Review of Sensorless Methods for Brushless DC

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Abstract

This paper provides a review of the literature addressing sensorless operation of brushless DC machines. Methods explained are state-of-the-art, including those 1. using measured currents, voltages, fundamental machine equations, and algebraic manipulations, 2. using observers, 3. using back-emf methods, 4. sensorless starting techniques, and 5. with novel techniques not in the previous categories.

1. Introduction

The purpose of this paper is to provide a review of sensorless methods of BLDC. This paper will focus on the analytical and qualitative basis of these methods.

2. Sensorless Review

Many reviews of sensorless operation of AC drives have been seen in the literature [1,18,27,32,53]. Of publications presently available in IEEE and IEE conference proceedings and transactions, past sensorless techniques have been grouped into 5 categories:

1. those using measured currents, voltages, fundamental machine equations, and algebraic manipulations,
2. those using observers,
3. those using back-emf methods,
4. sensorless starting techniques, and
5. those with novel techniques not falling into the previous four categories.

2.1. Methods Using Measurables and Math

The first type above consists of three sub-types, 1. those using the voltages and currents to calculate flux linkages, 2. those using the difference between a model's prediction of a voltage or current and the actual value, and 3. those using the machine equations, meausurables, known machine parameters, and algebraic manipulations to calculate position and speed. The first sub-type is seen in [1,2,21,26,34,40,49,64]. The voltage equation of the machine,

\[ V = R I + \frac{d \Psi}{dt} \]

where,

\[ V \] is the voltage vector,
\[ I \] is the current vector,
\[ R \] is the resistance matrix, and
\[ \Psi \] is the flux linkage vector.

This equation is then manipulated to obtain,

\[ \Psi = \int \left( V - RI \right) dt. \]

Knowing the initial position, machine parameters, and the relationship of flux linkage to rotor position, the rotor position is estimated. Determining the rate of change of the flux linkage from the integration results, the speed is determined [34]. A variation includes using the previous position data and polynomial fitting, extrapolating to obtain the next step position prediction [21]. An advantage of the flux-calculating method, as shown in [34] is that line-line voltages may be used in the calculations and thus no motor neutral is required. Others, have combined the flux-calculating method with advanced control strategies such as a state observer to compute the speed feedback signal [2], an adaptive control [26], and \( H^2 \) and \( H^\infty \) controls [40].

The second sub-type was published primarily by one author, N. Matsui, in [17,20,39,45,70]. Using a d-q model of the machine, the actual d-q transformed currents and voltages, and those on a hypothetical axis offset from the d-q axis by a small angle \( \Delta \theta \), the output voltages of the model, on the hypothetical axis, and those on the actual d-q axis were compared. The differences between the calculated voltages of the hypothetical axis considering \( \Delta \theta = 0 \), and the actual d-q axis voltages on the hypothetical axis,

\[ \Delta v = v_{dq} - \psi_{hypothetical,dq} = K E \dot{\theta} \sin(\Delta \theta), \]

yielded \( \Delta \theta \), the actual change in rotor position from the previously known position, as with \( \Delta \theta \approx 0, \Delta \theta \approx \Delta v \).

The third sub-type uses machine parameters and equations, meausurables, and algebra, reference frame theory
and transformations to calculate position and speed [25]. Initially the measured voltages and currents are transformed to rotor and stator d-q reference frames. Relations between stator and rotor reference frames, denoted with superscripts s and r, respectively, are

\[
\begin{align*}
\nu_d &= \nu_d^s \cos \theta_r - \nu_q^s \sin \theta_r \\
\nu_q &= \nu_q^s \sin \theta_r + \nu_d^s \cos \theta_r \\
i_d &= i_d^s \cos \theta_r - i_q^s \sin \theta_r \\
i_q &= i_q^s \sin \theta_r + i_d^s \cos \theta_r
\end{align*}
\]

These relations allow the substitution of stator reference frame quantities in the rotor reference frame voltage equations. Once the substitutions are made and the equations relating rotor and stator reference frames are totally in terms of stator variables, in the stator frame, abc frame equivalents are substituted in these equations, which are then manipulated to yield rotor angle. Knowing that the machine is non-salient, the equivalent rotor flux, and manipulating the same equations, the rotor speed is also determined. The final equations for position and speed are as follows:

\[
\begin{align*}
x[k+1] &= A x[k] + B u[k] \\
y[k] &= C x[k] + D u[k]
\end{align*}
\]

which must be completely observable [73].

