Direct Torque and Flux Regulation in Sensorless Control of an Induction Motor

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Abstract

In this paper, a novel scheme for direct torque and flux control of induction motors (IM) is presented. The scheme does not use a transformation to the rotor flux related frame and it guarantees the asymptotic stability of the torque and flux tracking. To perform controller design a torque-flux dynamic model with stator voltage as an input is obtained and separation of the torque and flux regulation is achieved. The controller is elaborated as a part of the sensorless control scheme of IM and it shows robustness with respect to rotor time constant changes. For slow reference commands a control scheme based on the steady state solution of the torque-flux model is also analyzed. It is significantly simpler than the dynamic scheme and has better robustness properties with respect to the current noise. The controllers can also be applied if the rotor speed signal is available. The analysis of the stator current and voltage constraints determines the allowable set of the torque and flux reference commands. Simulation results are presented to verify the proposed approach.

1 Introduction

In the vector control of induction motors (IM) the separation of the torque and flux regulation is achieved in the rotor flux related frame [1]. In the indirect field oriented (IFO) scheme the orientation of the rotor flux is obtained by integrating the rotor speed signal and commanded slip frequency. Although the IFO scheme stability is difficult to prove analytically it is widely used in practical applications.

An alternative way of the vector control implementation is based on the rotor flux observers. An observer that employs natural rotor flux dynamics was designed in [2]. Using this observer, an adaptive controller for position tracking of the robotic manipulators was elaborated in [3]. A robust nonlinear controller with flux observer and real time adaptation of the rotor resistance was developed in [4]. Sliding mode flux and speed observers and torque controllers of IM were designed in [5] with conditions of asymptotic stability of control being obtained.

In this paper, a novel scheme for direct torque and flux regulation of IM is described. The scheme does not use a transformation to the rotor flux related frame and it provides asymptotic tracking of the torque and flux commands. In the speed sensorless mode of operation the scheme shows robustness to the rotor time constant variation that could improve an electric drive efficiency as compared to the commonly used IFO scheme.

The controller design is based on a torque-flux dynamic model that has torque, flux magnitude and one auxiliary signal as state variables and the stator voltage as the inputs. In the first design method the natural dynamics of the torque and flux are cancelled out and the torque and flux are forced to the dynamics of the given reference models. The scheme is complimented with a rotor flux observer based on the rotor flux dynamic equation.

In the second design method a steady-state solution of the torque-flux model is calculated as a feedforward part of the control. In comparison with the first method the scheme is simpler and has better robustness to the current noise.

The paper is organized as follows. Section 2 discusses the dynamic model and control problem statement. The main results of direct torque and flux control design are given in Section 3. Section 4 is devoted to the analysis of the current and voltage constraints with controllers simulation results being presented in Section 5.

Relevant results obtained by the authors on speed sensorless control and parameters estimation of IM are given in [6]-[9].
2 Dynamic model of induction motors and control problems

Consider the dynamic model of induction motors (IM) in the stator frame [1]

\[
\frac{d\omega}{dt} = \frac{\mu T_I J_\lambda r - \alpha \omega + T_L}{m}, \quad (1)
\]

\[
\frac{d\lambda_r}{dt} = \left( -\frac{R_r}{L_r} + \frac{n_p \omega J}{L_r} \right) \lambda_r + \frac{R_r}{L_r} M I_s
\]

\[
\frac{di_s}{dt} = -\frac{M}{\sigma L_s L_r} \left( \frac{R_r}{L_r} I + \frac{n_p \omega J}{L_r} \right) \lambda_r - \frac{1}{\sigma L_s} (R_s + \frac{M^2 R_r}{L_r^2}) i_s + \frac{1}{\sigma L_s} v_s, \quad (3)
\]

where

\[
I = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}, \quad J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}.
\]

Here \( \omega \) is the angular speed of the rotor, \( \lambda_r, i_s \) and \( v_s \) are the rotor flux, the stator current and the voltage, respectively; \( T_e = \mu i_T J_\lambda r \) is the electromagnetic torque, \( T_L \) is the external load torque. Parameters \( R_r \) and \( R_s \) are the rotor and stator resistances, \( M \) is the mutual inductance, \( L_r \) and \( L_s \) are the rotor and stator inductances, \( \sigma = 1 - M^2/L_s L_r \) is the leakage parameter, \( n_p \) is the number of pole pairs, \( m \) is the moment of inertia of the rotor, \( \alpha \) is the damping gain, and \( \mu = 3n_p M/2L_r m \).

