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DSP BASED SPEED SENSORLESS CONTROL USING INDUCTION MOTOR

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ABSTRACT

Today, speed sensor less modes of operation are becoming standard solutions in the area of electric drives. The proposed algorithm has been implemented using a floating-point digital signal processor (DSP). Speed sensor-less control techniques can make the hardware simple and improve the reliability of the motor without the introducing the feedback sensor and it becomes more important in the modern AC servo drive. Computational Elements have been integrated on a single board SH65L type and interfaced to the power electronic converter, and the use of proper DSP optimizes the cost and computational properties. The novelty of the presented solution is the integration of a simple observer for both speed/flux and current control purposes, and the obtained results have been improved in comparison to the previous works. An overview of the test bench consisting of a digital control board, as well as computational algorithms and system benchmarks, is presented.

Keywords: Adjustable-speed drive, Analog to digital, Electromotive force. Induction motor. Field-programmable gate array, Field-oriented control, Digital signal processor, Predictive current controller, Personal computer.

Introduction

Adjustable electric drives are widely used in industry, with a typical ASD consisting of an IM and a power electronic converter, e.g., a voltage inverter. In most cases, the motor speed is commanded so that the control system requires an actual speed signal for closing the speed loop. In industrial applications, speed sensors, as well as sensor less solution, are used. Due to the possibility of existing speed sensor noises and for maintenance and economic aspects, the trend is to substitute speed sensors by computational solutions. Comprehensive reviews of the sensor less drives are given in [1], which shows that there are still some persistent problems associated with the sensor less control, indicating that new solutions are still needed [2]–[6]. In ASDs, the mature control approach of IM is the FOC method.
which is widely used in modern industrial drives. The integral part of numerous FOC systems is the stator current controller. Different methods for current control are presented in the literature, and a comprehensive survey is given in [8] and [9]. In the classical FOC solution, PI or hysteresis controllers are generally used. However, PCCs are reported to have better properties [10]–[15].

Another problem in electrical drives is system sensitivity to inaccuracy and changes of motor equivalent circuit parameters. Most of the FOC systems are very sensitive to such inaccuracies; therefore, some parameters should be estimated online, and a robust structure of the control is required. The real-time implementation of the system is an important task even for electric drives or power converters. In this paper, the real-time system of the electric drive is presented. The system has PCC implemented in the IM speed sensor less system with a FOC method. In the proposed scheme, FOC is used along with PCC and space vector PWM to ensure a constant switching frequency. The predictive current control algorithm previously presented modified by using an observer system instead of a simple load model. The use of the observer presented avoids the problems associated with system start-up indicated. To simplify the control algorithm while increasing the computational precision, the proposed PCC is combined with the speed/flux observer for a FOC IM drive. The proposed system requires a proper control board that ensures good computational power and reliable efficiency. In such solutions, FPGA and DSP are used, and the use of both elements ensures a system flexibility that can be easily interfaced with peripheral circuits and power converters.

![Figure 1. Speed sensor less ASD with IM and PCC.](image)

In the proposed control system, the next tasks have been implemented: PWM, PCC, FOC controllers, speed/flux observer, data acquisition, and communication with a PC. In contrast to other solutions, the same observer is used for realizing the PCC to estimate the necessity of additional computing of the motor EMF required by the control structure. This helped in reducing the needed computational time. The proposed sensor less control scheme of the IM system is simple and robust and
can operate with a very wide speed range, including extremely low and very high speeds. It is an incredible challenge to find one scheme for all possible speeds and operating points. Additionally, the solution is robust against motor parameter variation.

Proposed System
A. Basic Scheme

The structure of the proposed system is shown in Fig. 1. The IM is supplied from a frequency converter, which consists of a diode rectifier and a transistorized voltage inverter.

The FOC is used for a decoupled motor speed and flux control with PI controllers. Instead of PI or hysteresis stator current controllers; the PCC is used in the proposed scheme.

The motor current, transformed from the dq to the αβ coordinates, is controlled by PCC. Simultaneously, the PCC cooperates with the PWM, which assures a constant switching frequency of the inverter.

The inverter with PWM and PCC works as a controlled current source. The system works without a speed sensor, while only the inverter input voltage and output currents are measured.

Other variables required by the control system are calculated in a closed-loop observer system.

B. Estimation System

The rotor flux \( \psi_r \), speed \( \omega_r \), and EMF are estimated based on the voltage model of the IM with a combination of algebraic equations linking fluxes and currents [1] which was presented by authors in previous works to obtain two different observers. In this proposed sensor less drive, the observer presented by the authors in implemented. In the solution, it is also used for EMF calculation required in the Current controller (see Section II-C).

