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Direct Field Orientation Controller Using the Stator Phase Voltage Third Harmonic

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DIRECT FIELD ORIENTATION CONTROLLER USING THE STATOR PHASE VOLTAGE THIRD HARMONIC

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Abstract A new direct field orientation controller for induction machines 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 spatial saturation harmonic components rotating at synchronous frequency which 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 and a high frequency component due to the rotor slot ripple. Extensive experimental results showing the practical problems of detecting the third harmonic voltage signal are presented.

1. INTRODUCTION

The advantages of direct field orientation over the indirect type in overcoming sensitivity of the control to changes in machine parameters has been discussed very thoroughly in the literature [1-4]. However, implementation of the direct type of field orientation has been regarded as being difficult in practice by virtue of the delicate sensors (e.g. search coils or Hall effect sensors) needed for the control. These sensors, besides contributing a considerable amount to the total cost of the controller and affecting the reliability, often impose severe limitations on the range of machine operation.

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 induction machine operation under saturation and the consequent generation of harmonic components in the air gap flux has been presented in [5-7] and has been utilized to correct the time constant model in an indirect field oriented control scheme [8,9]. A new control principle which uses the third harmonic voltage to sense the air gap flux position and thus realize direct field orientation, is proposed in this paper. The controller requires 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 speed or position measurement in the torque loop.

2. SATURATION HARMONIC AIR GAP FLUX COMPONENTS

The direct field orientation strategies proposed in this paper are based detection of the third harmonic air gap voltage [5,6]. In these works 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 3rd, 9th, 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 and the rotor speed can then be derived from this third harmonic voltage signal in a very reliable manner since the useful signal is large and practically noise free.

The concept of synchronously rotating saturation harmonics is illustrated in Figure 1 which depicts the air gap flux components for a condition of saturation occurring in both the stator and rotor teeth. As the stator and rotor teeth begin to saturate, the teeth with the highest flux density will saturate first so that the flux distribution around the air gap will assume a flattened sinusoidal form with peak value B_{sat} as shown in the figure. In addition, for most machines the air gap flux density is also modulated by a high frequency component due to the existence of stator and rotor slots. This high frequency component is proportional to the rotor mechanical speed and can be utilized as a means to measure the speed, thereby eliminating the need for a tachometer [4].

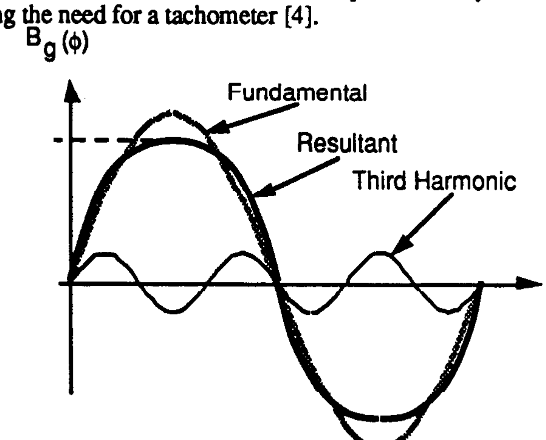

Fig. 1 Air gap flux density along the air gap with stator and rotor teeth undergoing saturation.

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. Also, if the rotor cage is assumed to be equivalent to a delta winding connection, the induction machine can be viewed as an ungrounded three phase wye-delta transformer where no circulation of zero sequence current is possible in the wye side. Therefore, the stator currents, and consequently the air gap MMF, will contain only the so called characteristic harmonics (5^{th}, 7^{th}, 11^{th}, and so on), while the air gap flux and, consequently, the phase voltages contain the predominant third harmonic and higher frequency slot components.

When the three phase voltages are summed, the fundamental and characteristic harmonics are canceled and the resultant wave form contains mainly a third harmonic together with higher frequency components due to the rotor slots [5], [6]. The amplitude of the induced third harmonic phase voltage is a function of the saturation level which 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 voltage exists and it may be used to determine the fundamental air gap flux linkage of the machine, λ_{m1} .

