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Adaptive Observers for Sensorless Control of an Induction Motor

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Abstract

In this paper, novel adaptive speed and flux observers are designed as a part of the speed sensorless control scheme of an induction motor (IM). The design is based on a special form of IM equations with filtered current and voltage command signals as inputs. The speed tuning algorithm incorporates the stator current observer with gradient or sign speed adjustment schemes. The proposed flux observer is based on an algebraic relation involving measured and filtered signals and it shows better performance than the conventional flux observers in the low speed region and during transients. Various simulation and experimental results are presented to demonstrate the potential and practicality of the approach.

1. Introduction

In recent years, there has been a lot of research on design of speed and flux observers for sensorless control of induction motors (IM) [1] - [6].

A high gain speed observer is suggested in [1] with the robust rotor flux estimator being implemented based on the flux natural dynamics. Two speed observers that compliment each other in different modes of operation of IM are analyzed in [2]. Sliding mode speed and flux observers and torque controllers of IM are designed in [3] with conditions of asymptotic stability of control being established. A method for speed and flux estimation using an adaptive control technique is elaborated in [4] while the major result on observer stability is not proven accurately. The speed observer based on introducing of the pseudo current and voltage signals is presented in [5] and it is further elaborated to estimate the rotor resistance in [6].

A variety of rotor flux estimation schemes assuming that the speed signal is available is constructed and categorized in [7]. Flux observers utilizing flux dynamics with the additional posititon/velocity tracking error and the torque and flux tracking error are suggested in [8] and [9], respectively. A novel scheme for the torque and flux control of an induction motor that is based on the direct tracking of the torque and flux signals is elaborated in [10].

The speed and flux observers presented in this paper are based on a special form of the IM model where stator current dynamics are separated from the unobservable rotor flux. The model is obtained by assuming that the speed is a slowly changing variable relative to the flux and current signals. Next using the separated model (SM) for the stator current as a reference the rotor speed identification scheme is designed. The gradient and sign tuning models are considered with the last one having better performance in low-speed region and in the case of the current noise. The speed identifiability condition is analyzed showing its equivalence to the persistance excitation of the electrical frequency.

An alternative speed (AS) observer is elaborated using a higher order tuning system. The AS observer better tracks speed transients although it is more susceptible to a noise.

Using the SM for the stator currents an algebraic equation that defines the rotor flux is obtained. The equation is used to construct a rotor flux observer that does not involve integration.

The paper is organized as follows. Section 2 discusses the dynamic model and control problem statement. The main results of adaptive speed and flux observers design are given in Section 3 with experimental results being presented in Section 4.

2. Dynamic model of an induction motor and control problems

Consider the dynamic model of IM in the stator frame [11]

$$\begin{aligned}
\frac{d\lambda_r}{dt} &= \left(-\frac{R_r}{L_r}I + n_p\omega J\right)\lambda_r + \frac{R_r}{L_r}Mi_s \qquad (1)\\
\frac{di_s}{dt} &= -\frac{M}{\sigma L_s L_r}\left(-\frac{R_r}{L_r}I + n_p\omega J\right)\lambda_r - \\
&\quad -\frac{1}{\sigma L_s}\left(R_s + \frac{M^2 R_r}{L_r^2}\right)i_s + \frac{1}{\sigma L_s}v_s \qquad (2)\\
\frac{d\Theta}{dt} &= \omega \qquad (3)\\
\frac{d\omega}{dt} &= \mu i_s^T J\lambda_r - \frac{\alpha\omega}{m} + \frac{T_L}{m}, \qquad (4)
\end{aligned}$$

where

$$I = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix} \qquad J = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix}$$

λ_r, i_s, v_s — rotor flux, stator current and stator voltage command
Θ – angular position of the rotor
ω — angular speed of the rotor
R_r, R_s — rotor and stator resistance
M — mutual inductance
L_r, L_s — rotor and stator inductance
$\sigma = 1 - M^2/L_sL_r$ — leakage parameter
n_p — number of pole pairs
m — moment of inertia of the rotor
α — damping gain
T_L — external load torque
$\mu = 3n_pM/2L_r$
$T_e = \mu i_s^T J\lambda_r$ — electromagnetic torque.

