Zero-Speed Tacholess IM Torque Control: Simply a Matter of Stator Voltage Integration

Kevin D. Hurst, Member, IEEE, Thomas G. Habetler, Senior Member, IEEE, Giovanni Griva, and Francesco Profumo, Senior Member, IEEE

Abstract—Most industry experts agree that the next generation of commercial drives will include some sort of sensorless torque control. Achieving a modest level of control in the very-low-speed range greatly increases the competitive value of the drive and expands its range of applications. After reviewing many of the previously presented sensorless control methods, this paper shows that the simple method of stator-flux orientation can provide zero-speed torque control equally as well as more complex approaches, which rely on elaborate mathematical models to improve the operating range. This paper experimentally demonstrates that, by using only a slight amount of low-pass filtering in the integration of the stator voltage, a reasonable flux estimate can be obtained for stator-flux-field-oriented torque control. With this simple method, adequate torque control has been demonstrated over a wide range of speed and load conditions, and even at zero rotor speed with moderate or heavy load torque.

Index Terms—Induction machine drives, sensorless control.

I. INTRODUCTION

The need for tacholess speed and torque control of induction machines (IM’s) has become widely recognized because of the cost and fragility of a mechanical speed sensor, and because of the difficulty of installing the sensor in many applications. For these reasons, most industry experts agree that the next generation of commercial drives will include some sort of sensorless torque control. Typically, the dynamic control performance of a sensorless drive does not match that of a standard vector drive with a speed sensor. In many applications, however, high-bandwidth torque control is not as critical as maintaining at least low-bandwidth control over the widest possible speed range. In applications such as cranes, hoists, presses, and traction drives, for instance, maintaining the desired torque down to zero speed (locked rotor) is highly desirable. Achieving a modest level of control in the very-low-speed range, in fact, greatly increases the competitive value of the drive and expands its range of applications.

These advantages can be realized at very little cost only if a mechanical sensor is not required. After reviewing many of the previously presented sensorless control methods, this paper will show that the simple method of stator-flux orientation can provide zero-speed torque control equally as well as more complex approaches, which rely on elaborate mathematical models to improve the operating range.

II. MOTOR FLUX ESTIMATION

Numerous methods have been proposed for providing torque control in a sensorless IM drive [3], [5], [7], [9]–[16]. In one approach, the line current or voltage is measured, in order to detect rotor-speed-related harmonics caused by machine saliencies [1]–[4]. After using these harmonics to find the rotor speed, the controller can calculate the rotor-flux position and, thus, provide field-oriented torque control. Since these harmonics depend on parasitic characteristics of the machine, however, they can be difficult to detect in some cases without design modification or signal injection. For instance, several techniques have been proposed for the analysis of speed-related harmonics caused by rotor slotting [1]–[3]. While this technique has been successfully employed on some IM’s, other machines exhibit such small rotor slot harmonics that the desired harmonics can be obscured by noise in the low-speed range. Another method of this type finds the flux position by detecting magnetic saliencies caused by main flux saturation [4], but this technique operates poorly in most standard IM’s, in which these salient effects are minimized.

A more common approach used to provide sensorless control involves estimating the flux (and/or speed) from the terminal voltage and current measurements. By obtaining the position of any flux in the machine, field-oriented torque control can be achieved. The most basic method for estimating the flux is by using the stator voltage model, which integrates the stator voltage, in order to estimate the stator flux, shown by [5]

$$\lambda_{qds} = \int (V_{qds} - \hat{R}_s I_{qds}) \, dt$$

where $\lambda_{qds}$ is the estimated stator flux, $R_s$ is the stator resistance, and $V_{qds}$ and $I_{qds}$ are the measured terminal voltages and currents, respectively. This method has been demonstrated by many researchers to work well over most of the speed range. Fig. 1 shows a block diagram of a sensorless IM drive based only on stator voltage integration, as described by (1). At low speeds, however, voltage measurement noise
and integration drift can pose a significant problem, although these limitations can be largely resolved with some simple modifications discussed below. In order to assess the performance of a controller based on (1), one must first address the host of schemes which attempt to improve the low-speed performance of this simple flux estimator.

