tag{1}$$

$$T_e = \frac{3}{2}\frac{P}{2}(\lambda_f i_{ts} - \lambda_t i_{fs})$$

Here one cannot find a decoupled relation between the current components i_{fs}, i_{ts} and the flux components λ_f and λ_t respectively. As one can see from the torque relation, a direct control scheme may be implemented, which control λ_f and i_{ts} simultaneously. An observer block is needed to implement both position and flux amplitude detection, from the third harmonic voltage v_3, and also speed estimation. Depending

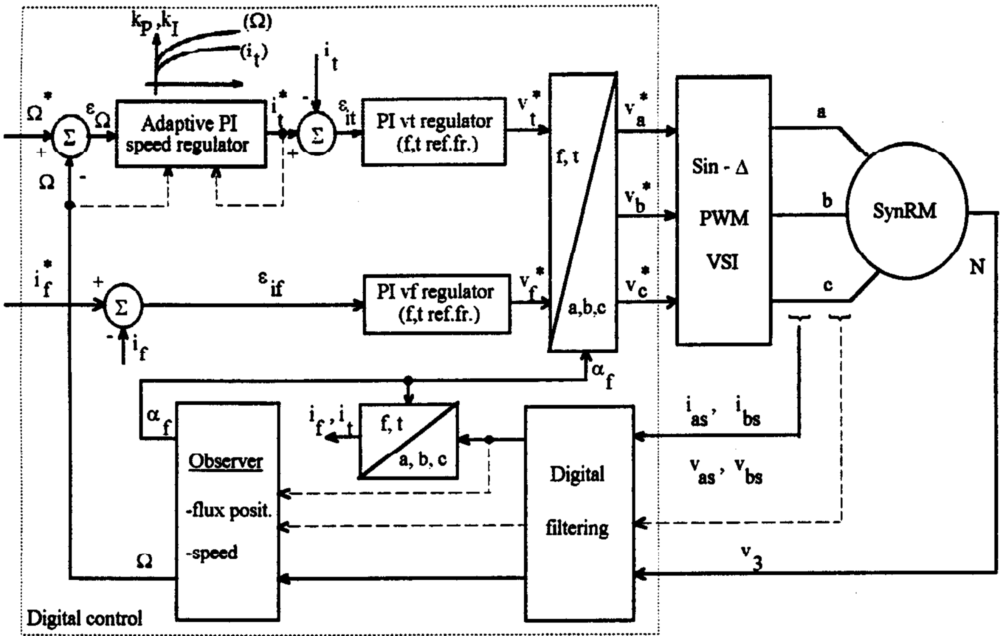

Figure 2. Current control of the SynRM in a flux reference frame.

on the estimation scheme (see next paragraph), the stator voltages and currents may also be used in the estimation process. A constant flux control strategy ($\lambda_f^* =$ constant) may be used. Thus, the i_t current component becomes the torque control variable. The main drawback of the suggested scheme lies in the problems associated with the flux fundamental amplitude detection from the v_3 signal. The nonlinear dependence $\lambda_f = f(\lambda_3)$ must be measured and stored into the digital control system. Furthermore, poor flux estimates are expected at low speeds, where the v_3 signal is very poor due to its harmonic components.

The control scheme from figure 2 was then proposed. Instead of a flux control on the field axis f, one imposes the current component, i_{fs}. A constant i_{fs}* reference current was considered, but different approaches could also be used (for optimal control implementation).

Using the notations from Figure 1, one finally has that the flux expression is given by

$$\lambda_f = i_f \frac{L_d}{\cos^2\delta + \sin^2\delta\tan\delta}\, i_f\, i_t \qquad (2)$$

and the torque is

$$T_e = k\lambda_f i_t = \frac{3P}{4} L_d \frac{1}{\cos^2\delta + \sin^2\delta\tan\delta}\, i_f\, i_t \quad (3)$$

These relations may be used in order to estimate the effect of the control strategies as compared with the (d, q) control approach ($i_{ds}^* =$ constant), and with the (f, t) controller (λ_f^* = constant).