The first control goal is for the electromagnetic torque to follow the reference value \( T_{ref} \)

\[
\lim_{t \to \infty} (T_e - T_{ref}) = 0.
\]

To achieve this goal the flux magnitude is to be kept at a certain level and the second control goal is

\[
\lim_{t \to \infty} (|\lambda_r|^2 - F_{ref}) = 0,
\]

where \( F_{ref} \) is the square of the flux reference value. It is assumed that the reference values \( T_{ref}, F_{ref} \in C^1[\mathbb{R}^+] \) and they should be selected accounting for constraints on the voltage and current signals. The detailed analysis of the effect of voltage and current constraints on control design is given in Section 4.

In the direct torque and flux regulation the value of the rotor flux or its estimate is used. Thus the next problem is to construct an observer for the rotor flux. In many applications it is desirable to avoid measurements of the rotor position or speed (such sensors make the system expensive and less reliable). Thus another problem of estimation of the rotor speed from the available for the measurement stator current and stator voltage command arises.

This paper is devoted to the design of torque and flux controllers assuming that the rotor speed observer and flux observer are designed and motor parameters are known or properly estimated [6]-[8].

3 Direct torque and flux control

In this section the direct torque and flux regulation schemes are designed based on the new dynamic equations for the electromagnetic torque and rotor flux. The scheme does not require a transformation to the d-q rotating frame and, in the simplified version, it behaves as a PI regulator of the torque and a PD regulator of the rotor flux.

3.1 Dynamic equation for the electromagnetic torque and flux magnitude

To derive the dynamic equations for the electromagnetic torque the derivative of \( T_e \) is calculated

\[
\frac{dT_e}{dt} = \mu \left( i_T^T J_\lambda r - \frac{L_s}{\sigma L_r} \frac{d\lambda_r}{dt} \right).
\]

Multiplying equations (2) and (3) by \( i_T^T J \) and \( \lambda_T^T J \) respectively, and then substituting the results into (6) yield the following equation for \( T_e \):

\[
\frac{dT_e}{dt} = -k T_e - n_p \omega \left( \frac{M}{\sigma L_s L_r} |\lambda_r|^2 + \lambda_T^T \frac{L_s}{i_T^T} \right) - \frac{\mu}{\sigma L_s} \lambda_T^T J v_s
\]

Dynamic equation for \( |\lambda_r|^2 \) is obtained in a similar way. Multiplying (2) by \( \lambda_T^T \) and utilizing the equalities

\[
\frac{d|\lambda_r|^2}{dt} = 2 \lambda_T^T \frac{d\lambda_r}{dt}, \quad \lambda_T^T J \lambda_r \equiv 0
\]

yields the expression for the derivative of \( |\lambda_r|^2 \):

\[
\frac{d|\lambda_r|^2}{dt} = -2 \frac{R_r}{L_r} |\lambda_r|^2 + 2 \frac{R_r}{L_r} M i_T^2 \lambda_r.
\]

Equation (8) does not contain the voltage as an input and the dynamic equation for the fictitious variable \( (i_T^T \lambda_r) \) should be added. The derivative of \( i_T^T \lambda_r \) may be calculated by multiplying equations (2) and (3) by \( i_T^T \) and \( \lambda_T^T \), respectively, and then summing them up

\[
\frac{di_T^T \lambda_r}{dt} = -k i_T^T \lambda_r + \frac{R_r M}{L_r} |i_s|^2 + \frac{M}{\sigma L_s L_r} \frac{R_r}{L_r} |\lambda_r|^2 + \frac{n_p \omega}{\mu} T_e + \frac{1}{\sigma L_s} \lambda_T^T J v_s.
\]