Next, a modal matrix must be determined for the system. A good reference to obtain the knowledge of this procedure is [74]. Using the modal matrix, P, the system is transformed into an observable form:

\[
\begin{align*}
x[k] &= P x'[k] = Q^{-1} x[k] \\
x'[k] &= P^{-1} x[k]
\end{align*}
\]

where

\[
\begin{align*}
A' &= QAQ^{-1} \\
B' &= QB \\
c'^T &= c^T Q^{-1}
\end{align*}
\]

Next, the full-order observer is designed using the designers choice of eigenvalues. The eigenvalues are usually chosen to be slightly faster (of a larger negative value) than those of the actual system so the state estimate error approaches zero as time approaches infinity. If the observer eigenvalues are chosen to be an order of magnitude larger, or more, than the actual system’s, then the estimated values will converge to the actual values within a sufficiently short time [24]. The observer takes the form

\[
\begin{align*}
z'[k+1] &= F z'[k] + g y[k] + H' u[k]
\end{align*}
\]

where

\[
\begin{align*}
F &= A' - g' c'^T \\
H' &= B' - g' d'^T
\end{align*}
\]

with the observer output equation

\[
w[k] = P z'[k] = Q^{-1} w[k]
\]

which allows the transformations

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\[ Z[k] = P\zeta[k] = Q^{-1}z[k] \]
\[ \zeta'[k] = P^{-1}\zeta'[k] = Q\zeta'[k] \]
which gives
\[ \zeta[k+1] = F\zeta[k] + g\zeta[k] + Hu[k] \]
where
\[ g = Q^{-1}g \]
\[ F = Q^{-1}A - Q^{-1}Q_d = A - gT \]
\[ H = Q^{-1}B - Q^{-1}gT = B - gT \]
which is a full-order observer in \( x[k] \), meaning that \( \zeta[k] \) converges to \( x[k] \).

Disturbance observers implemented in the cited papers, were reduced-order observers used with system states which were assumed quasi-static (fast sampling and calculating times). The slight variations of the state variables from one sampling to the next is termed a disturbance, which results in observations of small changes in rotor position angles. Reduced-order observers depend on some of the system outputs being linear transformations of the system states. By coupling some of the system states to the output of the observers through linear transformations (or actual measurements if the states are readily available for measurement), only the states of interest need be estimated by the lower-order observer. This reduces the computational burden of the observer [41]. In [1,5,24], reduced-order observers estimate the back-emf which is used to obtain rotor position and speed. In [41], the rotor position and speed are estimated as the outputs of a non-linear reduced-order observer. Sliding-mode observers, offspring of sliding-mode control, utilize a varying switching function in order to confine the state estimation error to move toward zero on a phase-plane sliding surface [3]. In [16] and [47], the sliding-mode observers were based on the observation of d-q transformed stator currents. From the resulting current estimations, the back-emfs were estimated, producing speed and position estimates. In [48], the full-order system model was utilized and a reduced-order sliding-mode observer described. In [3], a full-order sliding-mode observer was described, and modifications were made in order to overcome chattering problems, inherent in sliding-mode systems. In general, the sliding mode system starts with the state-space system [3,16,47,48]

\[ \hat{X} = f(X, U) \]
\[ Y = CX \]
where the usual definitions are made. In BLDC systems (usually considering the PMSM or the fundamental of the trapezoidal BLDC), \( X \) consists of the d-q transformed stator currents, speed, and position, i.e. \([i_d, i_q, \omega, \theta]\), or just the currents, as previously stated. The state observer takes the same form with the addition of a term containing the sign of the estimation errors:

\[ \dot{\hat{X}} = f(\hat{X}, U) - K \cdot \text{sgn}(S), \]
where
\[ S = \text{sgn}(X) = \begin{cases} 1 & X \geq 0 \\ -1 & X < 0 \end{cases} \]

\[ S \] is the vector of state estimation errors, i.e. \([\hat{x}_1 - x_1, \hat{x}_2 - x_2, \ldots, \hat{x}_n - x_n]^T\].

