Sensorless Speed and Position Control of Induction Motor Drives

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Abstract — Controlled induction motor drives without mechanical speed sensors at the motor shaft have the attractions of low cost and high reliability. To replace the sensor, the information on the rotor speed is extracted from measured stator voltages and currents at the motor terminals. Vector controlled drives require estimating the magnitude and spatial orientation of the fundamental magnetic flux waves in the stator or in the rotor. Open loop estimators or closed loop observers are used for this purpose. They differ with respect to accuracy, robustness, and sensitivity against model parameter variations. Dynamic performance and steady-state speed accuracy in the low speed range can be achieved by exploiting anisotropic effects of the machine.

1. INTRODUCTION

AC drives based on full digital control have reached the status of a mature technology. The world market volume is about 12,000 millions US$ with an annual growth rate of 15%.

Ongoing research has concentrated on the elimination of the speed sensor at the machine shaft without deteriorating the dynamic performance of the drive control system [1]. Speed estimation is an issue of particular interest with induction motor drives where the mechanical speed of the rotor is generally different from the speed of the revolving magnetic field. The advantages of speed sensorless induction motor drives are reduced hardware complexity and lower cost, reduced size of the drive machine, elimination of the sensor cable, better noise immunity, increased reliability and less maintenance requirements. The operation in hostile environments mostly requires a motor without speed sensor.

A variety of different solutions for sensorless ac drives have been proposed in the past few years. Their merits and limits are reviewed based on a survey of the related literature.

2. CONSTANT VOLTS-PER-HERTZ CONTROL

To avoid the excitation at the eigenfrequencies of induction machines, a gradient limiter reduces the bandwidth of the command signal as shown in Fig. 1. The v/f characteristic ensures that the stator flux is maintained at its nominal value. A preset minimum value of the stator voltage accounts for the resistive stator voltage drop.

Since v/f-controlled drives operate purely as feedforward systems, the mechanical speed may \( \omega \) differ from the reference speed \( \omega_s^* \). A load current dependent slip compensation scheme can be employed to reduce the speed error.

Constant volts-per-hertz control ensures robustness at the expense of reduced dynamic performance, which is adequate for applications like pump and fan drives, and for low-cost drives. These contribute a substantial share of the market for sensorless ac drives.

3. ROTOR FIELD ORIENTATION

3.1 Model reference adaptive system

The model reference approach (MRAS) makes use of the redundancy of two machine models of different structures that estimate the same state variable on the basis of different sets of input variables [2]. The stator model in the upper portion of Fig. 2 serves as a reference model, the rotor model in the lower portion as the adjustable model. Its tuning signal \( \hat{\omega} \) is obtained through a proportional-integral (PI) controller from a scalar error signal \( e \), computed from the phase displacement between the two estimated flux vectors, \( \Psi_S^* \) and \( \Psi_R^* \). As the error signal \( e \) gets minimized by the PI controller, the tuning signal \( \hat{\omega} \) approaches the actual speed of the motor.

The stator model uses a delay element instead of an integrator to eliminate an accumulation of the drift error. A bandwidth limiter at the input of the rotor model ensures equivalence of the two models.
3.2 Feedforward control of stator voltages

Okuyama [3] derives the commanded stator voltages from a steady-state machine model as the basic reference signals. Their components in rotor field coordinates are marked by the shaded frames in Fig. 3. It is thus the machine which, through its model, lets the inverter duplicate the actual terminal voltages. This process can be characterized as self-control.

A superimposed $i_d$-controller compensates for the parameter inaccuracy of the steady-state model, while an $i_q$-controller ensures the correct alignment of the rotor field angle with the rotor flux vector. Transient deviations from the correct field alignment are eliminated by the additional correction signals A and B [1, 3].

Torque rise time of this scheme is reported around 15 ms; speed accuracy is within $\pm 1\%$ above $3\%$ rated speed and $\pm 12$ rpm at 45 rpm [3].

3.3 Rotor field orientation with improved stator model

A sensorless rotor field orientation scheme based on the stator model is described by Ohtani [4]. The upper portion of Fig. 4 shows the classical structure in which the controllers for speed and rotor flux generate the current reference vector $i_\text{s}^*$ in field coordinates. This signal is transformed to stator coordinates and processed by a set of fast current controllers CR PWM. A possible misalignment of the reference frame is detected as the difference of the measured $q$-axis current from its reference value $i_q^*$. This error signal feeds a PI controller, the output of which is the estimated mechanical speed $\hat{\omega}$. It is added to an estimated value $\hat{\theta}$. The rotor frequency component from the reference values $i_q^*$ and $\Psi_r$. The integration of $\hat{\omega}$ produces the field angle $\delta$.

