High Dynamic Speed Sensorless AC Drive with On-Line Parameter Tuning and Steady State Accuracy

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Abstract — Controlled speed sensorless ac drives have reached a stage of development permitting good dynamic performance above 3% of rated speed. However, the accuracy of the rotor speed estimation under load remains sensitive to parameter errors of the internal machine model. This paper presents an approach that ensures high steady-state speed accuracy in addition to high dynamic performance. To eliminate the speed estimation error, the machine parameters are adapted on-line, based on the evaluation of rotor slot harmonic effects. A stator flux oriented control scheme is implemented in a digital signal processor system to demonstrate the robustness of the speed estimation to parameter variations. Experimental results demonstrate that the control system advantageously combines high dynamic performance with accuracy of speed estimation.

1. INTRODUCTION

Ac machine drives based on field oriented control operate at high dynamic performance and good speed accuracy, provided a speed sensor is used on the machine shaft. Such drives are widely employed in different kinds of industrial applications. During the past few years, a strong interest has developed to eliminate the speed sensor, while maintaining the performance of the control system unaltered. The advantages of speed sensorless ac drives are lower cost, reduced size of the machine set, elimination of the sensor cable, and increased reliability.

A large variety of different schemes for speed sensorless ac drives have been proposed [1-7, 9, 10]. Depending on the respective approach, very good dynamic performance can be achieved over a fairly large speed range. However, the speed accuracy is generally sensitive to model parameter mismatch, especially in the low speed range and under load.

A preferred method for speed sensorless ac drives uses stator field oriented control [2, 3]. In this scheme, the stator flux space vector is estimated from measured machine terminal quantities to provide the field transformation angle. A dynamic decoupling system operates in a feedforward mode to enable independent control of machine torque and flux [2].

The flux estimation accuracy is fairly good above a few Hz stator frequency. However, variations of machine parameters have a strong influence on the accuracy of the estimated mechanical rotor speed under load. Particularly at lower speed is the stator flux estimation sensitive to an inaccurate stator resistance value in the estimation model. Also the leakage inductance value, though not that decisive, should be properly matched to the actual leakage inductance of the machine. Inaccurate model parameters lead to misalignment of the field oriented coordinate system, which impairs the dynamic performance of the drive system. Possibly more important for many applications is the steady-state accuracy of the speed control, which is poor with detuned model parameters.

On the other hand, the method of speed estimation based on the evaluation of rotor slot harmonics does permit high speed accuracy in the steady-state, however at the expense of poor dynamic performance [9, 10].

This paper presents a novel solution for high-performance, high-accuracy sensorless ac drives. High dynamic performance is achieved using a machine model for fast estimation of the stator flux vector. The speed accuracy problem under load is widely eliminated by a parameter estimation and model adaption scheme which exploits the rotor slot harmonic effect.

2. STATOR FIELD ORIENTATION

2.1 Control structure

Fig. 1 shows the basic signal flow graph of a stator field oriented control system. The flow graph uses the complex notation which has been found expedient for the representation of ac machine systems [8, 1].

For the dynamic analysis, the induction machine is described in a synchronous reference frame, the real axis of

![Fig. 1: Block diagram of a speed sensorless ac drive system with stator field oriented control; the superscripts (F) and (S) refer to the respective coordinate systems.](image-url)
which is aligned with the stator flux vector. The following equations hold:

\[
\begin{align*}
\psi_s &= \psi_s + \frac{d\psi_s}{dt} + j\omega_s\psi_h \\
0 &= \tau_t i_t + \frac{d\psi_t}{dt} + j\omega_t\psi_t \\
\psi_s &= l_h i_s + l_t l_t \\
\psi_t &= l_h i_s + l_t l_t
\end{align*}
\]

where
\[
\begin{align*}
\omega_s &\quad \text{stator frequency,} \\
\omega_t &\quad \text{rotor frequency,} \\
l_h &\quad \text{stator inductance,} \\
l_t &\quad \text{stator inductance,} \\
l_h &\quad \text{mutual inductance.}
\end{align*}
\]

Note that time is a normalized quantity expressed as \(\tau = \omega_s R\), where \(\omega_s R\) is the rated stator frequency.

The existence of a fast current control system justifies assuming the stator currents as independently forced. The machine equations can then be simplified by substituting (3) and (4) into (2). This eliminates the rotor flux from (2).

