

Direct Field Orientation Controller Using the Stator Phase Voltage Third Harmonic

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Abstract—This paper presents the development and implementation of a direct field orientation controller (DFOC) 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. The control utilizes 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 and a high frequency component due to the rotor slot ripple. It is shown that the third harmonic voltage component can be effectively used to estimate both amplitude and position of the air gap flux. Two methods for estimation of the air gap flux from the third harmonic voltage are discussed in the paper. A complete induction motor direct field orientation control is designed and implemented in the laboratory. Extensive experimental results showing the DFOC drive system performance are presented and discussed.

I. INTRODUCTION

THE ADVANTAGES of direct field orientation over the indirect type in overcoming the controller sensitivity 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 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 has been demonstrated. A discussion of the 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 for parameter detuning in indirect field oriented control schemes [8, 9]. A new control principle that 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,

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a low cost medium performance controller can be realized without the need for explicit speed or position measurement in the torque loop. The basis for this work was established in [11] where two different methods to implement field orientation using the stator third harmonic voltage were proposed and simulated. In the present work, the authors propose a more direct way to estimate the air gap flux, eliminating the need for the integration of the third harmonic voltage component. Moreover, the method implemented here allows the estimation of the air gap flux position with greater accuracy and wider bandwidth.

The following sections describe: the origin of the saturation third harmonic stator voltage component, and how the air gap and rotor fundamental flux components are estimated from the third harmonic voltage signal; two ways of estimating the absolute location of the air gap flux, including estimation between third harmonic zero crossings; and the implementation and experimental results on an induction motor field-oriented drive system.

II. SATURATION HARMONIC COMPONENTS

The direct field orientation strategy proposed in this paper is based on detection of the third harmonic air gap voltage [5, 6]. In those 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 is 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 Fig. 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 a function of the rotor mechanical speed and can be utilized as

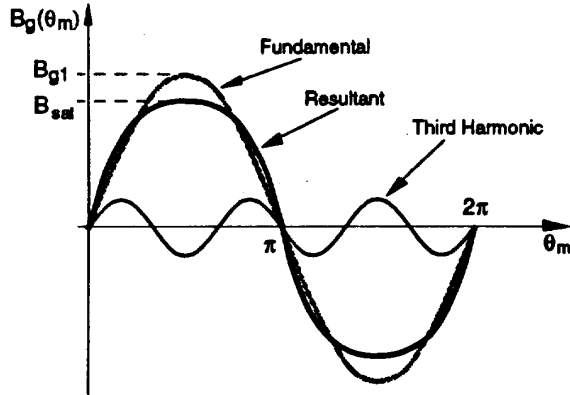


Fig. 1. Air gap flux density along the air gap with stator and rotor teeth undergoing saturation.

a means to measure the speed, thereby eliminating the need for a tachometer [4].

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 (5th, 7th, 11th, 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 cancelled and the resultant wave form contains mainly a third harmonic and 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} .

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

A. 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. Fig. 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

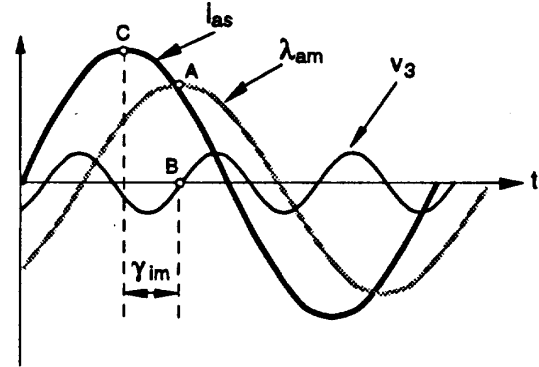


Fig. 2. Air gap flux linkage and its third harmonic component referred to one of the line currents.

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 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} , that 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.

B. Rotor Flux Orientation

Another possibility of 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.

$$\lambda_{qr}^s = \frac{L_r}{L_m} \lambda_{qm}^s - L_{lr} i_{qs}^s \quad (1)$$

$$\lambda_{dr}^s = \frac{L_r}{L_m} \lambda_{dm}^s - L_{lr} i_{ds}^s \quad (2)$$

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].

C. 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 information.

Thus, a practical aspect related to the implementation of these schemes concerns the estimation of the amplitude of the fundamental air gap flux linkage from the amplitude of the third harmonic flux. A non-linear function relating the amplitudes of fundamental and third harmonic air gap flux components is derived from the conventional induction motor no-load test and the results store in a look-up table for controller access. This function is machine dependent and requires supplementary experimental and/or computational effort.

In addition, experiments made at low speeds reveal that the signal/noise ratio for the third harmonic voltage signal decreases considerably. The same happens when the motor operates at a lower than rated flux level since the amplitude of the signal reduces due to the lower degree of saturation. Hence, an accurate estimation of the flux in these conditions requires a well designed variable filtering network that can be implemented via hardware and/or software.