\( n \) is the observer order, and the superscript \( T \) denotes a transpose operation. In order for the observer to operate in the sliding-mode, the following condition must be met:

\[ \frac{1}{2} \frac{d}{dt} S^T S < 0. \]

Defining \( \dot{X} - X = e \), where \( e \) is the state estimation error, the general error model and dynamics are described by

\[ \dot{e} = f'(e, U) + K \cdot \text{sgn}(e) \]

where
\[ f' \] is the system representing the difference \( f(\hat{X}, U) - f(X, U) \), and disturbances due to parameter variations have been neglected. Now, in order for convergence, \( K \) must be continuously selected to insure that
\[ e^T e < 0, \text{ as } \frac{1}{2} \frac{d}{dt} S^T S = e^T e. \]

### 2.3. Methods Using Back-Emf Sensing

#### 2.3.1. Terminal Voltage Sensing

In field-oriented operation of the BLDC, phase back-emf is aligned with phase current. Switching instants of the converter can be obtained by knowing the zero-crossing of the back-emf and a speed-dependent period of time delay [75]. The BLDC motor has a trapezoidal flux density distribution in the air-gap, and thus the induced back-emf in the stator windings is trapezoidal. Monitoring the phase back-emf when the particular phase current is zero (the silent phase), the zero crossing is detected. Low pass filters are used to eliminate...
higher harmonics in the terminal voltages. The zero-crossing instants are decoded and appropriately time-shifted to produce switching patterns.

Since back-emf is zero at rest and proportional to speed, the terminal voltage sensing method is not possible at low speeds. As the speed increases, the average terminal voltage increases, and the frequency of excitation increases. Capacitive reactance in the filters varies with the frequency of excitation, introducing a speed-dependent delay in switching instants, which disturbs current alignment with the back-emf and field orientation thus, causes problems at higher speeds. With this method, a reduced speed operating range is normally used, typically around 1000-6000 rpm.

Another method which similarly has a narrow speed range is a method using phase-locked loop circuitry. The phase-locked loop has a narrow speed range due to the capabilities of the phase detector, and is sensitive to switching noise. This method, provides gating signals much like hall-effect sensor outputs by using a phase-locked loop circuit which locks onto the back emf of the inactive phase during every 60° period.

2.3.2. Third Harmonic Back-emf Sensing

The third harmonic of the back-emf can be used in determination of the switching instants in the wye connected 120 degree current conduction operating mode of the BLDC motor [37]. This method is not as sensitive to phase delay as the zero-crossing method, as the frequency to be filtered is three times as high. The reactance of the filter capacitor in this case dominates the phase angle output of the filter more so than with the lower frequency. On the other hand, greater than 2/3 stator winding pole pitch is required. Provided that this condition is satisfied, the third harmonic of the rotor flux can link the stator winding and induces a third harmonic voltage component which can be detected between the motor neutral and the negative inverter bus terminal.

The terminal voltage equation of the BLDC motor for phase a can be expressed as

\[ v_a = R_i a + L_s \frac{di_a}{dt} + e_a \]

Back-emf voltage, \( e_a \), is comprised of many voltage harmonic components,

\[ e_a = E(\cos \omega_r t + k_3 \cos 3\omega_r t + k_5 \cos 5\omega_r t + k_7 \cos 7\omega_r t + \ldots) \]

The third harmonic of the terminal voltages is acquired by the summation of the terminal voltages.

\[ v_{a3} + v_{b3} + v_{c3} = 3E k_3 \cos 3\omega_r t + v_{\text{high freq}} \]

The summed terminal voltages contains only the third and the multiples of the third harmonic due to the fact that only zero sequence current components can flow through the motor neutral. This voltage is dominated by the third harmonic, which does not necessitate much filtering. To obtain switching instants, the filtered voltage signal which provides the third harmonic voltage component is integrated to find the third harmonic flux linkage.

\[ \lambda_{r3} = \int v_3 dt \]

The third harmonic flux linkage lags the third harmonic of the phase back-emf voltages by 30 degree. The zero crossings of the third harmonic of the flux linkage corresponds to the commutation instants of the BLDC motor. To acquire correct commutation instants, sensing the positive or negative going zero-crossing of the back-emf is essential.

The third harmonic method provides a wider speed range (100-6000 rpm) than the zero-crossing method, does not introduce as much phase delay as the zero-crossing method and requires less filtering.

2.3.3. Freewheeling Diode Conduction

This method uses indirect sensing of the zero-crossing of the phase back-emf to obtain the switching instants of the BLDC motor [65]. In the 120 degree conducting wye-connected BLDC motor, one of the phases is always open-circuited. For a short period after opening the phase, there remains phase current flowing, via a freewheeling diode. This open phase current becomes zero in the middle of the commutation interval, which corresponds to the point where back-emf of the open phase crosses zero. The biggest downfall of this method is the requirement of six additional isolated power supplies for the comparator circuitry for each free-wheeling diode.