The resulting equation is

\[
\tau_t \frac{d\psi_t}{dt} + \psi_t = -j\omega_t (\tau_t \psi_t - \tau'_t l_t i_s) + \tau'_t l_t \frac{di_s}{dt} + l_t i_s, \quad (5)
\]

where \(i_s\) is an independent variable, and

\[
\begin{align*}
\tau_t &= l_t / \tau_t & \text{rotor time constant,} \\
\tau'_t &= \sigma \tau_t & \text{transient rotor time constant,} \\
\sigma &= 1 - \frac{l_h^2}{l_s T_t} & \text{total leakage factor.}
\end{align*}
\]

Equation (5) is referred to in a synchronous reference frame. It is visualized in the upper portion of the signal flow graph Fig. 2. The lower portion represents the mechanical subsystem of the machine, [8]. \(T_e\) is the electromagnetic torque, and \(T_L\) is the load torque.

Fig. 2: Signal flow graph of the induction motor at stator field orientation, forced stator currents

Introducing a stator field oriented reference frame reduces the stator flux vector to its real component, \(\psi_s = \psi_s\). The imaginary component \(\psi_{sq}\) is then equal to zero. This condition is symbolized in the signal flow graph Fig. 2 by dotted lines representing the signals \(\psi_{sq} = 0\) and, following from that, \(d\psi_{sq}/d\tau = 0\). The first one of these equations, \(\psi_{sq} = 0\), establishes the condition that enforces stator flux orientation. The condition can be read from the balance of the imaginary components at the upper summing point in Fig. 2 as

\[
l_s \left( \tau_t \frac{d\psi_{sq}}{d\tau} + i_{sq} \right) = \omega_t \tau_t (\psi_{sq} - \sigma_l i_{sd}), \quad (6)
\]

taking into account that the imaginary output from the summing point is a zero signal.

On condition that stator field orientation exists, the dynamic machine structure Fig. 2 can be simplified as shown in the shaded area of Fig. 3. The diagram reveals that the torque-producing current \(i_{sq}\) produces an undesirable input \(-\omega_t \tau'_t i_{sq}\) to the stator flux channel. A compensating signal \(i_c\) is therefore derived from the torque current \(i_{sq}\) which is added in a feedforward fashion to the flux command current. The \(\tau'_t\)-delay in the compensation channel nullifies the differentiating element in the \(i_{sd}\)-input to the machine structure. If the machine parameters \(\tau'_t\) and \(\omega_t\) are accurately estimated, the compensation eliminates the cross-coupling between the torque producing current \(i_{sq}\) and the stator flux \(\psi_s\), [2].

Fig. 3: Machine control with stator field orientation using an external dynamic decoupler

2.2 Estimation of system variables

In the absence of a speed sensor, the stator flux vector \(\psi_s\) and the angular mechanical velocity \(\omega\) need to be estimated. The stator flux signal is generated by open integration of (1) in stationary coordinates, \(\omega_s = 0\), yielding

\[
\psi_s = \int (u_i - \tau_t i_s) d\tau. \quad (7)
\]

The corresponding signal flow is visualized in the upper left portion of Fig. 4.

The stator frequency can be expressed as

\[
\omega_s = \frac{d\theta}{d\tau} = \frac{d}{d\tau} \left( \tan^{-1} \left( \frac{\psi_{sq}}{\psi_{sd}} \right) \right) \quad (8)
\]

from which

\[
\omega_s = \frac{1}{\psi_s^2} \left[ \psi_s \frac{d\psi_s}{d\tau} \right]_z \quad (9)
\]

is obtained. Note that only the \(z\)-component of the vector
product is taken in (9). \( \delta \) is the stator field position angle. This angle is also used as the field transformation angle.

The angular mechanical velocity is defined as

\[
\dot{\omega} = \omega_s - \dot{\omega}_r. \tag{10}
\]

In this equation, \( \dot{\omega}_r \) is marked as an estimated quantity by the "\(^\approx\)"-symbol. While the stator frequency \( \omega_s \) in (10) can be estimated with good accuracy, the rotor frequency signal \( \dot{\omega}_r \) is prone to estimation errors.