The stator model is used to estimate the rotor flux vector $\Psi_r$. The band-limited integration by means of a first-order delay entails a severe loss of gain in the estimated flux component $\Psi_{r1}$ at low stator frequency, and a large field angle error. This tendency is compensated by an added flux component $\Psi_{r2}$, obtained from the reference value $\Psi_r^{\ast(S)}$ in stator coordinates. The latter component dominates the resulting flux vector $\hat{\Psi}_r$ at low frequency, replacing it by its reference value $\Psi_r^*$ in a smooth transition. This results in $\hat{\Psi}_r = \Psi_r^*$ at low frequency, which deactivates the rotor flux controller in effect. However, the field angle $\hat{\delta}$ as the argument of the rotor flux vector is still under control through the speed controller and the $i_q$-controller, although the accuracy of $\hat{\delta}$ reduces. Field orientation is finally lost at very low stator frequency. Only the frequency of the stator currents is then controlled. The currents are forced into the machine without reference to the rotor field. This provides robustness and certain stability, although not dynamic performance.

As the speed increases from a very low value, rotor flux estimation becomes more accurate and closed loop rotor flux control is resumed. The correct value of the field angle is re-adjusted as the $q$-axis current, obtained from the $\Psi_{r}\text{-estimator}$ Fig. 5, now relates to the correct rotor flux vector. The $i_q$-controller then adjusts the estimated speed, and in consequence also the field angle, for a realignment of the reference frame with the rotor field.

At 18 rpm, speed accuracy is reported to be within $\pm 3$ rpm. Torque accuracy at 18 rpm is about $\pm 0.03$ pu. At 0.1 pu. reference torque, improving significantly as the torque increases. Minimum parameter sensitivity exists at $\tau_1 = \tau_r$ [4].

3.4 Adaptive Observers

The accuracy of the open loop estimation models described
in the previous chapters reduces as the mechanical speed reduces. The limit of acceptable performance depends on how precisely the model parameters can be matched to the corresponding parameters in the actual machine. It is particularly at lower speed that parameter errors have significant influence on the steady-state and dynamic performance of the drive system.

The robustness against parameter mismatch and signal noise can be improved by employing closed-loop observers to estimate the state variables, and the system parameters.

3.4.1 Full order nonlinear observer

A full order observer can be constructed from the machine model, shown in the upper portion of Fig. 6. The model outputs the estimated values \( \hat{i}_s \) and \( \hat{\Psi}_r \) of the stator current vector and the rotor flux linkage vector, respectively.

Adding an error compensator to the model establishes the observer. The error vector computed from the model current and the measured machine current is \( \Delta i_s = \hat{i}_s - i_s \). It is used to generate correcting inputs to the electromagnetic subsystems that represent the stator and the rotor in the machine model. Kubota et al. [5] select the complex gain factors \( G_s(\hat{\omega}) \) and \( G_r(\hat{\omega}) \) such that the two complex eigenvalues of the observer \( \lambda_{1,2 \text{obs}} = k \lambda_{1,2 \text{mach}} \), where \( \lambda_{1,2 \text{mach}} \) are the machine eigenvalues, and \( k > 1 \) is a real constant. The value of \( k > 1 \) scales the observer by pole-placement to be dynamically faster than the machine. Given the nonlinearity of the system, the resulting complex gains \( G_s(\hat{\omega}) \) and \( G_r(\hat{\omega}) \) in Fig. 21 must depend on the estimated angular mechanical speed \( \hat{\omega}_m \).

The rotor field angle is derived from the components of the estimated rotor flux linkage vector \( \hat{\Psi}_r \). The speed signal \( \hat{\omega}_r \) is required to adapt the rotor structure of the observer to the mechanical speed of the machine. The signal is obtained through a PI-controller from the current error \( \Delta i_s \). In fact, the term \( \hat{\Psi}_r \times \Delta i_s \) represents the torque error \( \Delta T_r \) caused by the current error \( \Delta i_s \). If a model torque error exists, the estimated speed signal \( \hat{\omega}_r \) gets corrected by the PI controller, thus adjusting the \( \hat{\omega} \)-input to the rotor model. The phase angle of \( \hat{\Psi}_r \), which is the estimated rotor field angle, then approximates the true field angle that exists in the machine. The correct speed estimate is reached when the current error \( \Delta i_s \) and hence the torque error \( \Delta T_r \) reduce to zero.

The control scheme is reported to operate at a minimum speed of 0.034 p.u. or 50 rpm [5].

4. STATOR FIELD ORIENTATION

Control at stator field orientation is preferred in combination with the stator model. This model directly estimates the stator flux vector. Using the stator flux vector to define the coordinate system is then a straightforward approach.

A fast current control system renders the stator current vector a forcing function, making the electromagnetic subsystem of the machine behave like a complex first-order system. It is only characterized by the dynamics of the rotor winding.

Other than rotor field orientation, stator field orientation does not inherently provide decoupled control of torque and flux. In the signal flow graph Fig. 7, the torque command exerts an undesired influence on the stator flux. Xu et al. [6] propose a decoupling arrangement, shown in the left of Fig. 7, to eliminate the cross-coupling between the \( q \)-axis current and the stator flux. The internal influence of \( i_q \) is cancelled by the external decoupling signal, provided that the estimated signals and parameters match the actual machine data.