By introducing the new state vector \( Z = [z_1, z_2, z_3]^T = [T_e, |\lambda_r|^2, i_T^T \lambda_r]^T \) equations (7)-(9) can be rewritten in the following form:

\[
\frac{dZ}{dt} = A(\omega) Z + \begin{bmatrix}
- \frac{n_p \omega}{\sigma L_s} \lambda_T^T J v_s \\
0 \\
\frac{R_r M}{L_r} |i_s|^2 + \frac{1}{\sigma L_s} \lambda_T^T J v_s
\end{bmatrix},
\]

\[
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\]
where

\[
A(\omega) = \begin{bmatrix}
-k - \frac{\alpha M n p \omega}{\sigma L_e L_r} & -\mu n p \omega \\
0 & -\frac{2R_e}{L_r} & \frac{2R_e M}{L_r} \\
\frac{\nu p \omega}{\mu} & \frac{M R_e}{\sigma L_e L_r} & -k
\end{bmatrix}.
\]

Finally, perform decomposition of the stator voltage command

\[
v_s = \frac{\lambda_r}{|\lambda_r|} v_{sd} + \frac{J \lambda_r}{|\lambda_r|} v_{sq}
\]

which will be used in the sequel.

### 3.2 Rotor flux and torque control

To achieve the first control goal (4) of electromagnetic torque tracking the stator voltage command is chosen such that the torque \(T_e\) satisfies the following equation

\[
d(T_e - T_{ref}) = -LT_p(T_e - T_{ref}) - LT_I \int_0^t (T_e - T_{ref}) dt.
\]

where \(L_{TP} > 0\) and \(L_{TI} \geq 0\) are gains of the PI regulator selected to guarantee the exponential convergence of \(T_e\) to the reference value with an appropriate performance. This can be done by cancelling the natural dynamics of \(T_e\) in (7) and setting the needed one corresponding to equation (12). For this purpose the right-hand side of (7) must satisfy the equality

\[
-kT_e - \mu \lambda_r^T \left( \frac{M}{\sigma L_e L_r} \lambda_r + i_s \right) n p \omega - \frac{\mu}{\sigma L_s} \lambda_r^T J v_s = 0.
\]

Taking into consideration decomposition (11) the last expression uniquely determines the \(v_{sq}\) component of the voltage signal

\[
v_{sq} = \frac{\sigma L_r}{\mu |\lambda_r|} \left[ kT_e + \frac{d T_{ref}}{dt} + \lambda_r^T \left( \frac{M}{\sigma L_e L_r} \lambda_r + i_s \right) \mu n p \omega \right]
\]

\[
+ \frac{\sigma L_r}{\mu |\lambda_r|} \left[ -L_{TP}(T_e - T_{ref}) - L_{TI} \int_0^t (T_e - T_{ref}) dt \right].
\]

Thus only the component \(v_{sq}\) of the stator voltage that is orthogonal to the rotor flux is responsible for torque control. Note that the magnitude of the rotor flux in the denominator of (13) can make the voltage signal high if the flux is not regulated properly.

The addendum with the derivative of the reference torque \(dT_{ref}/dt\) guarantees that the electromagnetic torque converges to \(T_{ref}\) for nonconstant \(T_{ref}\). If this term is omitted, then the gains \(L_{TP}, L_{TI}\) of the PI controller should be chosen properly (big enough) to preserve the servo property of the closed loop system.

The second control goal employed in the torque control of IM is the tracking of the rotor flux magnitude. In this case the squared norm of the rotor flux \(|\lambda_r|^2\) is to be kept at a desirable, e.g., constant level \(F_{ref}\).

To make the control input \(v_s\) appear explicitly in the equation of flux dynamics, differentiate equation (8)

\[
\frac{d^2 |\lambda_r|^2}{dt^2} = \frac{L_{TP}}{L_r} \left( -\frac{d}{dt} |\lambda_r|^2 + M \frac{d}{dt} (T^T \lambda_r) \right).
\]

Utilizing (9) the final formula for the rotor flux \(|\lambda_r|^2\) is obtained

\[
\frac{d^2 |\lambda_r|^2}{dt^2} = \frac{2R_e}{L_r} \left[ -\frac{d}{dt} |\lambda_r|^2 + M - \left( k \lambda_r^T \lambda_r + \frac{R_e M}{L_r} |i_s|^2 \right) + \frac{M^2 R_e}{\sigma L_e L_r} |\lambda_r|^2 + \frac{\mu n p \omega}{\mu} T_e + \frac{1}{\sigma L_s} \lambda_r^T v_s \right].
\]