An estimation for \( \dot{\omega}_r \) can be derived from the condition for field orientation (6). The equation is rewritten as

\[
\dot{\omega}_r = \frac{l_s}{\tau_r} \left( \frac{\sigma \tau_l \frac{di_{sq}}{dt} + i_{sq}}{\psi_{sd} - \sigma l_s i_{sd}} \right), \tag{11}
\]

where \( \tau_r = l_r/r_r \) and \( \sigma \tau_l = \tau_l^r \). The equation permits determining the rotor frequency from the stator current vector signal referred to in stator field coordinates. The graphic representation of (11) contributes the lower portion to the signal flow diagram Fig. 4. As the essential output, (10) provides the estimated speed signal \( \dot{\omega} \).

![Stator flux, speed and rotor frequency estimation](image)

Fig. 4: Stator flux, speed and rotor frequency estimation

### 3. ERROR ANALYSIS

Fig. 4 shows that the estimation of the stator flux vector \( \psi_s \) is sensitive to variations of the stator resistance \( r_s \). The stator resistance value can be easily determined by conducting a test at dc level during initialization [11]. Above a few Hz stator frequency, the resistive voltage is small compared with the induced machine voltage. Even the temperature dependent variations of the stator resistance can be neglected in this region. Hence, the stator flux vector can be accurately estimated, opening access to the good dynamic performance of stator field oriented control. However, the steady-state speed accuracy under load is poor.

To analyze this, the sources of the speed estimation error shall be assessed. The estimated angular velocity \( \dot{\omega} \) of the rotor is derived from (10) and (11). No major problem exists in determining the stator frequency \( \omega_s \) on the righthand side of (10). The value of \( \omega_s \) can be accurately derived as the angular velocity of the stator flux vector. This is shown in the upper portion of Fig. 4.

The estimation error of the rotor frequency \( \dot{\omega}_r \) depends on various machine parameters. (11) shows that the error magnitude increases when the machine is loaded, \( i_{sq} \neq 0 \). The machine parameters in this equation depend on the operating conditions. These vary in a fairly wide range during operation:

- The rotor resistance \( r_r \) changes with the machine temperature.
- Variations of the stator inductance \( l_s \) are caused by changes of magnetization. Such variations can be large and almost sudden, occurring predominantly during transitions to the field weakening region.
- The variation of the leakage inductance \( \sigma l_s \) is due to saturation of the stator teeth. The saturation is a function of the load current. The extent of change varies from machine to machine.

It is concluded from these observations that the speed accuracy deteriorates primarily because the rotor frequency estimation (11) is inaccurate.

### 4. PARAMETER ADAPTION

It is now proposed to eliminate the speed estimation error by on-line tuning of the model parameters. The adaption is based on a rotor speed signal \( \dot{\omega}_r \) which is accurately determined in the steady-state by exploiting the rotor slot harmonic effect. There are two different rotor speed signals used, one from stator flux estimator and the other from the rotor slot harmonic effect. The signals serve for the adaptive tuning of the model parameters so as to reduce the estimation error.

#### 4.1 SPEED ESTIMATION BASED ON ROTOR SLOT HARMONICS

An induction motor has generally slots on the surfaces of the stator and the rotor iron core to accommodate the windings. The rotor slots produce slot harmonics [9] in the airgap field, which modulate the stator flux linkage at a frequency proportional to the rotor speed. On condition that the number of rotor slots is not a multiple of three, which is true for the most machines, the induced rotor slot harmonics voltage per phase can be written as:

\[
u_{hr} = \hat{u}_{hr} \sin(N_r \omega + \phi_h) \tau \quad (12a)
\]

or

\[
u_{hr} = \hat{u}_{hr} \sin(N_r \omega - \phi_h) \tau \quad (12b)
\]

where

- \( \hat{u}_{hr} \) amplitude of rotor slot harmonics voltage
- \( \omega \) angular velocity of the rotor
- \( N_r \) number of rotor slots per pole pair
- \( 2p \) number of poles

The induced rotor slot harmonics voltages along with other triplen components \( \nu_h \) are separated from the much larger fundamental emf by summing the three phase voltages in a wye connected winding arrangement. Referring to the definitions in Fig. 5 we have:

\[
u_h = u_a + u_b + u_c \quad (13)
\]

as the zero sequence voltage, in which all nontriplen com-
components, including the fundamental, are absent while the rotor slot harmonics voltages and other zero sequence components contained in the phase voltages add up. Hence with a machine having \( N_r = 20 \) rotor slots

\[
u_{th} = u_{thr} \sin(N_r \omega + \omega_s) t + \cdots.
\] (14)

Fig. 6 shows an oscillogram of the sum \( u_{th} \) of the three measured phase voltages, taken from an induction motor having \( 2p = 4 \) poles and \( N_r = 20 \) rotor slots. The measured waveform contains rotor slot harmonics voltages and other harmonic components, predominantly a third harmonic which originates from machine saturation [10].