As stated before, for a well designed, high performance SynRM drive, one has high values for the ratio $L_d/L_q = \mathrm{tg}\,\theta$. For the tested SynRM, the experimentally value of this ratio was around 10. [7]. Simple calculations for these cases lead to the conclusion that the position difference between the rotor and the flux axes, even for important variations of the torque (current) angle γ, is only few electrical degrees. At the same time, the ratio i_{qs}/i_t is practically constant, and thus one can use i_t^* as the torque reference instead of i_{qs}^*. A certain decrease of ratios i_{ds}/i_f and $\lambda_f/(L_d\, i_f)$ is observed, at increased values of the torque. Thus, one can expect a certain performance alteration for the proposed control scheme, for high torque values (acceleration periods and/or high load torque). In order to compensate this variation (as well as that observed in [9] at high speeds), for a constant i_f^* and using i_t^* as torque control variable, an adaptive speed regulator scheme was implemented (see Figure 2). At high levels of the torque, or at high speeds, the regulator coefficients are increased, thus compensating the decrease of the torque command at i_f^* constant. Experimentally obtained results validated this method.

3. POSITION AND SPEED ESTIMATION FROM THE THIRD HARMONIC SIGNAL

Two different approaches were used to detect the position and the speed of the magnetic flux λ_f of the SynRM. Both of them use the zero crossings of v_3 voltage for exact position up-dating, and different approaches for position detection between v_3 zero crossings.

3.1. Saturation harmonic air gap flux components

The direct field orientation strategies proposed in this paper are based on the detection of the third harmonic air gap voltage [2]. In this work the authors show that the resultant component of the air gap flux density when the machine is saturated includes the all odd harmonic components, including the triplens 3^{rd}, 9^{th} and so forth. These spatial harmonic components are synchronously rotating with the fundamental air gap flux component. Furthermore, it was shown that the third harmonic is the dominant harmonic component and that it is responsible for the induction of a third harmonic zero sequence voltage component in the stator phase voltages. The air gap flux can be derived from this third harmonic voltage signal in a very reliable manner since the useful signal is large and practically noise free.

If the machine phases are connected in wye without a neutral connection, no zero sequence components (triplen harmonics in a three phase system) will exist in the current. When the three phase voltages are summed, the fundamental and characteristic harmonics are canceled and the resultant wave form contains mainly a third harmonic component. The amplitude of the induced third harmonic phase voltage is a function of the saturation level that is dictated by the amplitude of the fundamental component of the air gap flux. Therefore, a function relating the third harmonic stator voltage and the air gap flux exists and it may be used to determine the fundamental air gap flux linkage of the machine, λ_{m1}.

3.2. Location of the air gap flux from the third harmonic signal

For the suggested measurement method, some practical implementation aspects must be considered. They are related with the use of the measured third harmonic voltage signal (v_3) in order to obtain the flux linkage information. One of these aspects concerns the estimation of the amplitude of the fundamental air-gap flux linkage from the amplitude of the third harmonic flux. This non-linear relation is machine dependent and requires supplementary experimental and/or computational effort. In addition, experiments made at low speeds of the machine yield a poor third harmonic signal (highly distorted signals), which makes difficult a correct estimation of the flux. When low order harmonics are encountered in the measured v_3 signal, one cannot apply digital filtering in order to eliminate them, without seriously affecting the characteristics of the third harmonic term. Thus, different approaches must be tested for the detection of these parasitic harmonic components.

Related to this aspect, the problem of the voltage integration in order to obtain the corresponding third harmonic flux (λ_3) has also several drawbacks. The integration of the offsets and noise that is present in the third harmonic voltage v_3, and special problems encountered with the signal during transient regimes lead to bad flux estimates. This observation applies for both analog and digital integration schemes, even if special filtering methods are used in order to improve the v_3 signal shape.

As a solution to this problem, an on-line estimation of the v_3 signal parameters may be used, thus allowing an analytical computation of λ_3. Simulations using a simple recursive least squares estimation scheme gave good results for both steady-state and transient regimes and allowed the on-line detection of the amplitude and phase values for different harmonic components of the measured v_3 signal. Good estimates of λ_3 were then analytically obtained. The main drawback of the method is the increased computational effort required from the digital control system in this case.