The observer equations
\[
\frac{d\hat{\psi}_s}{dt} = -k_r \hat{\psi}_r + u_s / \tau_s - k_ab \left( i_a - i_s \right) \tag{1}
\]
\[
\frac{d\hat{\psi}_r}{dt} = u_r - k_r \hat{\psi}_r \tag{2}
\]
\[
\hat{\psi}_r = \frac{(\hat{\psi}_s^{(1)} - \sigma L_s i_s)}{k_r} \tag{3}
\]
where \( \tau_s = \sigma L_s / R_s \) is the stator time constant; \( k_r = Lm / Lr \) is the rotor coupling factor; \( \sigma \) is the leakage coefficient; \( R_r, R_s, Lr, Ls, \) and \( Lm \) are motor parameters; and \( k_ab \) is the gain. The rotor flux \( \psi_r \) is calculated using two models simultaneously, i.e., model I \( (\psi_s, \psi_r) \) and model II \( (\psi_s, i_s) \), and the stator flux is calculated in model I, whereas the rotor flux is from model II. The estimated stator current \( i_s \), appearing in (3), is
\[
i_a = \frac{(\hat{\psi}_s^{(1)} - k_r \hat{\psi}_r)}{\sigma L_s} \tag{4}
\]
Fig. 2 shows the observer structure. The details of the observer structure, as well discussion on the \( k_ab \) selection, can be found in [25].

The flux magnitude and angle position are
\[
|\hat{\psi}_r| = \sqrt{\psi_{ro}^2 + \psi_{r\beta}^2} \tag{5}
\]
\[
\hat{\beta}_{\psi_r} = \arctan \left( \frac{\psi_{r\beta}}{\psi_{ro}} \right) \tag{6}
\]
The estimated motor mechanical speed signal
\[
\hat{\omega}_r = \hat{\omega}_{\psi_r} - \hat{\omega}_2 \tag{7}
\]
In (9), the slip relation contains the magnitude of the flux in the denominator. In the computation, the equation is secured from division by zero. The flux is zero only at the initialization and starting of the machine, so some calculation error will appear only at that time instant. The flux magnitude, position, and speed estimation structure is shown in Fig. 3.

The advantage of the observer flux (1)–(4) is that no information on rotor speed is required because the speed is estimated separately by the (5)–(9) relations, so speed estimation accuracy does not affect the accuracy of the flux observer, which eliminates errors associated with the computation or measurements, giving the estimation system the robustness presented in Section III.

C. PCC

The basic structure of the PCC implemented in the IM drive was previously presented, where a more detailed idea is explained. In this section, the PCC is presented shortly to underline only the difference between the previous and the proposed solution. The specification of the presented PCC as predictive comes from the fact that some of the variables in the controller algorithm are predicted. The cost function of the controller is classical—it is current regulation error.

![Figure 4. Notation in PCC for switching periods](image)

The relations of the PCC are based on the simplified model of the IM. For PCC derivation purposes, the IM was modelled as an inductance, and the EMF was connected in series, while
The small motor resistance was neglected. In the EMF was calculated using a simple equation of the IM model. The accuracy of the EMF calculation could be improved if a more detailed load model is used in PCC, e.g., when using the flux and speed closed-loop observer presented in the previous Section. The observer structure is extended in order to calculate the EMF simultaneously with the flux and speed computation. Therefore, in the PCC, the EMF calculation part is removed and substituted by the signals obtained from the observer system.

The stator current dynamic system is described by

\[
\frac{d}{d\tau} \begin{bmatrix} i_s \\ e \end{bmatrix} = \begin{bmatrix} u_{\text{com}}^i \nabla & e \end{bmatrix}.
\]  

(10)

Assuming the notation shown in Fig. 4 and for small \( T_{\text{imp}} \), it is possible to convert (10) to the next discrete form

\[
\frac{1}{T_{\text{imp}}} \begin{bmatrix} i_s(k-1) - i_s(k-2) \\ e(k-1) - e(k-2) \end{bmatrix} = \begin{bmatrix} u_{\text{com}}^i(k-1) - e(k-1) \nabla e(k-2) \end{bmatrix}.
\]  

(11)

Considering (11) for the period \( (k-1) \) to \( (k) \), the known values are the commanded voltage \( u_{\text{com}}^i(k-1) \) and the measured current \( i_s(k-1) \). Other variables such as \( i_s(k) \) and \( e(k-1) \) are unknown and should be predicted. In , the EMF value \( e(k-1) \) was simply predicted based on known samples of \( e(k-2) \) and \( e(k-3) \) as follows:

\[
\hat{e}(k-2) = \tau L_{i_s} (i_s(k-2) - i_s(k-1)) / T_{\text{imp}} + u_{\text{com}}^i(k-2)
\]  