3. LOCATION OF THE AIR GAP AND ROTOR FLUXES FROM THE THIRD HARMONIC SIGNAL

3.1 Air Gap Flux Orientation

A practical problem which arises when implementing the rotor flux orientation control scheme comes from the fact that the air gap flux is not absolutely located by the third harmonic voltage signal which comprises information concerning only the sine or cosine component of the air gap flux (d or q component). Therefore, it is necessary to extend the control methodology to obtain the two quadrature components of the air gap flux. Figure 2 shows the air gap flux fundamental and third harmonic components together with one of the stator line currents for a loaded machine. Clearly point B in the third harmonic wave locates the maximum of the fundamental component of the air gap flux (point A) which can then be referred to the stator current maximum value (point C) by the displacement angle γ_{im} . Hence, by detecting point A and measuring the angle γ_{im} the absolute position of the stator current maximum value (i.e. spatial position of the stator MMF) can be known.

With the fundamental of the air gap flux linkage located from the third harmonic voltage signal, a direct air gap field orientation strategy can be implemented as a first intuitive option. In this control scheme, the air gap flux is aligned with the d-axis of the d-q plane with the stator current components i_{qs} and i_{ds} being the command variables for the torque and flux respectively. Unfortunately, this type of field orientation scheme does not allow a complete decoupling between the command variables i_{qs} and i_{ds} , which is only achieved by the introduction of a decoupling network. This decoupling network, however, introduces the disadvantage of being dependent on sensitive machine parameters and also contributes to an increase in the complexity of the control algorithm. Another potential limitation of this type of controller relates to the static stability of the drive which will present a limited pull out torque if a current command is used as the flux control [10]. It is clear that the required decoupling networks can be readily incorporated if desired in the future.

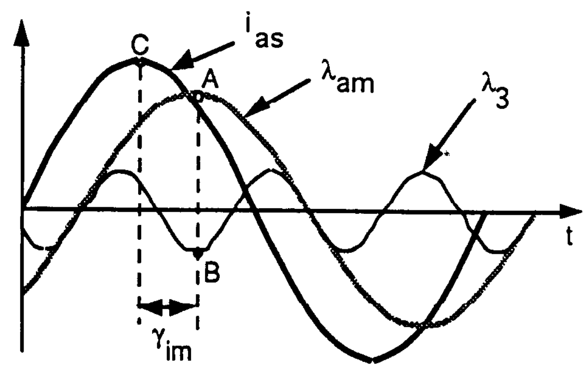

Fig. 2 Air gap flux linkage and its third harmonic component referred to one of the line currents.

3.2 Rotor Flux Orientation

Another possibility of air gap flux orientation is the rotor flux orientation strategy. In this case, an additional computation is necessary in order to obtain the rotor flux from the air gap flux, as described by Eqs. 1 and 2.

$$\lambda_{qr}^{s} = \frac{L_r}{L_m} \lambda_{qm}^{s} - L_{lr}\, i_{qs}^{s} \tag{1}$$

$$\lambda_{dr}^{s} = \frac{L_r}{L_m} \lambda_{dm}^{s} - L_{lr}\, i_{ds}^{s} \tag{2}$$

Although dependent on machine parameters, the rotor flux can be obtained with reasonable accuracy since the rotor leakage inductance, L_{lr} , and the ratio of total rotor inductance to air gap inductance, L_r/L_m, are only moderately dependent on the saturation level [11].

3.3 Implementation considerations

For any of the two previously proposed control reference frames, some practical implementation aspects must be considered. They are related with the use of the measured third harmonic voltage signal (v_3) used to obtain the flux linkage information.

Thus, a practical aspect related to the implementation of this scheme 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 component of interest (the third harmonic component, in this case). Thus, different approaches must be tested for the detection and elimination 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 noises which are 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 or 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.

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.

Several approaches which may be used in order to solve these problems are presented in the next section.

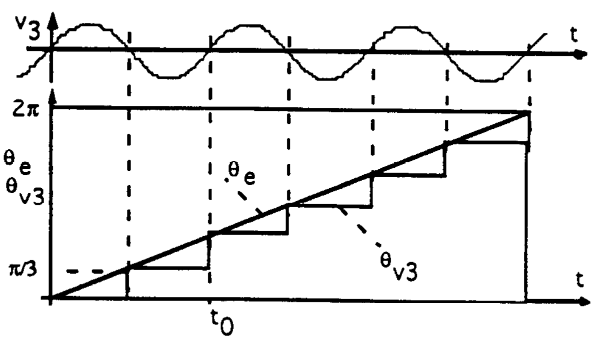

Fig. 3: Position detection from zero crossings of the v_3 signal.