Considering all these aspects related with the flux amplitude computation and the voltage integration, a simpler approach was selected for implementation. In this case, only the position of the flux is detected from the third harmonic stator phase voltage signal. Moreover, the position information is obtained directly from the v_3 signal, thus avoiding the integration process to obtain λ_3.

Experimental tests showed, unfortunately, that at very low speeds, the v_3 signal becomes very poor. The noise level is high, and also the harmonic content is substantial. Nevertheless, if properly filtered and averaged, the signal still contains useful information. This information is derived from the zero crossings of v_3, which may be used in order to have the position of the third harmonic voltage (and consequently of the air gap flux), with an accuracy of 60 electrical degrees. As one can see from Fig. 3, each zero crossing of v_3 means a $\pi/3$ increment of the third harmonic voltage position and, with the corresponding shift, of the third harmonic air-gap flux. Thus, an exact position information is available, in a very convenient manner.

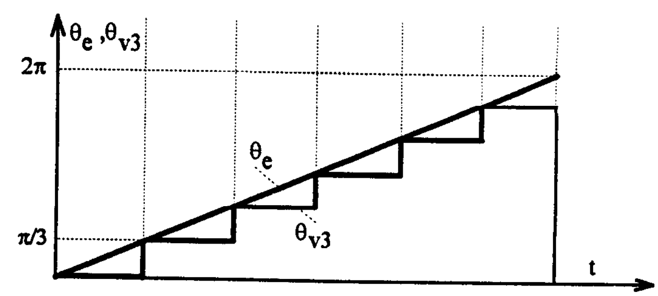

Figure 3. Position detection from zero crossings of the v_3 signal

Two practical problems arise when dealing with this approach:

(1) the "position transducer" thus implemented is a relatively coarse incremental one, and a procedure to absolutely locate the position must be implemented;

(2) an estimation procedure must be found, in order to express the position variation between the zero crossings of v_3.

3.3. Absolute position detection of v_3 signal

Two different means to absolutely locate the position of the flux from the v_3 signal for a three phase ac machine have been studied and tested.

(a) Use of the stator phase voltages. The first method uses the relative position of the stator phase voltages of the machine in order to detect the absolute position of v_3. An iR_s voltage drop correction is needed in this case in order to obtain correct results. The absolute position is detected at each zero crossing of v_3. Problems arise at very low speeds, where the high values of the harmonics in the stator voltages can lead to erroneous results.

(b) Use of a relative incremental scheme. The second method is similar to relative incremental transducers with a zero index signal. At the start, the machine is positioned in a known position, and any further zero crossings of v_3 are used to up-date the absolute position information. This approach is simpler than the previous one (no measurements of voltages and currents, nor iR_s corrections are needed). The main problem is that any parasitic zero crossing of v_3 leads to an erroneous position estimate.

Tests carried on the test SynRM give very good results for both methods. Finally, the second one was implemented. Satisfactory operation at speeds as low as 0.5 Hz (electrical) were attained, due to the fact that the v_3 signal zero crossings still give a correct information about the position of the flux, even if v_3 has a poor wave form at this frequency.

3.4. Estimation of the position between the zero crossings of v_3

As seen in Sec. 3.3, only at zero crossings of v_3 is the exact position available. Between these instants, special estimation techniques must be implemented in order to obtain an estimate of the position. Two different classes of approaches were tested.

(a) Use of the stator voltage vector position variation. In this approach, the phase variation of the stator voltage vector was computed in a stator reference frame, and was used to correct the position estimate of the flux, up-dated to the correct value at zero crossings of v_3. A low limit speed is reached, below which the system fails to operate (mainly due to the Ri_s drop, harmonic content of stator voltages, measurement noises, etc.), and the method must be combined with an open-loop control scheme, or a different position estimation scheme. Even if special corrections were used in order to obtain better signals at low speeds (for Ri_s, 5^{th} and 7^{th} harmonics elimination, special filtering techniques), below a certain frequency, the system fails to function.

(b) Use of position estimation schemes. An alternative approach is to implement a complete position estimation scheme based on the absolute position value detected at v_3 zero crossings, and on a dynamic model of the system. Two estimation schemes were tested and implemented.