The rotor slot harmonic components exhibit the dominating frequency

\[
\omega_{sl} = N_r \omega + \omega_s = (N_r + 1) \omega,
\] (15)

They are extracted by a band-pass filter, the center frequency of which is adaptively tuned to the rotor slot harmonic frequency \( f_{sl} \). The filter transfer function is chosen as

\[
F(s) = \frac{\tau_{sl} s}{(\tau_{sl} s + 1)^2},
\] (16)

since complex roots would have caused numerical problems with a fixed-point processor.

The signal flow graph Fig. 7 shows the corresponding speed estimation scheme as implemented in software. The filtered signal \( u_{thr} \) is digitized by detecting its zero crossing instants \( t_z \). A software counter is incremented by one count at each zero crossing, thus computing the digitized rotor position angle \( \epsilon \). The rotor speed is calculated by digital differentiation, the same way as from an incremental encoder.

Fig. 8 shows the rotor slot harmonics signal after band-pass filtering. The scheme works accurately in the steady-state, but yields a poor estimation during fast transients of the system, especially at low speed, since there are only 42 counts per revolution of the rotor available.

The accuracy of the extracted speed signal shall be discussed considering the example of an induction motor having \( 2p = 4 \) and \( N_r = 20 \). The evaluation of the rotor slot harmonics offers a signal which is equivalent to that of an incremental speed sensor having a resolution of \( p(N_r + 1) = 42 \) counts per revolution. Since \( \omega_s = \omega + \omega_r \), the maximum stator frequency error is obtained from (15) as \( \varepsilon_{sl \max} = 0.0012 \text{ p.u.} \), assuming \( \omega_{R} = 2.5\% \).

Although of high accuracy, the slot harmonics speed signal is inappropriate for direct feedback in a high-performance speed control system. This is owed to its low-frequency ripple content originated by the low number of rotor slots. A distorted feedback signal would deteriorate the dynamic performance of the drive system.

4.2 Rotor frequency estimation

Generally, a field oriented ac-drive system comprises of a current control loop and a speed control loop. The current control loop is normally much faster than the speed control loop so as to provide the desired fast torque response. Since the speed loop reacts much slower, the derivative \( \frac{di_{sq}}{d\tau} \) that contributes to the rotor frequency calculation in (11) can be neglected. Equation (11) is therefore rewritten as

\[
\dot{\omega}_r = \frac{l_s}{\tau_e} \frac{i_{sq}}{\psi_{sd} - \sigma_{ls} i_{sd}}.
\] (17)

The estimation error depends on the respective values of the stator inductance \( l_s \), the rotor time constant \( \tau_e \), and the leakage inductance \( \sigma_{ls} \). Practical considerations permit the following simplifications:

- The leakage flux \( \sigma_{ls} i_{sd} \) is only a fraction of the stator flux \( \psi_{sd} \). Hence the term \( \psi_{sd} - \sigma_{ls} i_{sd} \) is fairly insensitive to leakage inductance variations.
• Since \( \frac{I_s}{I_r} = \frac{l_s}{l_r} \), the accuracy of rotor frequency estimation depends primarily on the rotor resistance. Accordingly, (17) converts to

\[
\dot{\omega}_r = \frac{i_{sq}}{\psi_{sd} - \sigma l_s i_{sd}}. \tag{18}
\]

The equation shows that, at given load and magnetization, the estimated rotor frequency \( \dot{\omega}_r \) is direct proportional to the rotor resistance.

4.3 Model parameter adaption

An error signal is used for rotor resistance adaptation which is derived from two different rotor frequency signals. A first rotor frequency signal is obtained as \( \omega_{rs} = \omega_s - \omega_{sl} \), where \( \omega_{rs} \) is supplied by rotor slot harmonics extraction in the arrangement of Fig. 7. This signal is accurate and serves as a reference. The second signal is the estimated rotor frequency generated by the machine model on the basis of (18). The difference between the two signals is the error indicator. Fig. 9 shows that the magnitudes of the two signals are taken. This avoids that the sign of the error signal \( \Delta \dot{\omega}_r \) inverts in the generator mode.