4. ABSOLUTE POSITION LOCATION 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. Both give good results, each having different advantages without critical drawbacks.

4.1 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. If one does not consider for the moment any iR_s voltage drop in the stator phases, then at each zero crossing of v_3, one of the phase voltages is zero, and the other two have opposite signs. In a realistic approach, one must compensate the iR_s drop, for each phase. This means that the currents must be measured, an estimate of the stator resistance R_s must be found, and finally the estimated induced phase voltage can be obtained. One may obtain an estimate of the stator resistance using the measured value of the phase voltage at zero crossing of v_3, and of the corresponding phase current. This approach also has the advantage of on-line detection of changes in R_s due to temperature changes, etc.

After correcting the iR_s drop due to measurement noises and to harmonic components of the stator voltages (important mainly at low frequencies), some simple tests comparing the phase voltages must be implemented for determining which of the six zero crossing corresponds to the flux position of a given phase.

The approach considered has the advantage that it detects the absolute position at each zero crossing of v_3. On the other hand, problems arise at very low speeds where the high values of the harmonics in the stator voltages could lead to erroneous results when applying the proposed position detection tests. Even if these errors appeared for only one step, they obviously represent significant errors (being used by the control system) and seriously affect the behavior of the system. Special treatment of these cases is still possible. For example, one should not allow modifications of θ_3 having more than one $\pi/3$ radian step between two consecutive zero crossings of v_3. However, this solution is similar to the next, simpler approach.

4.2 Use of a relative incremental scheme

The second method is similar to relative incremental transducers with a zero index signal. Thus, at the start, the machine is first positioned in a known position, and any further zero crossings of v_3 are used to increment or decrement the absolute position information (software memorized), depending on the speed sign.

The approach is much easier to implement than the previous one since one does not need to implement any measurements of voltages and currents nor iR_s corrections. On the other hand, the main drawback is that of any relative transducer. That is, any parasitic zero crossing(s) of v_3 lead(s) to unacceptable position error, which continue to be maintained in the estimate.

Tests carried on a test IM machine give very good results for both proposed methods, and finally the last solution was implemented. Satisfactory operation at speeds as low as 1.2 Hz (electrical) were attained. The practical problem of reversing of the machine as well as numerous other transient regimes were also tested. As stated before, all these results are due to the fact that, even at very low speeds, if the v_3 signal has a poor waveform its zero crossings still give correct information about the absolute position of the air-gap flux of the machine.

5. ESTIMATION OF THE POSITION BETWEEN THE ZERO CROSSINGS OF V3

As mentioned, the estimated position information θ_{v3} (see Fig. 3), obtained from the zero crossings of v_3, gives very poor position information, with only six steps over a full period with $\pi/3$ radian accuracy. 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 as close as possible to the real position value. Two different approaches were tested.

5.1 Use of the stator voltage vector position variation

In the first approach, the phase information contained in the stator voltages was used. Thus, between zero crossings of v_3, one computes the position of the stator voltage vector, in a stator reference, and uses it to realize the position estimate. Due to the fact that several error sources affect the measured stator voltages, mainly at low speeds (iR_s drop, harmonic content, measurement noise), a better approach is to use the variations of the position of the stator voltage vector, to compute the estimated position.

Due to the same problems related with the stator phase voltages at low speeds, a limit speed is inevitably reached, below which the system fails to operate. In order to reach as low a speed as possible the method requires voltages and currents measurement, iR_s voltage drop correction, filtering of all data, implementation of the arc tangent function computation, etc. Below the limit speed, an open-loop control system must be implemented if the motor is to be reversed. This also requires an independent position estimation procedure within this low speed region and an accurate switching scheme from the open loop to the closed loop control and vice versa.

5.2 Use of position estimation schemes

An alternative approach is to implement a complete position estimation scheme based on the absolute position estimate and on a dynamic model of the system (motor + load). The first intuitive approach uses the mean speed of the field, computed by measuring the time length of a pulse set by the zero crossings of v_3. Thus, considering two consecutive zero crossings of v_3 at instants t_{j-1} and t_j, the mean speed on the time interval (t_{j-1},t_j) is

$$\Omega_j = \frac{\theta_{s3}}{t_{st}},$$

where $\theta_{s3} = \pi/3$ is the position step increment between two consecutive zero crossings of v_3 and t_{st} is the time length of the v_3 pulse. This estimate can be used for the next v_3 pulse (from t_j to t_{j+1}).