Any flux or speed estimation algorithm which relies on a mathematical model of the machine is known as a flux or speed observer. Equation (1) represents the simplest possible flux observer for an IM. Many other, more complex schemes have been devised which add mathematical machinery to this stator voltage model in an attempt to improve the accuracy of the flux estimation and/or to provide speed estimation. It is important to note that all of these observers depend first and foremost on some variation of (1), i.e., integration of the stator voltage.

The most successful contribution of these observers is the derivation of a reasonably accurate speed estimate. Originally presented in 1986 [6], this method finds the speed by comparing the flux estimate of (1) with a separate flux estimate derived from the rotor current model, given by

\[ \dot{\lambda}_{qdr} = \frac{L_m}{L_m} \frac{1}{\tau_r} \dot{\lambda}_{dq} - j\omega_r \lambda_{qdr} \]  

(2)

where \( p \) is the time derivative operator, \( \tau_r = L_r/R_r \) (the rotor time constant) and is the rotor speed in electrical radians. Many papers [7], [8] have presented speed observers which depend on other versions of (2), which itself is a form of the slip relation used in standard vector drives. All of these schemes boil down to one central concept: the error between the flux estimates of (1) and (2) forces the speed estimate \( \omega_r \) to converge to the true speed. Although this speed estimate typically has some steady-state error because of parameter detuning, it can provide sufficiently accurate speed control for many applications.

Many researchers have gone a step further and have developed schemes which, along with the speed observer, attempt to improve the flux estimation of (1) by incorporating the rotor current model of (2). The primary motivation for improving the model of (1) is that any errors in the voltage integration are magnified at low frequencies and, thus, the flux estimate can become unstable. Therefore, these schemes attempt to provide some error feedback which can stabilize the integral of (1). In the simplest case, by using the speed estimate of the previous time step, a flux estimate can be obtained from (2) which could be used to correct the estimate in (1) [9]. Starting with this idea, numerous observers have been proposed which attempt to estimate the speed and the flux simultaneously. As shown in [6], (2) is quite suitable for estimating speed only, but there is a fundamental shortcoming in these schemes in which one equation is used to estimate two unknowns, speed and flux. This process is obscured by some techniques; for instance, in [10], (2) is used to estimate the torque, and then this torque is used to find the speed estimate required by (2). In another case, (2) is rewritten in the rotor reference frame, but is still dependent on the speed estimate [11]. Other researchers have used a variety of theoretical concepts to combine (1) and (2), including full-order nonlinear observers [12], sliding-mode observers [13], extended Kalman filters [14], and reduced-order nonlinear observers [15]. All of these schemes, however, suffer from the same limitation—an attempt to obtain two separate flux estimates (for error feedback) and the speed estimate required by (2), i.e., using two equations to find three unknowns. The calculation of the speed from (2) is a necessary intermediate step, in order to estimate the flux from (2). Without heavy filtering, the positive feedback inherent in this type of algorithm can lead to control instability.

With sufficient filtering, however, this (underdetermined) error feedback scheme does provide an important benefit. The feedback which is generated by any of these methods does tend to stabilize the pure integral of (1). The instability of the integral calculation, which has also been termed “integral windup,” is different from control instability. The problem is that, without any feedback, the flux estimate in (1) can be destabilized by a variety of errors that become particularly detrimental at low stator frequencies, including measurement noise, digital approximation errors, parameter detuning, and dc offset in the measurements. Thus, any error feedback to the integral is preferable to pure integration.

One other method that should be mentioned uses the flux reference itself to provide error feedback [16]. Although it could be subject to stability problems, this method clearly demonstrates the value of providing some feedback to the integral, even if that feedback has no necessary dependence on the measured state of the machine. The low-speed performance improvement which has been shown by this approach can only
be a result of having some feedback, even that which does not
depend at all on any control theory or mathematical models.