By this technique, 45-2300 rpm sensorless operation has been achieved. This technique outperforms the previously mentioned back-emf methods at low-speeds.

2.3.4. Back-emf Integration

In this method position information is extracted by integrating the back-emf of the unexcited phase [30,52,66,69]. The integration is based on the absolute value of the open phase's back-emf. Integration of the back-emf starts when the open phase's back-emf crosses zero. A threshold is set to stop the integration which corresponds to a commutation instant. As the back-emf is assumed to vary linearly from positive to negative (trapezoidal back-emf assumed), and this linear slope is assumed speed-insensitive, the threshold voltage is kept constant throughout the speed range. If desired, current advance can be implemented by changing the threshold. Once the integrated value reaches the threshold voltage, a reset

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signal is asserted to zero the integrator output. To prevent the integrator from starting to integrate again, the reset signal is kept on long enough to insure that the integrator does not start until the residual current in the open phase has passed a zero-crossing. This type of control algorithm has been implemented commercially. The integration approach is less sensitive to switching noise, automatically adjusts to speed changes, but the low speed operation is poor. With this type of sensorless operation scheme, up to 3600 rpm has been reported [69].

2.4. Starting Techniques

To operate a BLDC without a rotor position sensor, requires some method of starting the motor. Published startup methods have belonged to one of five general types: 1. open-loop, 2. with known initial position, 3 using a machine interrogation and signal processing, 4. using computationally complex methods, and 5. those relying on a winding inductance which varies with rotor position due to saliency.

2.4.1. Open-loop Starting

Open-loop [29,65,69] starting the motor is accomplished by providing a rotating stator field which increases gradually in magnitude and/or frequency. Once the rotor field begins to become attracted to the stator field enough to overcome friction and inertia, the rotor begins to turn and the motor acts as a permanent magnet synchronous machine. This type of system may be satisfactory in certain applications, i.e. pump and fan drives. The disadvantage of this method is that the initial rotor movement direction is not predictable. When the stator field becomes just strong enough, the rotor could move in either direction. If then the speed of the stator field is slow enough, and the load torque demanded does not exceed the pull-out torque, the motor will operate synchronously in the desired direction. Initial direction in certain applications such as disk drives strictly require unidirectional motion, and thus this method is inadequate. Another problem exists if the stator field is rotating at too great a speed when the rotor field picks up. This causes the rotor to oscillate, requiring the stator field to decrease in frequency to allow starting.

2.4.2. Starting Methods Using a Known Initial Position

Methods which require a known position [1,48,64,72] to start the motor suffer from an additional constraint, e.g. a detent in which the rotor always freely stops, or a stopping circuit (and initial positioning) which always provides an initial position for the next movement. A simple way of setting an initial position is to excite one phase to cause the rotor to shift and lock into position.

2.4.3. Interrogation Techniques for Starting

Machine interrogation techniques [7,8,35,70] offer an excellent solution for drives where absolute position information is needed at startup. In these techniques, initial rotational direction is controlled. Saliency of the interrogated machine enhances the ability to detect rotor position. Usually voltage pulses are imposed on the machine terminals while monitoring the current peak or level after a time duration from the start of the pulse, often a period approximately equal to the electrical time constant of the windings. The flux of the permanent magnet, either aids or opposes the flux due to the pulse and the current level indicates their interaction. Possibilities include using the sign of the difference in the current levels providing a sign vector which is compared to values in a lookup table to obtain position information [7,8,32]. Alternatively, interrogation of the phase combinations with a salient machine using short (non-saturating) and long (saturating) pulses has been reported [35,70].

2.4.4. Control System Techniques for Starting

In [28], a Kalman filter estimates rotor position from measured voltages and currents. The machine equations of non-sinusoidal waveforms are in a d-q stator reference frame. At zero-speed the voltages and currents are zero and alone the Kalman filter cannot provide rotor position information. By injecting a d-axis current this condition is overcome and position information at standstill is obtained.

2.4.5. Starting Using Machine Inductance Saliency

Saliency in permanent magnet machines is usually present due to the use of interior permanent magnet (PM) machine design. The interior PM allows a shorter air gap which increases the flux intensity on the stator due to the rotor. A thin bridge which usually separates the magnet from the gap is often saturated during normal operation, increasing the reluctance of the radial area where the magnets (µ = 1) already exist. This allows a varying inductance profile for the motor which depends on rotor position. Sensorless operations utilizing the relationships between voltage pulses, current amplitudes, rate of current rises, etc. have often been used in switched reluctance and synchronous reluctance motor drives [54-69] where saliency is inherent, and are applicable in interior PM BLDC drives as well.