The estimation scheme works well in steady-state regime, but unfortunately provides poor estimates during fast transients of the system due to the fact that one uses the mean speed, and thus no acceleration information is available. Also, the estimated mean speed corresponds to the previous v_3 pulse and thus a dead time is present in the mean speed estimate.

Further improvements of this scheme, by introducing corrections which take into account a model of the machine and load, can lead to very good position estimates, even during transient operation. A modified estimation scheme based on these principles has been implemented. The scheme uses two correction terms for the speed estimate. Figure 4 presents the final estimator.

One correction term, $\delta\Omega$, is obtained from a block modeling the mechanical dynamic of the system. This term contains constant parameters model of the system, including the constant component of the load torque (modeled by the i_{q0} term).

A second correction term, $\delta\Omega_{er}$, is computed as the output of a PI regulator, having as input the mean position error, detected at v_3 zero crossings, (at time instant t_j). This term represents the correction needed for the load torque variations and/or modeling errors in the computation

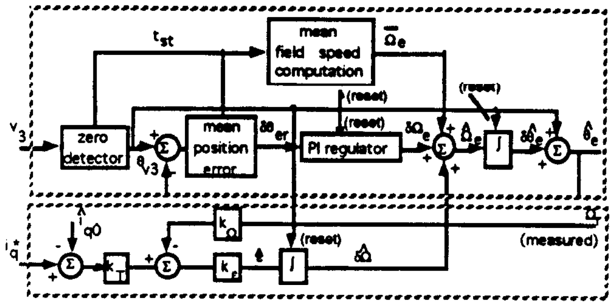

Fig. 4: Estimation scheme based on the mean speed of the field.

of the $\delta\Omega$ term.

A complete optimal observer, both for the air-gap field position and motor speed may also be implemented, similar to the estimation schemes presented in [12]. For such an approach, the motor speed may be estimated from the air-gap field speed, and the slip frequency detected from the machine model. In such case, one should pay particular attention to the parameter sensitivity of the estimation scheme. Problems may be encountered due to this fact, and further investigations must be done in order to validate such schemes.

6. DIRECT FIELD ORIENTATION CONTROLLER USING THE STATOR PHASE VOLTAGE THIRD HARMONIC

As stated in the preceding paragraphs, the air-gap flux linkage may be located from the third harmonic voltage signal (v_3). Thus, a direct air-gap field orientation strategy may be implemented. Since the goal of this study was to implement a cost-effective, medium performance direct field controller for an induction machine the air-gap flux has been aligned with the d-axis of the d-q plane for simplicity. In this case, the use of the proposed control approach gives acceptable results, even if this type of field orientation scheme does not allow a complete decoupling between the command variables i_{qs} and i_{ds}, which may only be achieved by the introduction of a decoupling network.

Similar considerations may be applied concerning the use of a speed information. A low cost transducer (tacho-generator) may be used, or only a speed estimation scheme, based on the machine model, may also be considered. The first approach was considered for the experimentally implemented scheme.

The general control scheme for the speed control of the machine is presented in Fig. 5. In this figure the torque control variable i_q^* is obtained from the speed error, using a PI regulator. The i_d^* and i_q^* reference signals are then used to close the current loops, in the air-gap flux d-q frame. A voltage source inverter is controlled by the phase reference voltage commands, obtained from the PI voltage regulators.

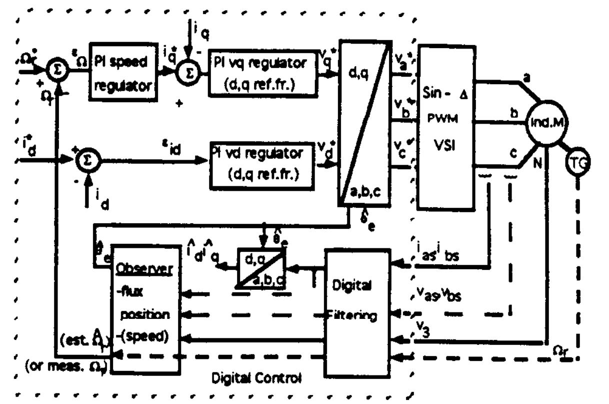

Fig. 5 The direct control scheme for the IM, using the third harmonic voltage signal.