(12)

\[
\hat{e}(k-3) = \tau L_{i_s} (i_s(k-3) - i_s(k-2)) / T_{\text{imp}} + u_{\text{com}}^i(k-3)
\]  

(13)

In this paper, unlike (12) and (13), the EMF is calculated in the flux and speed observer procedure as follows:

\[
\hat{e} = \frac{d\psi}{d\tau}.
\]  

(14)

The use of (14) is the benefit of the presented PCC—it minimally decreases the computational requirements for real systems because (12) and (13) are not needed. In the proposed speed sensor less drive, the calculation of (14) is an integral part of the used observer, i.e., in (2), therefore, no additional observer calculations are required.

The samples of \( \hat{e}(k-2) \) and \( \hat{e}(k-3) \) calculated by the observer are memorized and used in the successive calculations.

The change of position of the EMF vector is

\[
\Delta \varphi = (k-2) \Delta \varphi = \varphi(k-2) - \varphi(k-3)
\]  

(15)

The calculation of (15) requires two arc tangent calculations for obtaining \( \varphi(k-2) \) and \( \varphi(k-3) \). To simplify the calculation of \( \Delta \varphi \), the base mathematics relations of tangent functions for sum and difference of two angles are taken into account, and finally, the next relation is used

\[
\Delta \varphi = \arctan \left( \frac{\hat{e}(k-2) \hat{e}(k-3) + \hat{e}(k-3) \hat{e}(k-2)}{\hat{e}(k-2) \hat{e}(k-3) - \hat{e}(k-3) \hat{e}(k-2)} \right)
\]  

(16)

The controller structure is shown in Fig. 5.
The controller coefficient $W_{curr}$ was tuned to minimize the current regulation error. For that purpose, a lot of experimental tests were done for different $W_{1curr}$ values. The results are shown in Fig. 6.

In Fig. 6, the results were taken from the tests in which the current controller was working with FOC and a sensor less observer. The current regulation was tested for different motor speeds: 0.1, 0.5, and 0.8. Also, the influence of the load torque $T_L$ was tested and similar behaviours for $\omega_r = 0.1, 0.5,$ and $0.8$ were observed. It is noticeable that the motor and load torque speed has a minimal effect on the current controller operation; the gain selection close to 1.0 gives the minimal regulation error for different operating points. Fig. 7 shows an example of the PCC test results. In Fig. 7, one can find the results for both steady state and transients with PCC operation. Noticeably, in transients, the current regulation error is close to the error in steady state, and one can note the proper function of the PCC. The system responded to motor speed changes—decreasing and increasing. Additionally, the waveforms of the current present consistency.
between the commanded and actual current components. The base frequency of the current regulation error is the same as the frequency of the stator current because the controller operates in the stationary reference frame, so even small phase regulation error exists [8]. However, the real current follows the commanded one with a small error of less than 3%. In comparison to the authors’ previous solution, the results have improved.

**Investigations**

**Test Bench**

The structure and photograph of the test bench are shown in Fig. 8. In the experimental test bench, two identical 5.5-kW machines, the FSg132S-4, were used. Motor M was supplied by an MMB010-type voltage inverter with an open-source control board SH65L, and generator G was working with an industrial converter type 690+PB0055. The generator converter is working in a torque closed-loop control, with the desired load torque set by the test bench operator. Both machine inverters were connected with a dc link so that, in steady state, the whole test bench consumed little electrical energy, using only enough to cover the losses in the inverters and machines. The test bench parameters are presented in Table I.

**Control System**

The proposed control structure was implemented in an SH65L-type microprocessor system and an interface board. In the SH65L control system, the floating-point ADSP-21065L processor and the FPGA EPF6016 circuit, in addition to some auxiliary circuits, were used. The interface board contains an A/D converter, general-purpose I/O pins, and drivers for power converter transistors. The block structure of the control system boards is shown in Fig. 9. The ADSP-21065L is a low-cost general-purpose 32-b floating-point DSP and has 198 MOPS and configurable 544 kb of on-chip static RAM. The DSP uses a 30-MHz input clock. The clock frequency is equal to half the instruction rate—15 ns—so it is equivalent to 60 MHz. The DSP has a host processor interface (HPI) that can be easily used for superior communication. The HPI mechanism is particularly useful in the tests of the new control algorithm. The software code could be written in both Assembler and C languages with the help of the Visual DSP compiler.