The decoupling signal depends on the rotor frequency \( \omega_r \). An estimated value \( \hat{\omega}_r \) is therefore required. It is obtained from the estimator Fig. 8, which also supplies the remaining unknown system variables. The stator flux linkage vector is estimated by the approximate stator model. The angular velocity of the revolving field is then determined from the stator flux linkage vector. Although \( \hat{\omega}_r \) is computed from the estimated value \( \hat{\Psi}_r \), its value is obtained at good accuracy. The reason is that the uncertainties in \( \hat{\Psi}_r \) are owed to minor offset and drift components in measured currents and voltage signals. These disturbances exert little influence on the angular velocity at which the space vector \( \hat{\Psi}_r \) rotates.

The stator field angle \( \hat{\delta} \) is obtained as the integral of the estimated stator frequency \( \hat{\omega}_r \). The angular mechanical velocity of the rotor is computed as \( \hat{\omega}_r = \hat{\omega}_s - \hat{\omega}_r \), where \( \hat{\omega}_s \) is derived from the stator current vector \( i_s \) in stator field coordinates and from \( \hat{\Psi}_r \) using the condition for stator field orienta-

Fig. 7 Machine control at stator flux orientation using a dynamic feedforward decoupler
5. PERFORMANCE AT VERY LOW SPEED

5.1 Stator resistance

The important information on the field angle and the mechanical speed is conveyed by the induced voltage of the stator winding, independent of the respective method that is used for sensorless control. The induced voltage \( u_s = u_s - r_s i_s \) is not directly accessible by measurement. It must be estimated, either directly from the difference of the two voltage space vector terms \( u_s \) and \( r_s i_s \), or indirectly when an observer is employed.

In the upper speed range above a few Hz stator frequency, the resistive voltage \( r_s i_s \) is small as compared with the stator voltage \( u_s \) of the machine, and the estimation of \( u_s \) is performed with good accuracy. Even the temperature-dependent variations of the stator resistance are negligible at higher speed. If operated at frequencies above the critical low speed range, variations of the stator resistance are negligible at higher speed.

As the stator frequency reduces at lower speed, the stator resistance appears in series with the machine winding; its value is therefore added to the stator resistance of the machine model. Other than this, the influence of the threshold voltage is non-linear which requires a specific inverter model.

Representing the device forward voltage, the polarity of the threshold voltage \( u_{th} \) is uniquely determined by the direction of current flow. Differences in magnitude between the forward voltage of an active device and a recovery diode are minor and can be neglected. The threshold voltage is then constant. The contributions of the three phase components permit defining a threshold voltage vector \( u_{th} \). This vector has a constant magnitude, while its phase angle changes by \( +60^\circ \) or \( -60^\circ \) whenever one of the phase currents changes its sign. The stator voltage vector \( u_s = u^* - u_{th} \) therefore follows a distorted and discontinuous trajectory, even when the reference signal \( u^* \) of the pulsewidth modulator describes a circular trajectory. Fig. 10 shows an example at motoring operation. The fundamental amplitude of \( u_s \) is then less than its reference value \( u^* \); it is larger at regeneration.

The voltage trajectory exhibits strong sixth harmonic components. Since the threshold voltage does not vary with stator frequency as the stator voltage does, the distortions are more pronounced when the stator frequency, and hence also the stator voltages, are low. The latter may even exceed the commanded voltage in magnitude, which then makes correct flux estimation and stable operation of the drive impossible.

The inverter voltage vector \( u_{th} \) is modelled by the structure in Fig. 11. Its phase angle depends on the polarities of the three phase currents, of which only six possible combinations exist. These control the sector indicator \( \text{sec}(i_s) \), which locates the current space vector in one particular \( 60^\circ \)-sector of the
complex plane. The magnitude of $u_{th}$ is $2u_{th}$ for formal reasons [1]. The resistive voltage drop of the power devices $r_d i_s$ depends on the differential resistance $r_d$. The resulting compensation signal $u_{comp}$ accounts for the linear and nonlinear voltage drops in the inverter.

### 5.3 Stator flux estimation

Using a compensated inverter makes the controlling reference voltage vector $u^*$ an undistorted and accurate estimate $\hat{u}_s$ of the fundamental stator voltage vector. This enables an accurate estimation of the stator flux linkage vector even at very low stator frequency. A signal flow graph of a stator flux estimator is shown in Fig. 12.

The upper portion of Fig. 12 shows that the stator flux vector is obtained by integrating the induced voltage $\hat{u}_i = \hat{u}_s - \hat{r}_s i_s$, [8]. Using a pure integrator avoids the estimation error and bandwidth limitation associated with a low pass filter. The method necessarily incorporates the identification of a time-varying vector $u_{off}$ that represents offset and drift voltages. These are unavoidable when analog circuits are involved, as in stator current acquisition that provides the signal $i_s$.

The offset voltage estimator is shown in the lower portion of Fig. 12. It integrates the induced voltage $\hat{u}_i$ in a second, parallel flux estimation channel. The minimum and the maximum values per fundamental period $T$ of the resulting signal $\psi$, occurring at the respective time instants $t_{\min}$ and $t_{\max}$, are subsequently added, separately for their respective $d$-axis and the $q$-axis components, to form the components of the dc offset vector $\hat{u}_{off}$. It is apparent that the offset of a sinusoidal signal is zero if its minimum and maximum values sum up to zero. The offset vector is lowpass-filtered and then fed back to the input of the lower flux integrator in Fig. 12. Its input is forced to zero in a steady state, which is proof that $\hat{u}_{off}$ represents the dc component of $\hat{u}_i$.