It will now be demonstrated that there is a much simpler, and
generally more robust, means of achieving the same objective of
feedback to the voltage integral than by any of the above
methods. In the implementation of (1), a low-pass filter can
be substituted for the pure integral. This approach is suggested
by a couple of mathematical steps. First, consider a frequency-
domain version of (1), with the variables \( y \) and \( x \) substituted
for the flux estimate and stator voltage, respectively

\[
y = \frac{1}{s} x.
\]

Equation (3) can be expanded and rewritten as

\[
y = \frac{1}{s + \omega_0} x + \frac{\omega_0}{s + \omega_0} y.
\]

By adding to the right-hand side a negative feedback equal in
magnitude to the second term, the pure integral collapses to
a simple low-pass filter

\[
y = \frac{1}{s + \omega_0} y
\]

where \( \omega_0 \) is the filter cutoff frequency. By choosing \( \omega_0 \) to be
very small, (5) well approximates the pure integral of (3). It
can likewise be seen that the second term of (4), which is the
negative feedback used to create (5), also is a small quantity
when \( \omega_0 \) is small. This term precisely represents the effect of
the large feedback resistor typically used in op-amp integrators
in analog circuits. In both hardware and software integration,
therefore, a small amount of feedback is required to maintain
the stability of the integral in the presence of noise or offset
errors.

In the digital domain, a similar procedure yields an algorithmic
form of (5). Adding a small amount of negative feedback
to a pure, digital integral yields

\[
y[n] = \sum_{n=0}^{m} \alpha^n x[m-n]
\]

where \( x[m-n] \) represents the operand of (1), \( y[n] \) represents
the flux estimate, and \( \alpha \) is a gain slightly less than unity (about
0.999 for 1-Hz stator frequency). Therefore, (6) is a practical,
digital implementation of (5), which closely approximates the
ideal stator voltage model of (1). Essentially, (5) and (6) are
low-pass filters in which the pole is selected to be very close
to zero. At higher frequencies, \( \alpha \) should be reduced to account
for the increased dc gain of the filter.

Unlike in an analog integrator, the observer algorithm can
easily adjust \( \alpha \) according to the operating frequency, in order
to achieve optimum control performance. Also, the command
voltage is used in place of the measured voltage in the
integration, which is degraded by noise at low frequencies.
Furthermore, at high pulsewidth modulation (PWM) frequen-
cies (as in this case, about 10 kHz), the voltage command
waveform is a reasonable approximation of the actual voltage.
The value of the stator resistance and other parameters required
for field orientation using the stator voltage model are assumed
to be reasonably well known if the manufacturer provides the
motor and the drive as a unit.

The experimental results shown below demonstrate the
operation of this simple, modified integration algorithm. The
additional phase lag introduced by a small degree of low-pass
filtering does not affect the stability of a typical induction
motor drive, because of the huge inertia characteristic of the
machine. Furthermore, because of the negative feedback of the
modified algorithm, any noise, dc offset, or errors in the initial
conditions of the controller decay faster than the mechanical
time constant of the machine. Therefore, the system is more
stable with the modified algorithm of (6) than with the pure
integration of (1), with which such errors can be magnified.

Based on extensive experimentation and simulation, the
authors have come to the conclusion that the improvements in
flux and speed estimation described in [6]–[16] are mostly
a result of providing some feedback, rather than none, to the
voltage integral. Often, these schemes claim that some advantage
is gained by considering the machine model in generating
the error feedback, but, at very low frequencies, these models
decay down and the characteristics of the noise are difficult
to predict. In general, the advantage of using model-based
feedback instead of direct integral feedback [as in (6)] is highly
questionable because of the large increase in the complexity
of the model and the observer gains which must be tuned.
Furthermore, most proposed observers involve an increased
computational burden and sensitivity to parameter detuning.
Additionally, without sufficient filtering, the positive-feedback
nature of some of these algorithms can produce control insta-
Bility. For these reasons, it is the opinion of the authors that
the use of complex observers such as those described above is
difficult to justify when the same performance can be achieved
using the simple, robust algorithm of (6).