Another implementation aspect to consider is the integration of the third harmonic signal in order to obtain the flux component λ_3 . A non-ideal integration scheme has to be implemented what becomes a problem during transient regimes leading to inaccuracies in the flux estimation. 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 .

As mentioned earlier, at low speeds and lower than rated flux levels the signal/noise ratio of v_3 decreases considerably. Nevertheless, if properly filtered and averaged the signal still contains useful information. This information is derived from the zero crossings of v_3 , that may be used in order to estimate the position of the third harmonic voltage and consequently 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 in the third harmonic voltage and flux positions. Thus, an exact position information is available in a very convenient manner.

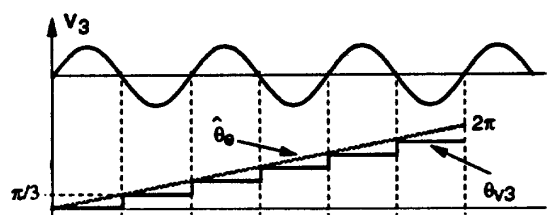


Fig. 3. Position detection from zero crossings of the v_3 signal.

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 .

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

IV. ABSOLUTE POSITION LOCATION OF AIR GAP FLUX

Two different means to absolutely locate the position of the flux from the v_3 signal for a three phase ac motor have been studied and tested. Both give good results, each having different advantages without critical drawbacks.

A. Use of the Stator Phase Voltages

The first method uses the relative position of the stator phase voltages to detect the absolute position of v_3 . If one does not consider for the moment any voltage drop across the stator impedances, 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 voltage drops for each stator phase. This means that the currents must be measured, an estimate of stator impedance must be found, and finally the estimated induced phase voltage can be obtained. One may obtain an estimate for the stator impedance 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 the stator resistance due to temperature changes, for instance. After correcting the stator impedance drop 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 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_{v_3} having more than one $\pi/3$ radians step between two consecutive zero

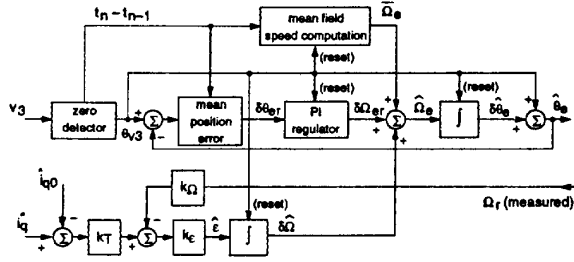


Fig. 4. Estimation scheme based on the mean speed of the field.

crossings of v_3 . However, this solution is similar to the next, simpler approach.

B. Use of a Relative Incremental Scheme

The second method is simpler similar to relative incremental transducers without an index signal. At starting, 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.

This approach is much easier to implement than the previous one since it does not require the implementation of voltages and currents measurements nor stator impedance voltage drop corrections. On the other hand, the main drawback is that of any relative transducer. That is, any parasitic zero crossing of v_3 leads to an unacceptable position error, which is 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 the machine as well as numerous other transient regimes were also tested. These test results showed that even at low speeds when the third harmonic voltage amplitude signal is small, its zero crossings still give a correct and useful information about the absolute position of the air gap flux of the machine.

V. ESTIMATION OF THE POSITION BETWEEN THE ZERO CROSSINGS OF V_3

As mentioned, the estimated position information θ_{v3} in Fig. 4, obtained from the zero crossings of v_3 , gives a coarse position information, with only six steps over a full period with $\pi/3$ radians accuracy. Only at the 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.

A. 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 , the position of the stator voltage vector is computed in a stator reference, and used to realize the position estimate. Due to the fact that several error sources affect the measured

stator voltages, mainly at low speeds (impedance voltage drop, harmonic contents, measurement noises), a better approach is to use the variations of the position of the stator voltage vector in order to compute the estimated position.

Due to the same problems related with the stator phase voltages at low speeds, a limit speed is reached below which the system fails to operate. In order to reach as low speed as possible the method requires voltages and currents measurement, stator impedance-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.

B. Use of Position Estimation Schemes

An alternative approach is to implement a complete position estimation scheme based on the absolute position estimate and on the dynamic model of the system (motor and 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_{n-1} and t_n , the mean speed on the time interval (t_{n-1}, t_n) is

$$\bar{\Omega}_n = \frac{\theta_{s3}}{t_n - t_{n-1}} \quad (3)$$

where $\theta_{s3} = \pi/3$ is the position step increment between two consecutive zero crossings of v_3 . The estimation scheme works well in steady-state regime, but unfortunately provides poor estimates during fast transients of the system since no acceleration information is available. Also, the estimate mean speed corresponds to the previous v_3 pulse and thus a dead time is present in the mean speed estimate.