The stator voltage command \(v_s\) is to be chosen in such a way that \(|\lambda_r|^2\) satisfies the following reference equation

\[
\frac{d^2 |\lambda_r|^2}{dt^2} = \frac{2R_e}{L_r} \left( -\frac{d}{dt} |\lambda_r|^2 - L_{FP} (|\lambda_r|^2 - F_{ref}) \right).
\]

In this case the controller will be of PD type. By proper selection of the gains \(L_{FD} > 0, L_{FP} > 0\) the exponential convergence of \(|\lambda_r|^2\) to \(F_{ref}\) can be achieved. It follows from (15), (16) and (11) that the corresponding \(v_{sq}\) component of the stator voltage equals

\[
v_{sd} = \frac{\sigma L_r}{M |\lambda_r|} \left[ \left( (L_{FD} - 1) \frac{2R_e}{L_r} - \frac{M^2 R_e}{\sigma L_e L_r} - L_{FP} \right) |\lambda_r|^2 + M \left( k + (1 - L_{FD}) \frac{R_e M}{L_r} \right) |i_s|^2 \lambda_r^T - \frac{R_e}{L_r} M^2 |i_s|^2 - \frac{M^2 R_e}{\sigma L_e L_r} \lambda_r^T v_s \right].
\]

The obtained formula for control of rotor flux magnitude shows that only the \(v_{sd}\) component of the stator voltage commands the variable \(|\lambda_r|^2\). Since different components of the stator voltage are responsible for torque control and flux magnitude control, these two goals can be achieved simultaneously. In this case stator voltage command must be chosen as (11) where \(v_{sd}\) and \(v_{sq}\) are defined according to (13) and (17).

The condition that \(F_{ref}\) is constant may be omitted. In this case it is possible to preserve the servo property of the controller by increasing the gains \(L_{FD}, L_{FP}\).

The value of the rotor flux \(\lambda_r\) used in this control algorithm is not available for measurement. Thus its estimate \(\hat{\lambda}_r\) must be utilized instead. Measuring or estimating the rotor speed, the rotor flux observer based on
equation (2) can be constructed. Then the overall controller ("PIDB" – PID Based controller) has the form

$$\frac{d\lambda_r}{dt} = (-L_r + n_p\omega J)\lambda_r + R_r M i_s$$

(18)

where \(\dot{v}_{sd}\) and \(\dot{v}_{sq}\) are computed from (17) and (13) respectively with \(\dot{\lambda}_r\) substituted for \(\lambda_r\) and \(\dot{T}_e\) substituted for \(T_e\). To avoid the start-up singularities in \(\dot{v}_{sq}\) and \(\dot{v}_{sd}\) caused by the division by \(\lambda_r\) the initial condition of the flux estimate \(\dot{\lambda}_r\) is set to a non-zero value.

By neglecting nonlinear terms (17) and (13) are simplified as follows

$$\dot{v}_{sd} = -L_{FD} (\dot{\lambda}_r^2 + M L_r T_e) - L_{FP} (\dot{\lambda}_r^2 - F_{ref})$$

(20)

$$\dot{v}_{sq} = -L_{TP}(T_e - T_{ref}) - L_{TP} \int_0^t (T_e - T_{ref}) \, ds,$$

(21)

where \(\dot{v}_r\) is used to denote a new set of feedback gains. From (8) follows that the expression in first brackets in (20) is the derivative of the squared flux norm. Thus controller (20), (21) is a combination of PI controller for torque regulation and PD controller for flux regulation. By selecting the gains in (20), (21) high enough the performance of the simplified controller (20), (21) can be made close to that of nonlinear controller (17), (13).

If the rotor speed is not measured, the proposed controller utilizes speed estimates obtained by using known techniques [6] - [8].

3.3 Torque and flux control based on a steady state solution

Another control scheme can be obtained using a steady-state solution of the torque-flux model (10). The steady-state based (SSB) scheme is significantly simpler than the dynamic scheme from the previous section and it has better robust properties with respect to the current measurement noise. The controller design utilizes the natural dynamics of IM. For simplicity, it is assumed that the rotor speed is constant. The idea of the controller design is to calculate a steady state solution of (10) with a pseudo constant input. Since the state matrix is constant and stable the steady state point exists and can be calculated as a solution of an algebraic equation.