(b1) The first approach uses the mean speed of the field, computed from the time length of a v_3 pulse, set by the zero crossings of v_3 [6]. This mean speed value needs correction terms in order to incorporate the dynamic of the system and the modeling errors. Figure 4 presents the final scheme that was implemented. The mean speed Ω is computed at each zero crossing of v_3, then is corrected with the $\delta\Omega$ term (modeling the mechanical dynamic of the system), and $\delta\Omega_{er}$ (correction for load torque variations and/or modeling errors). Each zero crossing of v_3 updates the computation process (all integral terms are zeroed).

b2) The second approach uses an estimation based on a position/speed/acceleration observer scheme. It is based on similar schemes presented in [12] for speed estimators. Figure 5 presents this scheme. As one can see, the (machine+load) model is used to obtain the estimates. The $k_T\, i_t^*$ term modulates the electromagnetic torque; $k_\Omega \Omega$ the viscous torque; the output of the PI position error regulator, $\varepsilon_{\theta er}$, modulates the (variable) load torque; the $k_{\Omega er}\, \delta_{\theta er}$ term introduces a supplementary correction to the speed estimate. Once again, an improved behavior of the scheme was obtained by restarting the integration process (PI position regulator and position estimate) at zero crossings of v_3, and forcing θ to the exact value, at the v_3 zero crossings.

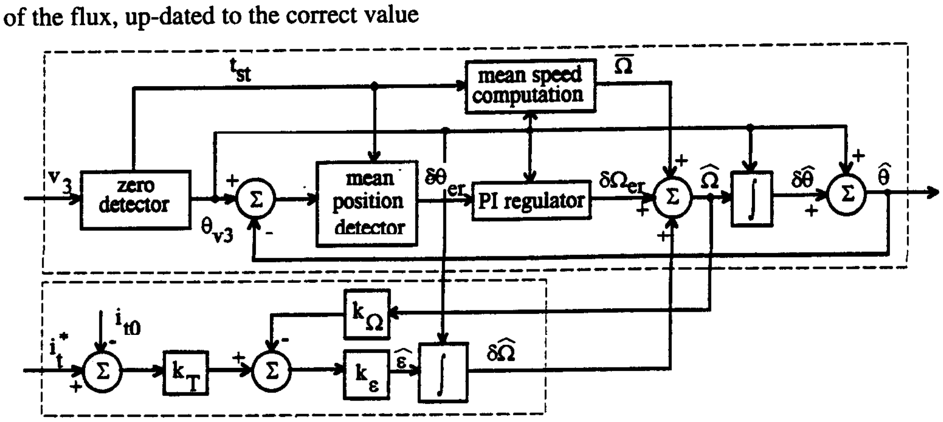

Figure 4. Estimation scheme based on the mean speed of the field.

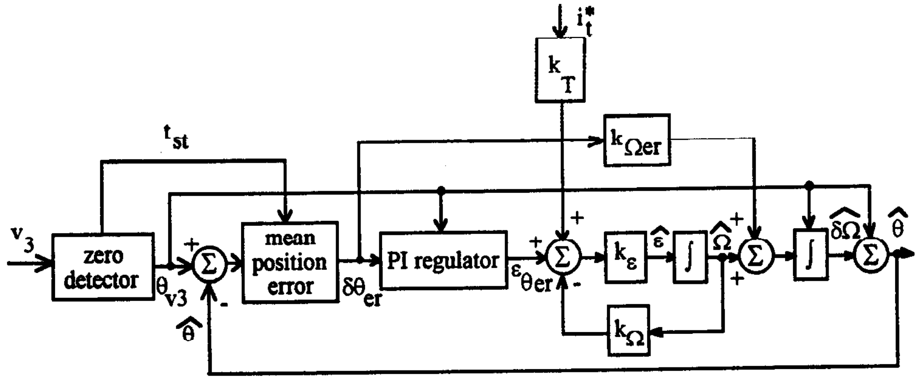

Figure 5. Optimal position observer based on the zero crossings of the v_3 signal.