In a general approach, the current command component i_d^* must be obtained using a decoupling network, in order to insure a constant air-gap flux signal. Considering the medium performance requirements accepted for this study, only a constant value for i_d^* was considered for the experimental tests.

7. EXPERIMENTAL RESULTS

Experimental tests were done based only on the estimation scheme proposed in Fig. 4, with the speed of the motor being measured from a speed transducer. The proposed control structure was tested on a 7.5 hp, 4 pole, 1755 rpm, 460 V, 9 A induction machine. 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 this 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 of 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.

Good results were obtained with the proposed estimation and control schemes. Figures 6 to 9 present some experimental data obtained for the tested induction motor. Figure 6 presents the start of the machine. In this case, at start, an initial premagnetization of the machine is performed, by first imposing the i_d current at its nominal value. Only after the stabilization of this current the machine is started. The starting position is considered to be the zero position. Quite good results are evident. A certain sensitivity of the scheme to the load parameters is to be expected, which more evident during this starting regime. Further studies must be done in this direction to optimize performance.

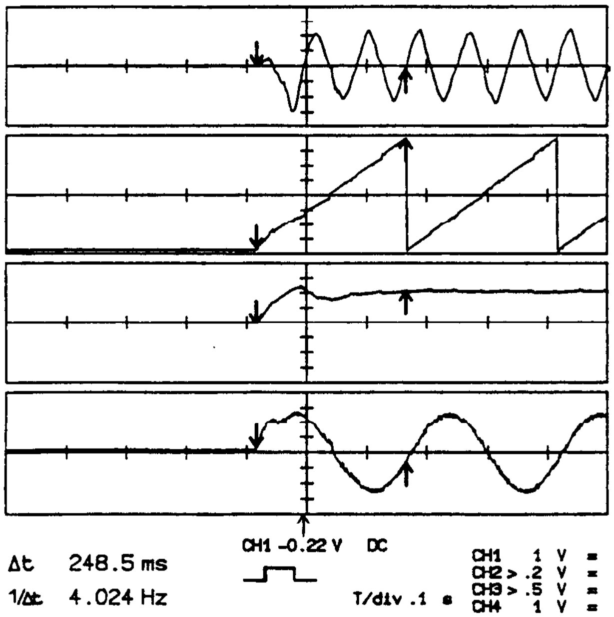

Fig. 6 Start of the IM: (ch.1) v3 signal; (ch.2) estimated field position; (ch.3) measured motor speed; (ch.4) stator phase current.

Figure 7 presents a steady-state operation at low speed (1.2 Hz electrical). As one can note, a good position estimate is obtained (channel 2), from the zero crossings of v_3 (channel 1); the phase voltage (channel 3) and current (channel 4) are also presented.

Figure 8 shows a reversal of the machine. The main problem is the region around the zero speed (very low speeds), where poor v_3 signals, and consequently, poor position estimates are obtained. If the motor speed only need cross this region, the scheme clearly functions very satisfactorily.

In order to verify the correctness of the position estimate, and the decoupling of the two current components, some separate tests with an open loop for the speed control were implemented. Thus, a bang-bang step i_q reference was imposed for the motor. Figure 9 shows the obtained results in this case. The oscillations were not around the zero speed but around a mean non-zero speed.

In the figure, the v_3 signal (channel 1), measured i_d current (channel 2), measured i_q current (channel 3) and measured motor speed (channel 4) are presented. The variations of the speed are quite steep so that the differences between the motoring and regenerating regions may be due to the static friction torque.

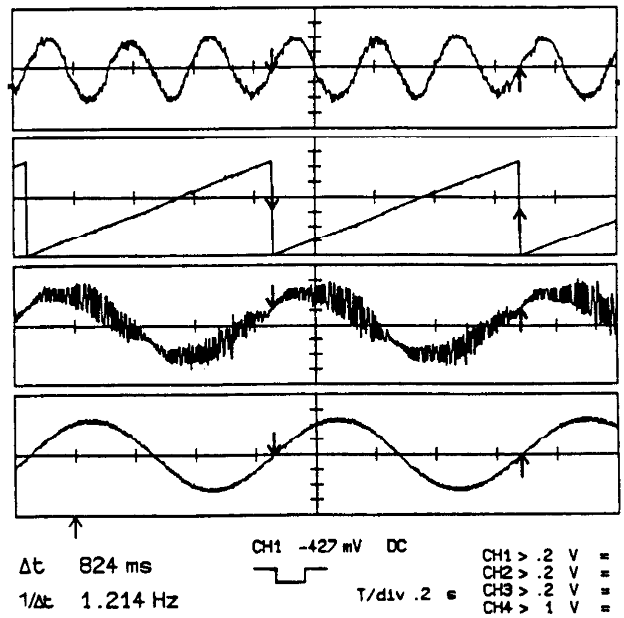

Fig. 7 Measured data at steady-state, low speeds of the IM: (ch.1) v3 signal; (ch.2) estimated field position; (ch.3) stator phase voltage; (d) stator phase current.