The error signal \( \Delta \dot{\omega}_r \) is then low-pass filtered to attenuate its ripple content. The filter time constant is chosen as high as \( T_f = \omega_{sl} r_1 = 0.7 \) s to eliminate dynamic error in \( \omega_{sl} \) during acceleration. The filtered signal feeds a PI-control, the output of which represents the model rotor resistance \( r_1 \) for rotor frequency estimation as per (18). The rotor frequency estimator is shown in the lower portion of Fig. 9.

Fig. 9 shows that the stator frequency signal \( \omega_s \) is obtained by numerical differentiation of the field transformation angle \( \delta \). This angle, in turn, is calculated from \( \mathbf{y}_s \) as per (8) using an \( \text{arctan} \) table. The procedure is more accurate and also numerically faster. The discretization error of the method in Fig. 4, (2), is proportional to \( s(\omega_s t_c) \), where \( t_c \) is the processor cycle time. For example, a 7% stator frequency error results at \( \omega_s = 3 \) and \( t_c = 400 \mu s \).

5. SOFTWARE IMPLEMENTATION

The control system was implemented in a TMS 320-C25 signal processor system. The software program consists of a main program which is executed every 114 \( \mu s \), and a fast interrupt service routine. The interrupt service routine activates a timer interrupt every 20 \( \mu s \), following which the rotor slot harmonics voltage is sampled and band-pass filtering is performed. The zero crossing point \( t_z \) is then detected. Other operations like speed estimation, parameter adaption, field orientation, as well as flux control, speed control and current control are completed by the main program.

6. EXPERIMENTAL RESULTS

The experimental verification of this fast and accurate sensorless speed control method was conducted in the laboratory. The drive system consists of a 7.5-kW induction motor being fed from a PWM inverter operated at 4 kHz switching frequency. A converter controlled dc machine serves as the load.

Fig. 10 and Fig. 11 show oscillograms that were recorded during speed reversal at higher speed and at low speed, respectively. The oscillograms also display accurate speed signals \( \omega \) which were measured by an incremental encoder to monitor the actual speed. This signal is not used in the control system. The two tracks in the center of Fig. 11 represent the components \( \psi_{sa} \) and \( \psi_{sb} \) of the estimated stator flux linkage vector. It can be observed from these signals that a positive offset error exists and that the software program limits the signals to unity maximum values.

![Fig. 10: Speed reversal at rated speed](image1)

![Fig. 11: Speed reversal at low speed](image2)
The sinusoidal waveforms get distorted which, through a ripple component in the field transformation angle $\delta$, affects the $i_{sq}$-signal as shown in the lowest track of Fig. 11.

A speed reversal from $-4500$ rpm to $+4500$ rpm which includes 1:3 field weakening is shown in Fig. 12.

The next oscillogram, Fig. 13, demonstrates that a speed signal derived from rotor slot harmonics is not suited to directly replace a tachometer signal in a high-performance drive. The slot harmonics signal exhibits a ripple content which acts as a disturbance on the speed controller. Fig. 13 was recorded from a drive system having a rotor field oriented control system.

Finally, Fig. 14 demonstrates the effectiveness of the rotor resistance adaptation at different speed settings for this test. The load torque was set to its rated value. There are considerable speed errors without rotor resistance adaptation. Note that the error is referred to the rated speed $\omega_{AR}$.

When the adaptation is activated, the speed errors reduce to less than 0.002 p.u. The overshoot of the $\omega^* = 2$ curve is a minor effect which is due to a lack of torque-gain adaptation at field weakening.

7. SUMMARY

State-of-the-art ac machine drives with speed sensorless control suffer from poor speed accuracy. The speed error is caused by fluctuations of the machine parameters, which change with varying machine temperature and with magnetic saturation. Although a speed measurement based on the rotor slot harmonic effect can provide the required accuracy, the ripple content of the extracted signal deteriorates the dynamic performance decisively.

Both a fast dynamic response and high steady-state speed accuracy are obtained when the rotor slot harmonic effect is utilized for parameter adaptation in a sensorless vector controlled drive system. The experiments show that the speed error is maintained below 0.2% throughout the base speed and field weakening range.

8. REFERENCES