The induced voltage is further contaminated by harmonic components. These are introduced by errors in the acquisition of the fundamental current component $i_s$, [1], by incomplete compensation of the dead time effect [9], and by parameter errors of the nonlinear inverter model. The harmonics of $\hat{u}_i$ are detected by the ac disturbance estimator shown in the center of Fig. 12. The estimator simply computes the instantaneous difference in amplitude between the estimated stator flux vector $\hat{\psi}_s$ and its reference signal $\hat{\psi}_s^*$. Proportional control by a gain constant $k_p$ is used for fast cancellation of the harmonic components.

The estimated field angle $\hat{\delta}$ is obtained as the argument of the stator flux vector $\hat{\psi}_s$ and the estimated speed as the derivative of the field angle $\hat{\delta}$.

### 5.4 Stator resistance adaptation

An important measure to improve the low-speed performance is the accurate on-line adaptation of the stator resistance, which is the most relevant parameter in sensorless control. Kubota et al. [10] use the observer structure Fig. 6 to determine the component $\Delta i_s$, $i_s / i_s$ of the error vector $\Delta i_s$ in the direction of the stator current vector $i_s$, which is proportional to the deviation of the model parameter $\hat{r}_s$ from the actual stator resistance. The identification delay of this method is reported as 1.4 s.

A faster algorithm relies on the orthogonal relationship in steady-state between the stator flux vector and the induced voltage [8]. The vector diagram Fig. 13 illustrates the principle. The space vector signals are processed in a synchronous reference $xy$-frame C, aligned with the stator current vector $i_s$. The relation $r_s = (u_{sx} - u_{ix})/u_x$ is easily verified, where $u_{ix} = u_i \sin (\gamma - \delta)$.

To facilitate the correct estimation of $r_s$, the vector $u_i$ of the induced voltage, obtained as the derivative of the stator flux vector $\psi_s$, must not depend on the stator resistance $r_s$. The reason is that $r_s$ is the variable to estimate. The estimated stator flux is therefore derived from the instantaneous reactive power, defined as $q = u_i \times \Delta i_s$|z|, which notation describes the $z$-component of the vector product of the stator voltage and stator current vector.

The estimated stator resistance value is then used as an input signal $\hat{r}_s$ to the stator flux estimator Fig. 12. It adjusts its parameter through a lowpass filter. The filter time constant $T_f = \alpha \times T_f$ is about 100 ms. The identification delay of this method is about 350 ms [8].
5.5 Low speed performance achieved by improved models
The oscillogram Fig. 14 shows the response to load step changes of rated magnitude while the speed is maintained constant at 5 rpm. This corresponds to operating at a stator frequency of 0.16 Hz ($\omega_s = 0.003$) during the no-load intervals. Fig. 15 demonstrates persistent operation at zero stator frequency. A load step of 120% rated magnitude demonstrates that correct field orientation was maintained.

6. SENSORLESS CONTROL THROUGH SIGNAL INJECTION

Signal injection methods exploit machine properties that are not reproduced by the fundamental machine model described in Sections 3 and 4. The injected signal excites the machine at a much higher frequency than that of the fundamental field. The resulting high-frequency currents generate flux linkages that close through the leakage paths in the stator and the rotor, leaving the mutual flux linkage with the fundamental wave almost unaffected. The high-frequency effects can be therefore considered superimposed to, and independent of, the fundamental behavior of the machine. High-frequency signal injection is used to detect anisotropic properties of the machine.

6.1 Anisotropies of an induction machine
A magnetic anisotropy can be caused by saturation of the leakage paths through the fundamental field. The spatial orientation of the anisotropy is then correlated with the field angle $\delta$, which quantity can be identified by processing the response of the machine to the injected signal. Other anisotropic structures are the discrete rotor bars in a cage rotor. A rotor may be also custom designed so as to exhibit periodic variations within a fundamental pole pitch of local magnetic or electrical characteristics, for example the dimensions of the rotor slots. Detecting such anisotropy serves to identify the rotor position angle, the changes of which define the shaft speed.

The case of a saturation-induced anisotropy is considered first. The fundamental field saturates the stator and rotor iron in the region of higher flux density, there producing higher magnetic resistivity of the local leakage paths. A transient excitation by an injected voltage $u_{tr}$ changes the leakage fluxes in the direction of the forcing voltage vector, $u_{tr} = \delta \varphi_{\sigma \Omega} / \delta t$. The transient currents change accordingly. Given the higher magnetic resistance in the saturated region, the local leakage paths require more magnetizing current than the neighboring zones of less fundamental flux density. The space vector $i_{tr}$ of the transient current therefore inclines in space towards the saturated region as seen from the leakage flux vector $\varphi_{\sigma \Omega}$.