III. EXPERIMENTAL RESULTS

For this paper, the value of sensorless torque control by
using only (6) has been demonstrated experimentally in two
separate laboratories using completely different equipment. In
both cases, the observer algorithm has been implemented as
part of a field-oriented controller, in order to demonstrate
closed-loop torque control in the low-speed range. At Georgia
Institute of Technology, Atlanta, the inverter space vector
PWM control, the current regulator, and the observer algorithm
operate with a 128-ms sampling/integration time step on a
Motorola DSP56001. The test motor is a four-pole 230-V
2-hp machine which is mechanically coupled to a magnetic
clutch, which provides rated torque, even at very low speeds.
A block diagram of the complete drive system and control
is shown in Figs. 1 and 2. In Fig. 1, the flux angle for field
orientation was calculated in this case using (6) in place of (1).
Voltage sensors were not used for this setup, because it was
found that the command voltages (calculated only in software)
corresponded closely to the actual voltages at 8-kHz switching
frequency. Furthermore, the use of the voltage reference avoids
the difficulty of voltage measurement at low speeds, when
the fundamental stator voltage is small. An exceptionally
accurate voltage measurement technique that also compensates
for conduction losses could provide further improvement in the
Torque control at zero speed is clearly demonstrated in Figs. 3 and 4. Fig. 3 shows the flux and torque estimation with locked rotor and 1.0-per-unit torque command, resulting in a slip frequency of 2.8 Hz. The nonsinusoidal quality of the flux estimate produces a torque ripple at the fundamental frequency, but the flux estimation and average torque output are still good and quite acceptable for many industrial applications. Fig. 4 shows the flux estimate with a 0.2-per-unit torque command, resulting in a slip frequency of 1.1 Hz. In this condition, the torque and flux control can still be indefinitely maintained. However, the flux is now clearly more distorted, due to errors in the flux calculation (observer). In fact, torque and flux control can even be maintained with locked rotor conditions down below 0.9-Hz stator frequency (Fig. 5), although the flux estimate now contains significant distortion. Fig. 6 shows the torque and flux control all the way down to 0.3 Hz, with a counterrotating rotor. It should be pointed out that, at this very low stator frequency, significant error exists in the field orientation, since the estimated torque is quite low, even though the slip frequency is high. With this condition, torque control was lost after about 30 s. This still represents satisfactory torque control for a reasonable period of time in many applications. It is also very important to note that many applications do not require the control of a low torque at zero speed, as is
the stator voltage, a reasonable flux estimate can be obtained only a small amount of low-pass filtering in the integration of frequency and 50-r/min rotor speed, while the bottom trace depicts the torque-producing component of the stator current. The stator-flux-oriented control algorithm, including space-vector PWM, is implemented on a TMS 32010 digital signal processor (DSP). The integration described by (5) is performed by analog circuitry. The result of this experiment is shown in Fig. 7. This paper has experimentally demonstrated that, by using only a small amount of low-pass filtering in the integration of the stator voltage, a reasonable flux estimate can be obtained for stator-flux field-oriented torque control. With this simple method, adequate torque control can be achieved, even at zero rotor speed, and down to 0.3-Hz synchronous frequency. The experimental results achieved in this paper show that this approach results in as wide an operating range as many other, more complex approaches that have been summarized in the paper.

IV. CONCLUSIONS

This paper has experimentally demonstrated that, by using only a small amount of low-pass filtering in the integration of the stator voltage, a reasonable flux estimate can be obtained for stator-flux field-oriented torque control. With this simple method, adequate torque control can be achieved, even at zero rotor speed, and down to 0.3-Hz synchronous frequency. The experimental results achieved in this paper show that this approach results in as wide an operating range as many other, more complex approaches that have been summarized in the paper.

REFERENCES


Kevin D. Hurst (S’92–M’96) received the B.S. degree in electrical engineering from Massachusetts Institute of Technology, Cambridge, and the M.S. and Ph.D. degrees from Georgia Institute of Technology, Atlanta, in 1987, 1993, and 1996, respectively.