Denote

$$v_1 = \frac{\mu}{\sigma L_s} \lambda^T J s, \quad v_2 = \frac{R_s M}{L_r} i_s^2 + \frac{1}{\sigma L_s} \lambda^T v_s.$$ 

(22)

It follows from (10) that the values of \(v_1\) and \(v_2\) corresponding to the steady state point \(Z_{st} = [T_{ref}, F_{ref}, z_3]^T\) for some value of \(z_3\) are defined by the algebraic equation

$$A [T_{ref}, F_{ref}, z_3]^T = -[v_1, 0, v_2]^T.$$ 

(23)

Its solution with respect to \(v_1\) and \(v_2\) is given by the following formulae

$$v_1 = k T_{ref} + \mu \left(\frac{M}{\sigma L_s L_r} + \frac{1}{M}\right) F_{ref} n_p \omega,$$

(24)

$$v_2 = -\frac{n_p \omega}{\mu} T_{ref} + \left(\frac{R_s}{M L_r} + \frac{R_s}{\sigma L_s M}\right) F_{ref}.$$ 

Tracking back our notations (22) and utilizing decomposition (11) the expressions for \(d\) and \(q\) components of the stator voltage command are obtained

$$v_{sd} = \frac{\sigma L_s}{\lambda_r} \left(v_2 - \frac{R_s M}{L_r} i_s^2\right), \quad v_{sq} = \frac{\sigma L_s}{\mu \lambda_r} v_1.$$ 

(25)

Estimates of the rotor flux should be used instead of the hard to measure rotor flux signal. To lessen the needed information about the rotor flux the magnitude of \(\lambda_r\) is substituted by its reference value \(\lambda_r = \sqrt{F_{ref}}\) in (25).

Simulation results show that the torque and flux controller based on the steady-state solution ("SSB" controller) has good robust properties with respect to the stator current noise. On the other hand, it appears to be sensitive to the speed estimation error with biased mean that results in long transients. This sets additional requirements on the speed observer used with this scheme. If the rotor speed is measured, the "SSB" controller becomes attractive for use due to its simplicity.

4 Analysis of current and voltage constraints

The values of the allowable stator current and voltage signals are limited by IM design. Thus stator current and voltage constraints are to be taken into account for when the values of the reference torque \(T_{ref}\) and reference rotor flux magnitude \(F_{ref}\) are chosen. In this section this problem is analyzed for the case of constant \(T_{ref}\) and \(F_{ref}\). In particular the following question is answered: do the stator voltage and current satisfy the constraints for the steady-state operation \(||(\lambda_r)|^2 = F_{ref}, T_e = T_{ref}||?\)

First consider constraints on the magnitude of the stator current. Let the constraints be set in the following way

$$|i_{sd}| \leq C_{is}, \quad |i_{sq}| \leq C_{is},$$ 

(26)

where \(i_{sd}\) and \(i_{sq}\) are direct and quadrature components of \(i_s\). It follows from the definition of \(T_e\) that
Thus in the steady state $|i_{eq}| = |T_{ref}|/\mu F_{ref}$. To satisfy constraint (26) $T_{ref}$ and $F_{ref}$ must be chosen according to the inequality

$$|T_{ref}| \leq \mu \sqrt{F_{ref} C_{is}}. \quad (27)$$

Similar inequality is obtained from equation (8). In the steady state $|\lambda_r|^2 = M|\lambda_r|i_{sd}$. Thus $i_{sd} = \sqrt{F_{ref}/M}$ and the correspondent inequality is

$$\sqrt{F_{ref}} \leq C_{is} M. \quad (28)$$

Unifying inequalities (27) and (28) yields the following result

$$\frac{|T_{ref}|}{\mu C_{is}} \leq \sqrt{F_{ref}} \leq C_{is} M. \quad (29)$$

From (29) follows that $|T_{ref}|$ should be less then

$$|T_{ref}| \leq \mu M C_{is}^2. \quad (30)$$

Inequalities (29) are illustrated by plots in Figure 1 which define allowable values for the reference flux and minimal stator current

$$C_{im} = \sqrt{\frac{T_{ref}}{\mu M}} \quad (31)$$

for different values of the reference torque. The $C_{im}$ (31) is defined by the intersection of the curves from the left and right sides of (29). The motor parameters used in calculation are given in Table 1.