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

$$\lim_{t\to\infty} T_e \to T_{ref} \tag{5}$$

To achieve the goal (5) the flux magnitude is to be kept at a certain level and the second control goal is

$$\lim_{t\to\infty} \lambda_r \to F_{ref}, \tag{6}$$

where F_{ref} is the flux reference value.

It is assumed that the reference values $T_{ref}, F_{ref} \in C^1[R^+]$ and they should be selected accounting for constraints on the voltage and current signals.

The torque and flux regulation (5), (6) is achieved by using standard indirect field oriented (IFO) control technique [11] when rotor speed is measured. 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 the problem of the estimation of the rotor speed from the available for the measurement stator current and stator voltage command arises

$$\lim_{t\to\infty} \hat{\omega} \to \omega \tag{7}$$

where $\hat{\omega}$ is a rotor speed estimate.

In direct field oriented (DFO) torque and flux regulation schemes the value of the rotor flux or its estimate is used. Thus the next problem is to construct an observer for the rotor flux with the convergence

$$\lim_{t\to\infty} \hat{\lambda}_r \to \lambda_r \tag{8}$$

where $\hat{\lambda}_r$ is a rotor flux estimate.

Since parameters of an induction motor may change during its operation and their exact values may be essential for the quality of control, the problem of online estimation of the motor parameters arises. In particular, rotor and stator resistances are to be estimated

$$\lim_{t\to\infty} \hat{P} \to P \tag{9}$$

where $P = [R_r, R_s]^T$ and $\hat{P}$ is its estimate.

This paper is devoted to the design of estimators for rotor speed and flux only assuming that motor parameters are known or properly estimated using standard technique.

3. Adaptive rotor speed and flux observers

3.1. Rotor speed observer

In order to design a speed observer an assumption is made that the rotor speed changes significantly slower relative to the rotor flux. This assumption allows us to consider the rotor speed as a slowly changing unknown parameter so that the adaptive identification techniques can be applied.

First differentiating (2) and eliminating λ_r gives

$$\begin{aligned}\frac{d^2 i_s}{dt^2} &= (\alpha_1 I + \omega\beta_1 J)\frac{di_s}{dt} + (\alpha_2 I + \omega\beta_2 J)i_s + \\ &\quad +(\alpha_3 I + \omega\beta_3 J)v_s + \alpha_4\frac{dv_s}{dt},\end{aligned} \tag{10}$$

where

$$\alpha_1 = -\frac{1}{\sigma L_s}\left(R_s + \frac{M^2 R_r}{L_r^2}\right) - \frac{R_r}{L_r}; \quad \beta_1 = n_p;$$

$$\alpha_2 = -\frac{R_r R_s}{\sigma L_r L_s}; \quad \beta_2 = \frac{n_p R_s}{\sigma L_s};$$

$$\alpha_3 = \frac{R_r}{\sigma L_s L_R}; \quad \beta_3 = -\frac{n_p}{\sigma L_s}; \quad \alpha_4 = \frac{1}{\sigma L_s}.$$

Adding $c\,di_s/dt$ for some $c > 0$ to both sides of (10) and formal division by $(d/dt + c)$ transform (10) to

$$\frac{di_s}{dt} = a(t) + \omega b(t) + \epsilon. \tag{11}$$

Here $\epsilon \to 0$ exponentially and functions $a(t)$ and $b(t)$ are linear combinations of the filtered stator current and voltage signals $i_{s0}, i_{s1}, v_{s0}, v_{s1}$

$$a(t) = (c + \alpha_1)i_{s1} + \alpha_2 i_{s0} + \alpha_3 v_{s0} + \alpha_4 v_{s1}, \tag{12}$$

$$b(t) = J(\beta_1 i_{s1} + \beta_2 i_{s0} + \beta_3 v_{s0}), \tag{13}$$

where

$$i_{s0} = \frac{1}{s+c} i_s; \quad i_{s1} = \frac{s}{s+c} i_s,$$

$$v_{s0} = \frac{1}{s+c} v_s; \quad v_{s1} = \frac{s}{s+c} v_s,$$

and "s" is a Laplace transform variable.