4. IMPLEMENTATION ASPECTS AND EXPERIMENTAL RESULTS

A four poles rotor, three phase, 7.5 hp, 1750 rpm rated speed, 230 V rated voltage, 10 A rated current SynRM was used for the tests. A Motorola 56001 DSP processor based digital system was used to implement the digital control scheme. Twelve bit accuracy A/D and D/A converters were used to interface the digital system with the analog transducers and the control equipment. A 2.5 kHz sampling frequency was used in order to implement the estimation and control procedures. Interrupts, generated by the processor timer, activate the control sequence of the program. Special care was given to a proper digital processing of the signals. A 6-th order Butterworth low pass filter was used for all the signals (of fundamental or third harmonic frequency). Its linear phase characteristic in the frequency range of interest (less than 1 kHz), preserves the time delays between the signal components at different frequencies. At low speeds, supplementary averaging procedures were also implemented in order to improve the v_3 signal shape. Due to the high noise content in the signal, a special adaptive locking procedure for the zero crossing detection was also needed at these low speeds in order to eliminate false zero crossing detection.

The different position estimation schemes proposed in Sec. 3 were tested. The use of the stator voltages vector position variation, to estimate the flux position between zero crossings of v_3 give good results at high and medium value speeds, but poor behavior at low speeds. A value of 2.7 Hz (electrical) was the determined as the limit frequency for which the system operated satisfactorily (see Fig. 6). Below this limit, due to the position estimation errors (caused by the highly distorted processed signals), the system becomes unstable. Also, an open loop starting procedure is needed in this case. Reversal of the machine is possible, if the zero speed region is crossed with a certain minimum acceleration.

Better results were obtained when using the estimation schemes form figures 4 and 5. The minimum frequency limit in this case, in steady-state regime, was 0.5 Hz, for the

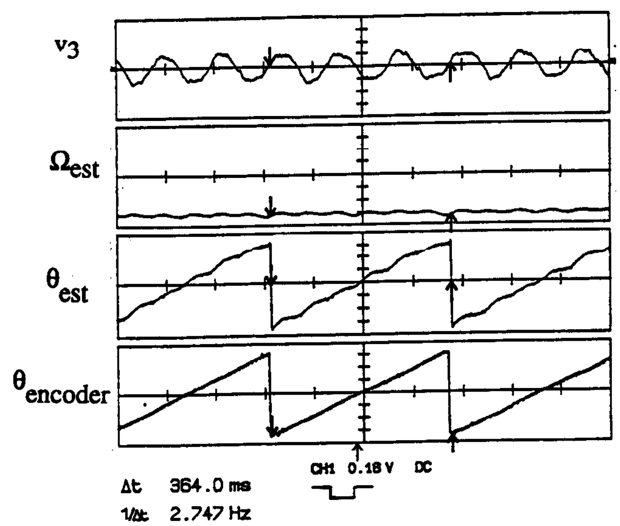

Figure 6. Position and speed estimation, using v_3 and the stator voltages; steady-state regime, low speed (v_3 signal - ch.1; estimated speed - ch.2; estimated position - ch.3; measured position - ch.4)

optimal estimation scheme from Figure 5 (see Figure 7). The estimates are very good in this case. The main problem at low speeds in these cases is the estimation schemes sensitivity for large variations of the load and/or machine parameters. The optimal estimator scheme is more sensitive than that based on the mean speed, but gives better estimates for constant parameters systems.

These schemes work also well at the starting of the machine. A constant voltage is first applied between two of the machine phases, until an initial pre computed position is reached. The control scheme works from the starting moment

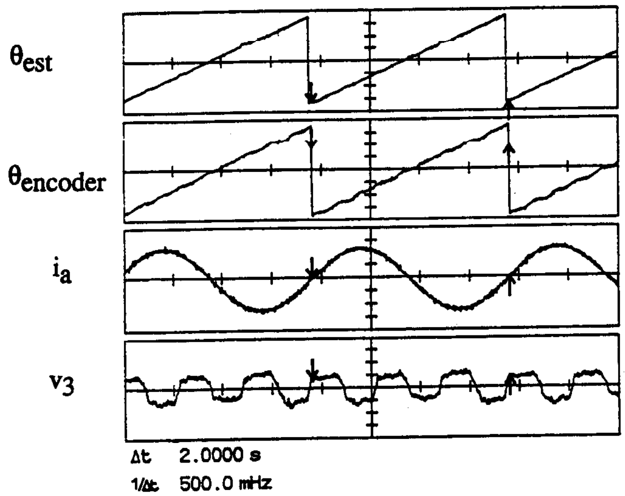

Figure 7. Position and speed estimation, using v_3 and the optimal estimator scheme; steady-state regime, low speed (estimated position - ch.1; measured position - ch.2; phase current - ch.3; v_3 signal - ch.4)