$$R_a \quad R_r \quad M \quad L_r \quad L_s$$

| 0.11 | 0.087 | 0.00081 | 0.0011 | 0.0011 |

Table 1

Now consider stator voltage constraints that are set in the same way

$$|v_{sd}| \leq C_{us} \quad |v_{sq}| \leq C_{us}. \quad (33)$$

Values of $d, q$ components of the stator voltage command at the steady-state are given by expressions (24), (25). Utilizing the equality $|i_{sd}|^2 = i_{sd}^2 + i_{sq}^2 = F_{ref}/M^2 + T_{ref}^2/(\mu^2 F_{ref})$ the expressions for the stator voltage command magnitude are obtained

$$v_{sq} = \sigma L_s \left[ \frac{k}{\mu} \sqrt{F_{ref}} + \sqrt{F_{ref}} \left( \frac{M}{\sigma L_s L_r} + \frac{1}{M} \right) n_{\omega} \right], \quad (34)$$

$$v_{sd} = \frac{\sigma L_s}{\sqrt{F_{ref}}} \left[ \frac{R_s}{\sigma L_s M} F_{ref} - \frac{R_r M T_{ref}^2}{\mu^2 L_r F_{ref}} - \omega \frac{n_{\omega} T_{ref}}{\mu} \right]. \quad (35)$$

The correspondent inequalities are

$$\frac{\sigma L_s}{\sqrt{F_{ref}}} \left| \frac{R_s}{\sigma L_s M} F_{ref} - \frac{R_r M T_{ref}^2}{\mu^2 L_r F_{ref}} - \omega \frac{n_{\omega} T_{ref}}{\mu} \right| \leq C_{us} \quad (36)$$

$$\sigma L_s \left[ \frac{k}{\mu} \sqrt{F_{ref}} + \sqrt{F_{ref}} \left( \frac{M}{\sigma L_s L_r} + \frac{1}{M} \right) n_{\omega} \right] \leq C_{us}. \quad (37)$$

Given the numbers $C_{us}$ and $C_{is}$ inequalities (29), (36) and (37) define all possible combinations of $\{T_{ref}, F_{ref}, \omega\}$ such that at the steady-state the stator current and stator voltage command satisfy constraints (26), (33).

5 Simulation results

The effectiveness of the proposed controllers has been verified via simulations in Matlab/Simulink. The rotor speed observer developed in [8] is utilized for the speed sensorless direct torque and flux regulation. Parameters of the "PIDB" controller are given in Table 2.

$$L_{FP} \quad L_{FD} \quad L_{TP} \quad L_{TF}$$

| 1000 | 10 | 300 | 5000 |

Table 2

Reference torque $T_{ref}$ is a piecewise constant, switching between 40N/m and 30N/m at 1Hz frequency, $F_{ref} = 0.012Wb^2$.

In Figures 2, 3 the "PIDB" controller (18)-(19) is examined as a part of the sensorless scheme when all motor parameters are known and there is no noise in the stator current measurements. Long transient at the beginning
is caused by the delay in the speed and flux estimation. Figures 4 show the comparison of "PIDB" and "SSB" controllers in the case of the rotor resistance $R_r$ increased by 40% over its nominal value and the stator current contaminated with the white noise ($\pm 3\,\text{A}, 1\,\text{kHz}$). Both sensorless schemes are robust with respect to variations of $R_r$. The "SSB" controller is less sensitive to the noise in the stator current than the "PIDB" controller.

6 Conclusions

- A novel scheme for the direct torque and flux control of IM is designed that is based on the dynamic torque-flux model where the separation of the torque and flux regulation is achieved. The controller provides PI regulation of the torque and PD regulation of the rotor flux and it guarantees the asymptotic stability of the torque and flux tracking.
- The scheme based on the steady-state solution of the torque-flux dynamic model is also designed. It is simpler than the first scheme and more robust to the current noise.
- The analysis of the stator current and voltage constraints defines the allowable set of the torque and flux reference commands.
- The direct torque and flux controller is complimented with a rotor flux observer. The real time implementation issues of the flux observer and the analytical design based on the zero-order-hold (ZOH) or Tustin approximations are presented in [7].
- The sensitivity analysis revealed that the sensorless direct torque and flux regulation schemes are insensitive to the rotor resistance variations. This improved robustness of the control could result in better efficiency of an electric drive as compared to the commonly used IFO scheme.

References