2.4.6. I.C.'s for Sensorless BLDC Control

In the past there have been several integrated circuits which enabled sensorless operation of the BLDC. These included Unitrode's UC3646, Microlinear's ML4425, and
Silicon Systems, Inc. 32M595. Each of these used back-emf methods and open-loop starting.

2.5. Unique Sensorless Techniques

The following sensorless methods are unique. These range from artificial intelligence methods to variations in the machine structure. There are few which fall into this class, and therefore each is briefly discussed.

A backpropagation neural network (BPN) was trained to provide a non-linear mapping between measured voltages and currents, and rotor position [6]. This rotor position was then used with the equations of section 2.1 to calculate the flux linkage vector. This vector was compared to the estimated flux linkage vector using the equations of section 2.3.1 and the measured voltages and currents. The difference was used as the error which was backpropagated to modify the BPN weights. An initial flux linkage vector (initial rotor position) was required to be known a priori.

A paper using fuzzy logic [85] proposed two fuzzy logic sub-systems. Using the conventional equations consisting of phase voltages and currents, the rotor position was calculated. With the knowledge of the relationships between these measurables and the rotor position, a fuzzy Mamdani type system was developed to produce rotor position estimates. It was noted that this could as easily been accomplished with look-up tables, however, for the desired resolution the size of the look-up tables becomes unmanageably large. The second fuzzy system used took as input the estimated rotor position and produced reference current values for two different drive strategies: unity power factor, and maximum torque per amp.

In [15], a sensorless method was developed using the calculation of flux linkage by integrating the back-emf. The motor was at the end of a 10 km cable. A difficulty in this sensorless method was that the back-emf had high-frequency oscillations superimposed on the voltage waveforms at points of current commutations. This was overcome by integrating the back-emf over the whole 60° silent phase period which produced pulses with amplitudes proportional to the speed and at positions where the peak of the pulses was 30° prior to a commutation instant.

Solutions to sensorless operation of BLDC by means of alterations in machine design were shown in [68,71]. In [68], an additional stator lamination with equally spaced slots around the periphery was added to the machine. Each of the equally spaced slots contained a small sensing coil. The local magnetic circuit variations for each of the sensing coils is effected by the permanent magnet rotor's position. A 20 kHz signal was injected through the coils, and the signal distortions were analyzed at the terminals of the sensing coils. The second harmonic detected provided position information. An artificial saliency was created in [71], by attaching small pieces of aluminum to the surface of the permanent magnets allowing eddy currents to flow (in the aluminum). The flow of eddy currents in the aluminum acted to increase the reluctance of the magnetic circuits creating saliency.

Two papers describe sensorless methods which rely on irregularities in the measurements [54,67]. These methods utilize small deviations, or inherent properties of the shapes of the current waveforms to obtain position information or enable a control which tends to provide a phase current / phase back-emf alignment. The method in [67] considers the shape of the D.C. bus current. The authors show that the shape of the D.C. bus current varies depending on the alignment of the phase current and phase back-emf. A clear difference in the distortion patterns is seen between a leading, aligned, and lagging operation. The method of control is to count the number of samples on a consecutive rising and falling half-cycle of the 6-pulse D.C. current ripple. When these are equal, the machine is operating at unity internal power factor. When unaligned, the difference in the rising and falling counts is used as a multiplier in the D.C. link voltage incremental update. The machine is operated as a synchronous machine with this control effort to maintain phase current / back-emf alignment.

The second method, in [54], works on the principle of machine saliency. In this paper, the crest of the current waveform, which is controlled by hysteresis chopping, is the medium of rotor position information gathering. The elementary equations of the rise and fall of current in the hysteresis waveform were examined. In these equations, it is easily seen that there is a dependency on the instantaneous inductance, as well as other machine parameters. Utilizing a fast sampling and these equations, while considering the other machine and drive parameters constant, and with measured current and voltages, the instantaneous inductance, and thus position is determined. This is possible only if the machine has inductance saliency.

3. Conclusion

A review of prior work in rotor position sensorless methods for brushless D.C. has been presented. The basis of these techniques has been presented in an effort to provide a reference for utilizing these methods. A new method of sensorless operation which builds on those presented is revealed in another paper in these proceedings.

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