He was commissioned as an officer in the U.S. Navy in 1991 and subsequently served for four years on the U.S.S. Fulton (AS-11). He is currently with the Allison Transmission Division, General Motors Corporation, Indianapolis, IN. Prior to this, he was a Senior Power Electronics Engineer with Sundstrand Aerospace, Rockford, IL.
Thomas G. Habetler (S’82–M’83–SM’92) received the B.S.E.E. and M.S. degrees in electrical engineering from Marquette University, Milwaukee, WI, and the Ph.D. degree from the University of Wisconsin, Madison, in 1981, 1984, and 1989, respectively. From 1983 to 1985, he was with the Electro-Motive Division, General Motors Corporation, as a Project Engineer. While there, he was involved in the design of switching power supplies and voltage regulators for locomotive applications. He is currently an Associate Professor of Electrical Engineering, Georgia Institute of Technology, Atlanta. His research interests are in switching converter technology and electric machine protection and drives.

Dr. Habetler was co-recipient of the 1989 First Prize Paper Award and the 1991 Second Prize Paper Award of the Industrial Drives Committee, and the 1994 Second Prize Paper Award of the Electric Machines Committee of the IEEE Industry Applications Society. He is an Associate Editor of the IEEE TRANSACTIONS ON POWER ELECTRONICS, and also serves as Publications Chair of the IEEE Power Electronics Society.

Giovanni Griva received the “Laurea” degree in electronic engineering and the Ph.D. degree in electrical engineering from the Politecnico di Torino, Torino, Italy, in 1990 and 1994, respectively. In 1995, he joined the Department of Electrical Engineering, Politecnico di Torino, as a Research Assistant. His fields of interest are power electronics conversion, integrated electronic/electromechanical design, high-performance speed servo drives, and applications of new power devices. He has published numerous papers in international conferences and technical journals.

Dr. Griva was the recipient of the IEEE Industry Applications Society First Prize Paper Award in 1992. He serves as a Reviewer for the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS and the IEEE TRANSACTIONS ON POWER ELECTRONICS.

Francesco Profumo (M’88–SM’90) was born in Savona, Italy, in 1953. He received the “Laurea” degree in electrical engineering from the Politecnico di Torino, Torino, Italy, in 1977. From 1978 to 1984, he was a Senior Engineer with the T&D Ansaldo Group, Genova, Italy. In 1984, he joined the Department of Electrical Engineering, Politecnico di Torino, where he was an Associate Professor until 1995. He is currently a Professor of Electrical Machines and Drives at Politecnico di Torino and an Adjunct Professor at the University of Bologna, Bologna, Italy. He was a Visiting Professor in the Department of Electrical and Computer Engineering, University of Wisconsin, Madison, during 1986–1988 and in the Department of Electrical Engineering and Computer Science, Nagasaki University, Nagasaki, Japan, during 1996–1997. His fields of interest are power electronics conversion, high-power devices, applications of new power devices, integrated electronic/electromechanical design, high-response speed servo drives, and new electrical machines structures. He has published more than 130 papers in international conferences and technical journals. He serves as a reviewer for the Proceedings of the Institution of Electrical Engineers, Part B, (U.K.) and the EPE Journal. He is member of the technical program committees of several international conferences in the power electronics and motor drives field. He has been a Coordinator or Partner for several projects in the frame of European Commission activities (Tempus, Comett, Joule, Human Capital and Mobility, Alfa, European Union S&T Grant Programme in Japan, and Leonardo da Vinci).

Dr. Profumo won the IEEE Industry Applications Society Second Prize Paper Award in 1991 and 1997 and the First Prize Paper Award in 1992. He is the Secretary of the Industrial Drives Committee of the IEEE Industry Applications Society and a Member at Large of the IEEE Power Electronics Society (PELS) AdCom Committee. He serves as a Reviewer for the IEEE Spectrum, IEEE TRANSACTIONS ON POWER ELECTRONICS, IEEE TRANSACTIONS ON INDUSTRY APPLICATIONS, IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS, and IEEE TRANSACTIONS ON MAGNETICS.

Dr. Profumo is a Registered Professional Engineer in Italy.