Equation (11) is used as a reference model for the rotor speed identification. The tuning model (stator current observer) is given by the following formula

$$\frac{d\hat{i}_s}{dt} = a(t) + \hat{\omega} b(t) - L(\hat{i}_s - i_s), \quad L > 0, \tag{14}$$

where $\hat{i}_s$ is the stator current estimate, $\hat{\omega}$ is the rotor speed estimate and L is a positive constant.

Let $e = \hat{i}_s - i_s$ be the stator current estimation error, then the dynamic equation for the error is obtained by subtracting (11) from (14)

$$\frac{de}{dt} = -Le + (\hat{\omega} - \omega)b(t) - \epsilon. \tag{15}$$

The speed estimate $\hat{\omega}$ is to be adjusted in such a way as to make the current estimation error and its derivative tend to zero. This will imply the convergence to zero of the expression $(\hat{\omega} - \omega)b(t)$. If the identifiability condition $|b(t)| \geq \delta > 0$ holds, then the speed estimate will converge to the actual rotor speed. Such an adjustment is realized by the following equation

$$\frac{d\hat{\omega}}{dt} = -\gamma e^T b(t), \tag{16}$$

with some positive tuning gain $\gamma > 0$.

The stator current observer (12)-(14) together with the adjustment mechanism (16) constitute the speed observer. Conditions under which the speed identification occurs are given in the following theorem.

Theorem 1 *Consider a dynamic model of an induction motor (1), (2) and the speed observer (12)-(14), (16). Let the stator voltage command v_s be a bounded continuous piecewise smooth function. Suppose that the identifiability condition*

$$|b(t)| \geq \delta > 0 \tag{17}$$

holds for all $t \geq t_0$ and some $\delta > 0$, $t_0 \geq 0$.
Then the speed identification occurs

$$\lim_{t \to \infty} \hat{\omega}(t) \to \omega.$$

Remark 1 The identifiability condition (17) can be substituted by one of the following equivalent persistent excitation conditions [12]:

PE1 There exist $\alpha > 0, T > 0, t_0 > 0$ such that for all $t \geq t_0$

$$\int_t^{t+T} b(s)b(s)^T ds \geq \alpha I.$$

PE2 There exist $\alpha > 0$, $T > 0$, $t_0 > 0$ such that for all $t \geq t_0$, $\xi \in R^2$ there exist $t_* \in [t, t+T]$ such that $|b(t_*)\xi| \geq \alpha|\xi|$.

PE3 There exist $C > 0$, $T > 0$, $t_0 > 0$ such that for all $t \geq t_0$ there exist $t_i \in [t, t+T]$, $i = 1, 2$, such that $[b(t_1), b(t_2)]^{-1} \leq C$.

Note that the condition **PE1** is the definition of the persistent excitation of a function $b(t)$.

Remark 2 The identifiability condition (17) is equivalent to the following inequality

$$\left| \int_0^t e^{-c\tau} \left(\beta_1 \frac{di_s}{dt} + \beta_2 i_s + \beta_3 v_s \right) d\tau \right| \geq \delta.$$

Since $(\beta_1 di_s/dt + \beta_2 i_s + \beta_3 v_s) = \delta_1 \omega_e$ [5], where ω_e is the electrical excitation frequency and δ_1 is the constant that depends on machine parameters, then the identifiability condition is violated when $\omega_e \equiv 0$, i.e., the electromagnetic field is not rotating.