The scheme can be improved by introducing corrections which take into account the motor and load models. The result is a good position estimation, 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. Fig. 4 presents the final estimator.

One correction term, $\delta\hat{\Omega}$ 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 term \hat{i}_{q0} .

A second correction term, $\delta\hat{\Omega}_{er}$, is computed as the output of a proportional-integral (PI) regulator, having as input the mean position error detected at v_3 zero crossings, (at time instant t_n). This term represents the correction needed for the load torque variations and/or modeling errors in the computation of the $\delta\hat{\Omega}$ 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 that case, one should pay particular attention to the parameter sensitivity of the estimation scheme. Problems may

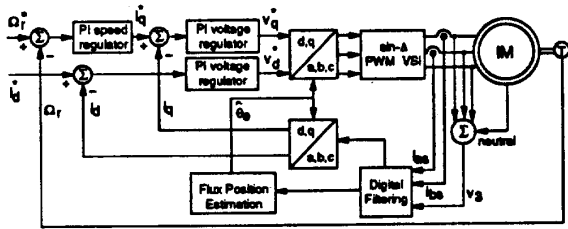


Fig. 5. The direct control scheme for the IM using the third harmonic voltage signal.

be encountered due to this fact, and further investigations must be done to validate such scheme.

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

As stated in the preceding paragraphs, the air gap flux 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 speed information. A low cost transducer (tachogenerator), or only a speed estimation scheme based on the machine model, may be considered. The high frequency slot ripple component in the third harmonic voltage signal v_3 can also be used to measure the motor speed as in [4]. The first approach was considered for the experimentally implemented scheme.

The control scheme implemented is presented in Fig. 5. In this figure the torque control variable i_q^* is obtained from the speed error via 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 reference voltage commands generated by the PI voltage regulators. As mentioned earlier, the current command i_d^* must be obtained using a decoupling network to insure decoupling between the d - q variables and consequently constant air gap flux. Considering the medium performance requirements accepted for this study, only a constant value for i_d^* was considered for the experimental tests.

VII. EXPERIMENTAL RESULTS

Experimental tests were carried out 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 IM. A Motorola 56001 digital signal processor (DSP) based digital system was used to implement the digital control scheme. Twelve bit accuracy A/D and D/A converters

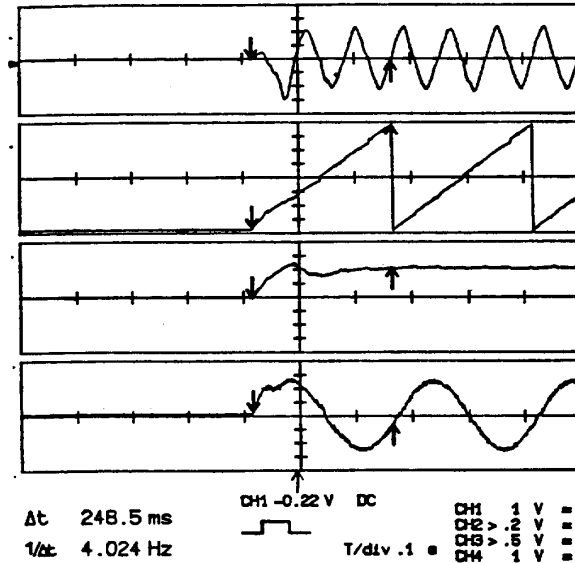


Fig. 6. Starting of the test induction motor. From top to bottom: v_3 signal, estimated field position, measured motor speed, and stator phase current.

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. Special care was paid to a proper digital processing of the signals. A 6-th order Butterworth low pass filter was used for the fundamental and third harmonic signals; 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. Extra averaging procedures were also implemented at low speeds in order to improve the v_3 signal. A special adaptive locking procedure for the zero crossings detection was also needed at low speeds, in order to eliminate false zero crossing detections.

Good results were obtained with the proposed estimation and control schemes. Figs. 6 to 9 present some experimental data obtained for the tested induction motor. For all experimental plots in this section the traces from top to bottom correspond to channel 1 to channel 4. The time and vertical scales are indicated at the bottom of the plots. The plot in Fig. 6 depicts the start of the machine. In this case, an initial pre-magnetization 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. Experiments show that a good starting performance is obtained. A certain sensitivity of the controller to the load parameters is to be expected, which is more evident at starting transient. Further studies must be done in this direction to optimize performance.

Fig. 7 presents a steady-state operation at low speed (1.2 Hz electrical). As one can see, 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 shown.