The leakage inductance is therefore not constant in a saturated machine. It varies with the angular displacement between the leakage flux vector $\varphi_{\sigma \Omega}$ and the fundamental flux vector. The scalar leakage $l_{\sigma}$ inductance in the stator equation $u_{tr} = l_{\sigma} \delta i_{tr} / \delta t$ of an isotropic machine converts into a transient inductance tensor $l_{tr}$ when an anisotropy exists. The stator equation becomes $u_{tr} = l_{tr} \delta i_{tr} / \delta t$. The tensor $l_{tr}$ reflects only the fundamental component of the spatial saturation function. It describes the physical phenomenon that the derivative $\delta i_{tr} / \delta t$ of the transient stator current has a different direction in space than the voltage vector $u_{tr}$ that originates the excitation.

There is generally more than one anisotropy present in an induction motor. The existing anisotropies have different spatial orientations, such as the actual angular position of the fundamental field, the position of the rotor bars within a rotor bar pitch, and, if applicable, the angular position within a fundamental pole pair of a custom designed rotor.

Multiple anisotropies make the transient inductance tensor more involved. Fig. 16 is a visualization of the inverse inductance tensor $l_{tr}^{-1}$ that describes the direction of a transient current change as a function of the exciting voltage: $\delta i_{tr} / \delta t = l_{tr}^{-1} u_{tr}$. It displayed over one electrical revolution of an
unloaded induction machine. The quasi-circles are caused by the rotor bar anisotropy; their number corresponds to the number of rotor bars per pole pitch. They displace along another, low-frequency trajectory under the influence of the saturation anisotropy. Their shapes vary, changing from a vertical ellipse through almost a circle to a horizontal ellipse. This is owed to the changing magnetic saturation of the leakage paths as the fundamental field rotates.

The response to an injected high-frequency signal necessarily reflects all anisotropies, field dependent and rotor position dependent. While intending to extract information on one particular anisotropy, the other anisotropies act as disturbances.

6.2 Signal injection

The injected signals may be periodic, creating either a high-frequency revolving field, or an alternating field in a specific, predetermined spatial direction. Such signals can be referred to as carriers, being periodic at the carrier frequency with respect to space, or time. The carrier signals, mostly created by additional components of the stator voltages, get modulated by the actual orientations in space of the machine anisotropies. The carrier frequency components are subsequently extracted from the machine currents. They are demodulated and processed to retrieve the desired information.

Instead of injecting a periodic carrier, the high-frequency content of the switched waveforms in a PWM controlled drive system can be exploited for the same purpose. The switching of the inverter produces a perpetual excitation of the transient leakage fields [11]. Their distribution in space is governed by the anisotropies of the machine. Measuring and processing of adequate voltage or current signals permits identifying their spatial orientations.

6.3 Injection of a revolving carrier

A polyphase carrier rotating at frequency \( \omega \) is generated by the voltage space vector \( u_c = u_c \exp(j \omega t) \). This signal is added to the controlling voltage of the pulsewidth modulator, thus creating the transient excitation of the machine. The modulation by the machine anisotropies reflects in a space vector \( i_c \) of carrier frequency \( \omega \), forming part of the measured stator current vector \( i_s \). It is separated by a filter from the fundamental current \( i_s \) of lower frequency, and from the switching harmonics of higher frequencies.

The separated space vector \( i_c \) consists of a positive sequence component \( i_p \), and a number of components \( i_{qj} \) that rotate at the angular velocity \( -\omega + 2\omega_p \), i.e. in negative directions. Of the latter components, one particular must be separated to extract the angular orientation \( \omega \) of the selected anisotropy. The others act as disturbances.

Rotating at the frequency of the carrier signal, the current vector \( i_c \) follows in fact elliptic trajectories. This is owed to the existence of its positive sequence and negative sequence components. There are as many ellipses superimposed as anisotropies exist.

The axis ratio of a saturation ellipse is \( l_{ad}/l_{dd} \), close to unity value that ranges between 0.9 and 0.96 [12, 13]. It is therefore difficult to identify the angular inclination of the ellipse and thus determine the angular orientation of the anisotropy. The characterizing component \( i_{ad} \) is very small, being superimposed by the larger positive sequence current vector \( i_p \) and contaminated by the effect of other anisotropies and disturbances. Finally, all these signals are buried under the much larger fundamental current \( i_s \), and under the switching harmonics.

To give an example, the current amplitudes \( i_p/i_{1R} \) and \( i_{ad}/i_{sR} \) from [12], referred to the rated fundamental current \( i_{sR} \) are shown in Fig. 17. The values are measured from an induction machine at zero fundamental excitation, \( i_1 = 0 \), such as to avoid saturation generating an additional anisotropy. However, the rotor has an engineered anisotropy of \( l_{ad}/l_{dd} = 0.91 \). There are three categories of negative sequence currents:

- The current \( i_2 \) at frequency \( -\omega + 2\omega \) is caused by the engineered rotor anisotropy. Its harmonic spectrum spreads between \( -\omega \) and \( -\omega + 2\omega_{max} \) when the machine speed \( \omega \) varies between 0 and \( \omega_{max} \), where \( \omega_{max} = 2\pi \cdot 10 \text{ Hz} \) is an assumed maximum value in Fig. 17. This frequency component carries the speed information; its magnitude \( i_2 = 0.022 i_{max} \) is extremely low.
- The current \( i_{slot} \) at frequency \( -\omega_c + N/p \omega \) is caused by the discrete rotor slots; it extends over the frequency range \( -\omega_c \) to \( -\omega_c + N/p \omega_{max} \), where \( N \) is the number of rotor slots and \( p \) is the number of pole pairs.
- The current \( i_q \) at frequency \( -\omega_c \) originates from winding asymmetries, and from gain unbalances in the stator cur-
rent acquisition circuits. This disturbance is in very close spectral proximity to the speed-related component \(i_2\); both converge to the same frequency at \(\omega = 0\). Also, \(i_u > i_2\) in this example.

If this machine was fully fluxed and loaded, another negative sequence current \(i_{\text{sat}}\) would appear at frequency \(-\omega_s + 2\omega_c\). Also this component has an extremely low spectral distance \(2(\omega_s - \omega)\) from the component \(i_2\), where \(\omega_s - \omega\) is the slip frequency.

The distribution of the significant negative sequence spectra in Fig. 17 indicates that it is extremely difficult to separate these signals by filtering [14].

### 6.4 Speed and position estimation based on anisotropies

Degner and Lorenz [12] use a dynamic model of the mechanical subsystem of the drive motor as part of a closed phase-locked loop (PLL) for spectral separation. A signal flow graph of the speed and rotor position estimator is shown in Fig. 18. The carrier generated space vector \(i_c\) is transformed to a \(+\omega_c\)-reference frame in which \(i_c\) shows as a complex constant. Its contribution is nullified through the feedback action of an integrator. The remaining signal \(i_c\) contains all negative sequence components. It is transformed to the \(-\omega_c\)-reference frame. This transformation shifts the frequency origin in Fig. 17 to \(-\omega_s\); the negative sequence components then appear as low-valued positive sequence signals.

The unbalance disturbance at frequency zero is compensated by an estimated vector \(\hat{i}_u = i_u \exp(j\hat{\phi}_u)\), and the disturbance generated by rotor slotting by an estimated vector \(\hat{i}_{\text{slot}}\). What remains is the current vector \(\hat{i}_2 = i_2 \exp(j(2\omega t + \phi_2))\), representing the rotor anisotropy as a second harmonic. This signal carries the important information, since \(2\omega t + \phi_2\) is twice the rotor position angle; \(\phi_2\) is a phase displacement introduced by signal filtering.

The mechanical system model in the upper right of Fig. 18 receives an acceleration torque signal, formed as the difference between the electromagnetic torque \(T_{\text{el}}\) and the load torque \(T_L\), both being represented by their estimated values. The feedforward signal \(T_L\) serves to improve the estimation dynamics. It is obtained by a separate load model. The estimated angular velocity \(\hat{\omega}\) of the rotor is the integral of the acceleration torque, where \(\tau_m\) is the normalized mechanical time constant. Integrating \(\hat{\phi}\) yields the estimated rotor position angle \(\hat{\theta}\).

The estimated angle \(\hat{\theta}\) controls two anisotropy models. The upper model in Fig. 18 forms part of the PLL. It computes the phase angle component \(2\omega t + \phi\) of the negative sequence current vector \(i_2\), while its magnitude \(i_2\) and phase displacement \(\phi_2\) are introduced as predetermined constant parameters. As commanded by the computed phase angle error \(e = \hat{i}_2 \times i_2\), the PID controller forces the resulting space vector \(\hat{i}_2\) to align with its reference vector \(i_2\), thus establishing \(\hat{\theta} = \theta\) as desired. The anisotropy model thus serves to impress on the estimated current vector \(\hat{i}_2\) the same rotor position dependent variations that the real machine, through its inherent anisotropy, forces on the negative sequence current component \(i_2\).

The rotor slot related current vector \(\hat{i}_{\text{slot}}\) is estimated by the anisotropy model in the lower portion of Fig. 13 in a similar fashion. The vector \(\hat{i}_{\text{slot}}\) is used to compensate the undesired disturbance \(i_{\text{slot}}\) that forms part of \(i_u\).

The saturation-induced anisotropy is not modelled in this approach, which limits its application to unsaturated machines. Another problem is the nonlinearity of the PWM inverter which causes distortions of the machine currents. These generate additional negative sequence current components that tend to fail the operation of the position estimator [15]. A general difficulty of all revolving carrier injection methods is the extreme low signal-to-noise ratio which is less than \(10^{-3}\) in the example of Fig. 17. This calls for special efforts to ensure that the low-level signals are sufficiently reproduced when doing the analog-to-digital conversion of the measured currents [16].

The nonlinearity of the PWM inverter, commonly known as dead-time effect, produces distortions of the pulselwidth modulation whenever one of the phase currents changes its sign. With the high-frequency carrier signal superimposed to the modulator input, the stator currents are forced to multiple zero crossings when the fundamental phase currents are close to zero. The effect causes severe current distortions that the established methods for dead-time compensation cannot handle.