The specific functions needed for power converter control require advanced PWM, timers, and A/D conversions. To extend the control ability, the DSP was interfaced with an FPGA Flex 6016 and an A/D converter AD7864. The EPF6016 is an FPGA Flex 6000 family programmable circuit. The FPGA architecture is based on logic arrays that contain LABs composed of ten LEs communicating through a local interconnecting structure. The EPF6016 has 1320 Les that respond to 16 000 typical gates and 204 programmable I/O pins. The AD7864 is a four-channel simultaneous-sampling high speed 12-b A/D converter [38], with the conversion time for one channel being 1.65 μs. It has a configurable input range, parallel interface to the processors, one 5-V supply voltage, and very low power consumption (90 mW). The A/D conversion is dedicated specifically to ac motor control and power electronic converters. The selection of specific DSP, FPGA, and A/D converter was determined by economic and performance aspects.
The obtained system has very high computational power as could be noticed in the results in Section III-D.

**Control Algorithm**

The control algorithm is divided into two parts: software and hardware. The software procedures are performed in DSP, and the hardware procedures are performed in FPGA. The cooperation between DSP and FPGA provides a flexible control structure. The DSP is working in the software loop and interrupts the system. In the time-independent software loop, the communication with PC is performed. The loop is interrupted every 150 μs to provide the essential control procedures, which are strictly time dependent. Under each interruption, the following procedures are performed:

1. Measurement of signal recalculations and transformations;
2. Calculations of speed/flux/EMF observers;
3. FOC structure with PI controllers;
4. Predictive current control;
5. Auxiliary I/O control.

The use of the FPGA system makes it possible to realize parts of a control system by hardware, which releases the processor from some tasks. An FPGA in the experimental setup realizes the following functions:

1. Times the switching-on of each transistor for one switching period;
2. Provides a dead time;
3. Controls the breaking transistor;
4. Services the A/D converters;

For an ac drive, it is essential to synchronize the control algorithm and the PWM calculations with the sampling of the measurement signals. This allows for the maintenance of the same sampling frequency for the A/D converter with the control algorithm calculation steps. The PWM system operates with a switching period of $T_{imp} = 150 \mu s$, which consists of four cycles: one-half of the time of the passive vector duration $t_0$, the first active vector $t_1$, the second active vector $t_2$, and one-half of the zero vector period $t_0$. The PWM operation is approximated in Fig. 10.

The A/D converter is sampled with the switching frequency. The EOC signal generates interruptions for the DSP. Next, the DSP performs complete calculations for one step of the control algorithm and sends the new values for PWM to the FPGA circuit. Writing new times to the FPGA has to be realized before finishing an actual switching period.

![Figure 9. Structure of the control boards.](image-url)
Real-Time Algorithm Implementation

The realization of the presented system is strictly dependent on the possibility of the real-time implementation. It is obvious that each step of the control algorithm calculation has to be less than the assumed switching frequency. In Table II, the number of times that the whole algorithm was executed, as well as some basic procedures, is presented. The benchmarks given in Table II have been measured by a digital oscilloscope after the observation of the DSP programmable output signal. The used observer processing time was compared with that of previously used observers. Due to the time, it is comparable with an estimator and better than a Luenberger-type observer used in the system presented. One can see that, despite the control algorithm complexity, the total calculation time is close to 40% of the switching period. This allows for an increase of the PWM frequency or for an extension of the calculation procedures in the future work.

Experimental Results

The discussed algorithm has been tested experimentally with different speeds and loads. Some representative results are shown in Figs. 11–16. All the waveforms are presented in per unit related to the base...
values: voltage of 400 V, current of 19 A, speed of 1500 r/min, torque of 36 N · m, and flux of 1.3 Wb.

In the experiments, we carried out tests to compare the observer presented in Section II-C with a previous version presented by the authors. No significant differences were observed in the drive steady state and transients; it was, however, noticed that, if the observer from Section II-C is implemented, the drive is more robust due to changes in the leakage inductance. The nominal value of $L_σ$ is 0.089 p.u.; with this solution, the minimal acceptable level was 0.001 p.u., while in the drive with the observer, only 0.33 p.u. was obtained, which indicates the better properties of the observer structure presented in this paper. Due to the embedded drive robustness, no additional procedures are required for parameter online estimation at different speeds, including extremely low and high speeds; hence, the proposed solution seems applicable for real-time implementation.

**Scope of the Study**

The requirements modern electric drives can be satisfied if a control scheme with good performance is used. The trend of speed sensor elimination requires some complex calculation and advanced control solutions. Operation in real time needs high computational power of the microprocessor system.

**Conclusion**

The system selection is dependent on economic and computational requirements. The proposed system provides simplicity and good control performances while keeping the robustness against the motor parameter changes without using online parameter estimation. A low-cost control system satisfies the system requirements by proper DSP and FPGA selection. For that reason, the proposed solution is appropriate for the industrial applications.

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**References**