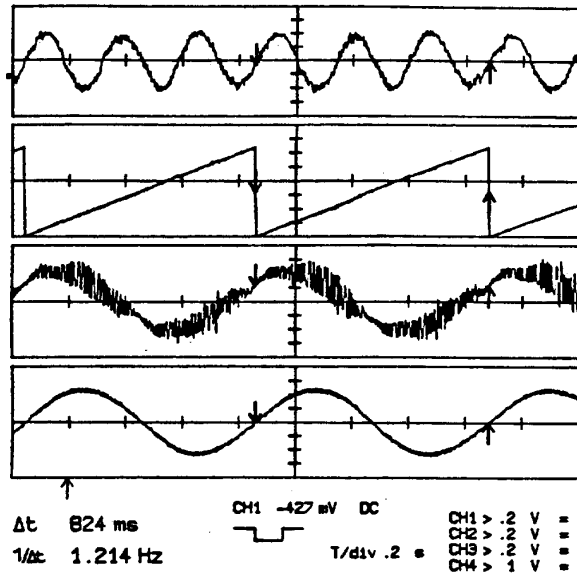


Fig. 7. Steady state operation at low speed. From top to bottom: v_3 signal, estimated field position, stator phase voltage, and stator phase current.

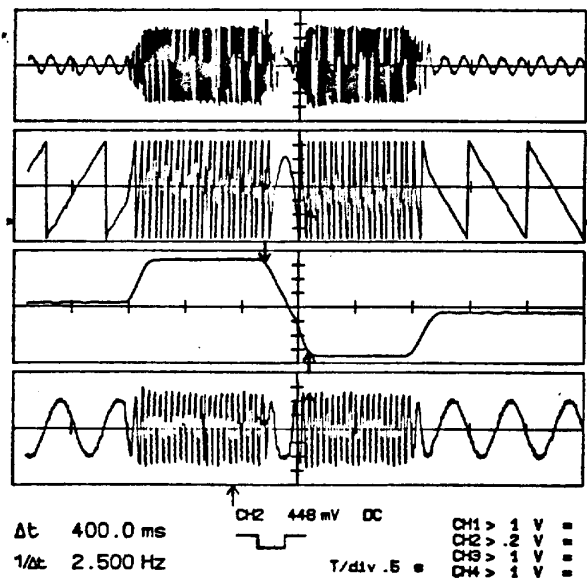


Fig. 8. Speed reversal transient. From top to bottom: v_3 signal, estimated field position, measured motor speed, and stator phase current.

Fig. 8 presents a reversal of the motor. The main problem is the region around the zero speed (very low speeds) when extra filtering is necessary for the third harmonic voltage signal. If the motor speed only cross this region, the position estimation scheme as presented in this work functions satisfactorily and no extra filtering is necessary.

Open speed loop tests were performed in order to verify the correctness of the position estimate, and the decoupling of the two current components. The reference for the torque component i_q^* is defined as a square wave and Fig. 9 shows the results obtained in this case. The figure shows from top to

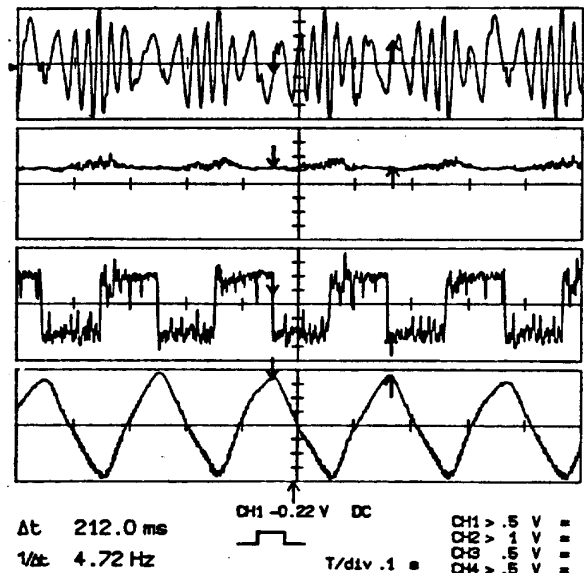


Fig. 9. Open loop operation with constant i_d^* . From top to bottom: v_3 signal, measured i_d current components, measured i_q^* current component, and measured motor speed.

bottom the v_3 signal, measured i_d current, measured i_q current and measured motor speed. The variation of the speed slopes or acceleration are evident. It is a consequence of the non-total decoupling and also possibly due to differences in the static friction torque.

VIII. CONCLUSION

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 via the slot ripple component in the third harmonic voltage signal may be implemented. The method is based on the concept of locating the fundamental component of the air gap flux from the third harmonic voltage component induced in the stator voltages when the machine operates in a 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 a star connected stator winding with access to the neutral connection.

Further analysis and experimental work must clearly be done in order to completely validate the proposed estimation scheme, to study its sensitivity to the motor and load parameters, and to find its 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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Thomas A. Lipo For photograph and biography see page 432 of this issue.