Being time-discrete events, the current distortions are difficult to compensate in a frequency domain method. An offline identification method was proposed by Teske [15]. The compensation is done here using sets of time-variable profiles over one electrical revolution, one profile for every operating point in terms of load and excitation level, and separately for the respective \(d\)- and the \(q\)-component.

Current publications on revolving carrier methods show that numerous side effects require the signal processing structures to get more and more involved, while the dependence on pa-

![Fig. 18 Speed and rotor position estimator using a PLL to identify the response to an injected revolving carrier](image)

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rameters or on specific off-line commissioning procedures persists.

### 6.5 Injection of an alternating carrier

Revolving carriers scan the whole circumferential profile of all anisotropies that exist in a machine. This serves to determine the characteristics of only one particular anisotropy. The subsequent processing steps fully identify its spatial orientation without using a-priori knowledge. An alternative class of methods relies on injecting not a rotating, but alternating carrier in a specific, though time-variable spatial direction. The direction is selected in an educated guess to achieve maximum sensitivity in locating the targeted anisotropy. It is an advantage of these methods that already existing knowledge is just updated by acquiring an incremental error per sampling period.

### 6.6 Balance of quadrature impedances

The approach of Ha and Sul [17] aims at identifying a field angle while the machine operates at low or zero speed. According to the vector diagram Fig. 19, an alternating high-frequency ac carrier signal of amplitude $u_c$ is added to the $d$-axis component of the control input of the pulsewidth modulator. The voltage signal excites the machine in the direction of the estimated $d$-axis. This direction may have an angular displacement $\gamma = \delta - \delta$ from the true $d$-axis, the location of which is approximately known from the identification in a previous cycle.

The injected voltage adds an ac component $i_c$ to the fundamental stator current components of the machine, represented in Fig. 19 by the space vector $i_{c1}$. Owing to the anisotropic machine impedance, the high-frequency ac current $i_c$ develops at a spatial displacement $\gamma$ with respect to the true field axis of the machine.

When the machine is operated in saturated conditions, its impedance $Z_c$ at carrier frequency $\omega_c$ is a function of the circumferential angle $\alpha$ in field coordinates. The impedance has a maximum value $Z_q$ in the $d$-axis, and a minimum value $Z_d$ in the $q$-axis. The identification of the $d$-axis is based on the assumption of a symmetric characteristic $Z_c(\alpha) = Z_c(-\alpha)$. An orthogonal $xy$-coordinate system is introduced in Fig. 18, having its real axis displaced by $-\pi/4$ with respect to the estimated $d$-axis. Its displacement with the true $d$-axis is then $-\pi/4 - \gamma$.

The identification procedure is illustrated in the signal flow graph Fig. 20, showing the current control system and the generation of the ac carrier in its upper portion. The shaded frame in the lower portion highlights the field angle estimator. Here, the measured stator current $i_s$ is bandpass-filtered to separate the ac carrier current $i_c$. The current $i_c$ and also the excitation signal $u_c \cos \alpha$, are transformed to $xy$-coordinates, and then converted to complex vectors that have the respective rms amplitudes and conserve the phase angles. The complex high-frequency impedance $Z_c$ is formed which is a function of the transformation angle $\delta - \pi/4$; seen from the field oriented coordinate system in Fig. 19, the transformation angle is $-\pi/4 - \gamma$. For reasons of symmetry, the real and imaginary components $Z_q$ and $Z_d$ would be equal if accurate field alignment, $\gamma = 0$, existed. A nonzero error angle $\gamma$ makes $Z_q$ increase, and $Z_d$ decrease. Hence an error signal $\varepsilon = Z_q(\gamma) - Z_d(\gamma)$ can be constructed which adjusts the estimated field angle $\delta$ to an improved value by means of a PI controller. Fig. 20 shows that this angle is used for coordinate transformation. In a condition of accurate field alignment, $\delta \rightarrow \delta$, from which $\gamma \rightarrow 0$ follows.

It is documented in [17] that the difference between the impedance values $Z_d$ and $Z_q$ is small when the machine is fully saturated. The reduced error sensitivity then requires a high amplitude of the injected signal.

### 6.7 Evaluation of elliptic current trajectories

The carrier injection methods described so far suffer from certain drawbacks. We have the poor signal-to-noise ratio and the parameter dependence of the revolving carrier methods, and the low sensitivity of the quadrature impedance method.
Linke [18] uses an alternating ac carrier injected in the estimated direction of the $d$-axis to extract a saturation anisotropy signal at high signal-to-noise ratio. An alternating carrier voltage can be decomposed into two identical components that rotate in opposite spatial directions. The associated current trajectories assume elliptic shapes under the influence of existing anisotropies [1]. The spatial orientations of these ellipses deviate from the true field axis as a function of the error angle $\gamma = \Delta - \delta$ between the true and the estimated field axis.