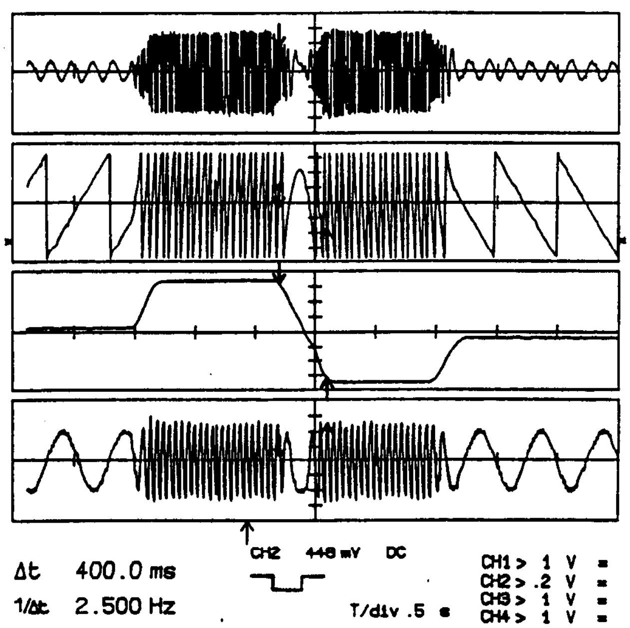

Fig. 8 Reversal of the machine: (ch.1) v3 signal; (ch.2) estimated field position; (ch.3) measured motor speed; (ch.4) stator phase current.

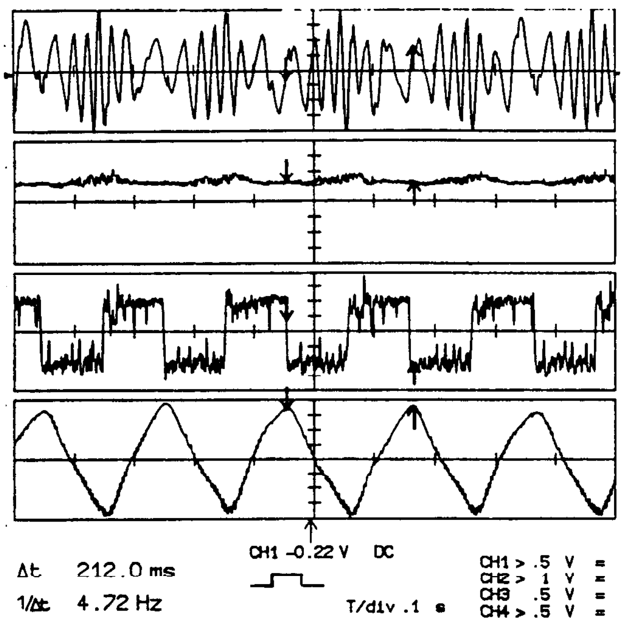

Fig. 9 Open loop speed operation, bang-bang i_q^*, constant i_d^*: (ch.1) v_3 signal; (ch.2) measured i_d current; (ch.3) measured i_q current; (ch.4) measured motor speed).

8. CONCLUSIONS

A simple and low cost scheme for direct field orientation control of an induction machine has been proposed. Low-cost speed transducers may be used, or a complete estimation scheme, including the estimation of the motor speed, may be implemented. 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 phase voltages are summed, the resultant signal contains a dominant third harmonic component followed by the rotor slot ripple which can be used for purposes of speed control. Utilization of this controller requires the stator be star connected with access to the neutral connection.

Further analysis and experimental work must clearly 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 performance limits. However, relative to the goal of this study, that of obtaining a cost-effective, simple and reliable measurement and control structure, one may conclude that this goal is possible with the approach of this paper.

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