The relative incremental position estimation scheme proposed in par.3.3 is used as the machine is started. False zero crossings of v_3 are eliminated. One uses adaptive control coefficients, as presented in par. 2.3. With proper control parameter adjustments, the only restriction is to have a reference speed greater than the limit value (i.e. 0.5 Hz). Figure 8 presents such a starting regime of the machine.

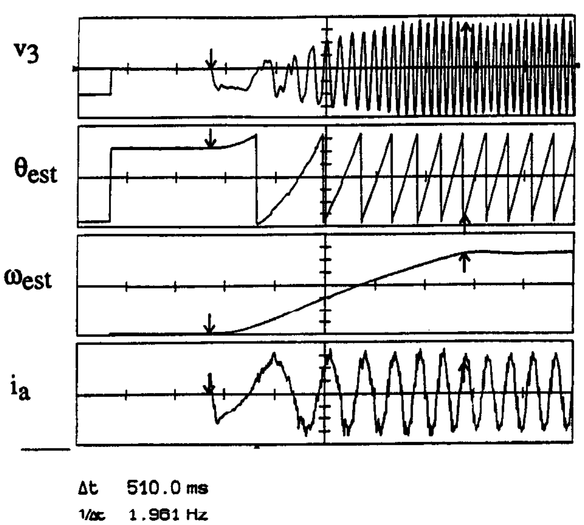

Figure 8. Alignment and starting of the machine (v_3 - ch.1; estimated position - ch.2; estimated speed - ch.3; phase current - ch.4)

The start of the machine, followed by some reversing regimes, is also presented in Figure 9. Very good results are obtained. The control scheme is used during the zero speed region crossing, as well.

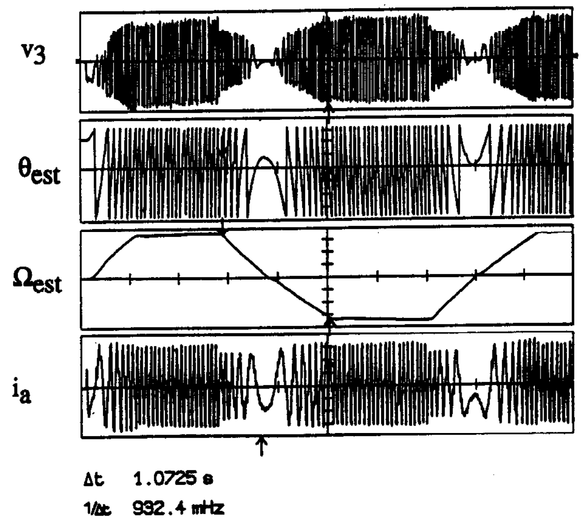

Figure 9. Start, acceleration and reversing of the machine (v_3 - ch.1; estimated position - ch.2; estimated speed - ch.3; measured phase current - ch.4)

5. CONCLUSIONS

A simple and low cost scheme for direct field orientation control of a synchronous reluctance motor has been proposed. The need for both position and speed transducers is eliminated. The method is based on the concept of locating the fundamental component of air gap flux from the third harmonic voltage component induced in the stator phase voltages when the machine is in saturated condition. When the stator phases voltages are summed, the resultant signal contains a dominant third harmonic component, which can be used to detect the air gap flux position and implement a direct control scheme. Utilization of this method requires the stator to be star connected with access to the neutral connection.

Further analysis and experimental work must be done in order to completely validate the proposed estimation schemes, to study their sensitivity with the motor and load parameters and to find their ultimate performance limits. Anyway, one may conclude that the main goal of this study proved to be reachable, i.e. to obtain a cost-effective, simple and reliable measurement and control structure.

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