The signal flow graph Fig. 21 illustrates the field angle estimation scheme. The bandpass filtered carrier-frequency current $i_c$ is rotated by $\omega_c t + \delta$ which makes its positive sequence component $i_p$ a complex dc value. A lowpass filter suppresses all accompanying ac components like negative sequence currents and switching harmonics. The imaginary part of $i_p$ is $-\sin 2\gamma$, which is proportional to the error angle $\gamma = \Delta - \delta$ for small error values. This signal is sampled at about 1 kHz. It feeds an I-controller to create the estimated field angle $\hat{\delta}$ in a closed loop. In doing so, reference is made to the injected carrier signal to build the transformation term $\omega_c t + \delta$.

As the acquired signal is a dc value in principle, the sampling frequency can be chosen independently from the carrier frequency. This ensures good and dynamically fast alignment with the field axis without the need of choosing a high carrier frequency. Also the dynamics of the speed and torque control system is not impaired as the carrier signal does not appear in the torque building $q$-current component. Therefore, the measured $q$-current need not be lowpass filtered, as is required when a rotating carrier is used. According to Fig. 21, such filter is only provided for the component $i_d$ in the excitation axis.

The method yields a signal of high signal-to-noise ratio, permitting operation at low carrier level. A 100-mA carrier current was found sufficient for field angle estimation in a 1-kW drive system.

### 6.8 High-frequency excitation by PWM switching

The switching of a PWM inverter subjects the machine to repetitive transient excitation. The resulting changes $di/d\tau$ of the transient machine currents are influenced by the anisotropies of the machine, characterized by the transient inductance tensor $I_T$. The transient current derivative is separated from the fundamental component by taking the associated zero sequence voltage $\Delta h_d/d\tau$ from a wye-connected stator winding [19]. This voltage is proportional to the characteristic component of the anisotropy signal in that particular phase axis in which a switching has occurred. Reconstructing the complete spatial orientation of an anisotropy requires therefore the evaluation of a minimum of two switching events in different phase axes. The switching instants must be executed within a very short time interval, such that the angular orientation of the anisotropy remains almost unchanged.

The acquired axis components of the anisotropy signal compose a rotor position vector $p(\theta_N)$. The angle $\theta_N$ indicates the angular position of the rotor within one rotor bar pitch. The oscillogram Fig. 22 shows a full revolution of $p(\theta_N)$. This emphasizes the high spatial resolution that this method provides. Also noteworthy is the high level of the acquired signals, which is around 35 V. A full mechanical revolution is performed when $\theta_N/N$ increments by $2\pi$, where $N$ is the number of rotor bars.

The anisotropy signals are obtained by instantaneous sampling of speed-independent values. The influence of the saturation-induced anisotropy can be suppressed by spatial filters [20]. Rotor position acquisition is possible at sampling rates of several kHz [13]. The spatial resolution and the signal-to-noise ratio are very high. This permits implementing precise incremental positioning systems for high dynamic performance. However, the incremental position is lost at higher speed when the frequency of the position signal becomes higher than twice the sampling frequency.

Other than continuous carrier injection methods, which are frequency domain methods, PWM excitation constitutes a sequence of non-periodic time-discrete events, and hence requires time-domain methods for signal processing. The absence of spectral filters enables a faster dynamic response. The oscillogram Fig. 23 shows a positioning cycle that requires maximum dynamics at 120% rated torque. The high magnetic saturation during the acceleration intervals temporarily reduces the amplitude of the position signals; the posi-

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**Fig. 21** Signal flow graph of a field angle estimation scheme based on injecting an alternating carrier

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**Fig. 22** Trajectory of the rotor position vector $p(\theta_N)$ within one rotor bar pitch

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**Fig. 23** Oscillogram of an anisotropy signal during a position cycle.
tion accuracy remains unaffected, as the relevant information is contained in the phase angles.

7. SUMMARY AND PERFORMANCE COMPARISON

A large variety of sensorless controlled ac drive schemes are used in industrial applications. Open loop control systems maintain the stator voltage-to-frequency ratio at a predetermined level to establish the desired machine flux. They are particularly robust at very low and very high speed, but satisfy only low or moderate dynamic requirements. Small load dependent speed deviations can be compensated incorporating a speed or rotor frequency estimator.

High-performance vector control schemes require a flux vector estimator to identify the spatial location of the magnetic field. Field oriented control stabilizes the tendency of induction motors to oscillate at transients, which enables fast control of torque and speed. The robustness of a sensorless ac drive can be improved by adequate control structures and by parameter identification techniques. Depending on the respective method, sensorless control can be achieved over a base speed range of 1:100 to 1:150 at very good dynamic performance. It is a particular attraction of the fundamental model that runs on a simple hardware platform. Also stable and persistent operation at zero stator frequency can be achieved with these models, provided that all drive system components are precisely modelled and their parameters correctly adapted. Accurate speed estimation in the very low speed region, however, is difficult since the fundamental model becomes unobservable.

Improved low speed performance can be achieved by exploiting the anisotropic properties of induction motors. The spatial orientations of the anisotropies are related to the field angle, and to the mechanical rotor position. These can be identified either by injecting high-frequency carrier signals into the stator windings and process the response of the machine, or by making use of the transients that a PWM inverter generates. These methods have recently emerged. They bear great promise for the development of universally applicable sensorless ac motor drives.

8. REFERENCES