Remark 3 Speed estimation law (16) can be substituted by the sign-adjustment scheme

$$\frac{d\hat{\omega}}{dt} = -\gamma sign(e^T b(t)), \quad \gamma > 0. \tag{18}$$

Though its stability is not proved analytically, it shows better performance than (16) in the low-speed region and in the case of the noise in the current measurements.

Remark 4 The adaptation algorithms of the gradient type (16) may not be robust with respect to small disturbances. To overcome this problem the robust modifications of (16) with a deadzone or σ-modification are used [13].

Proof of Theorem 1
The proof is based on the Lyapunov function candidate

$$V(e, \hat{\omega}, t) = \frac{1}{2}e^2 + \frac{1}{2\gamma}|\hat{\omega} - \omega|^2 + \int_t^{\infty} \frac{\epsilon^2}{4L} ds.$$

Taking into account the expression (16) its derivative along the solutions of (15) equals to

$$\dot{V}(e, \hat{\omega}, t) = -Le^2 - e^T\epsilon - \frac{\epsilon^2}{4L} = -\left(\sqrt{L}e + \frac{\epsilon}{2\sqrt{L}}\right)^2 \leq 0. \tag{19}$$

The statement of the theorem technically follows from (19).

The proposed speed observer depends on three parameters ($c > 0$, $L > 0$, $\gamma > 0$). Their values may substantially influence the observer performance especially when the speed estimate is used in the closed loop control or the speed changes are fast. In general increasing the parameter γ improves the estimation convergence, but too high values of γ may cause the overshoot and increase the observer sensitivity to the noise. Conversely, very large values of parameters c and L may result in slow convergence. Thus they should be chosen rather small. At the same time if c and L are excessively small, oscillations of the rotor speed estimate with slow attenuation may be observed. There is no any precise rule for tuning these parameters, but mentioned above observations should be taken into account.

There is always some noise in the stator current measurements. Several methods can be proposed to lessen the observer sensitivity to the noise. The first one is to use the adjustment mechanism (18) instead of (16). According to simulation results the sign-scheme (18) shows better robustness to the current noise. Another way is to filter out the rotor speed estimate with a low-pass filter. It is also possible to filter out the inputs (i_s, v_s) using the same filter. One should take into account that parameters of the chosen filter will influence the performance of the observer.

3.2. Alternative rotor speed observer

In this section, the speed observer using the tuning system of the higher order is derived. In part, the observation scheme is similar to the one designed in the previous section because the observer utilizes the same methodology of using filtered signals instead of the stator current derivatives.

Introducing the new state space vector $z = [i_{s0}^T, i_{s1}^T]^T$ rewrites the (10) in the state space form

$$\frac{dz}{dt} = Az + \omega g(t) + f(t), \quad (20)$$

where

$$A = \begin{bmatrix} 0 & I \\ \alpha_2 I & \alpha_1 I \end{bmatrix}, \quad f(t) = \begin{bmatrix} 0 \\ \alpha_3 v_{s0} + \alpha_4 v_{s1} \end{bmatrix},$$

$$g(t) = \begin{bmatrix} 0 & 0 \\ \beta_2 J & \beta_1 J \end{bmatrix} z + \begin{bmatrix} 0 \\ \beta_3 J v_{s0} \end{bmatrix}.$$

Since $z(t)$, $f(t)$, $g(t)$ are functions of signals i_{s0}, i_{s1}, v_{s0} and v_{s1}, then they may be computed from the measured current and voltage command.

Considering the equation (20) as a reference model the tuning system will be

$$\frac{d\hat{z}}{dt} = A\hat{z} + \hat{\omega} g(t) + f(t) - C(\hat{z} - z), \quad (21)$$

where $\hat{z}$ is an estimate of z, $\hat{\omega}$ is an estimate of the speed ω and C is 4×4 matrix such that $\mathcal{A} = A - C$ is Hurwitz.

The equation for the error $e_z = \hat{z} - z$ is

$$\frac{de_z}{dt} = \mathcal{A} e_z + (\hat{\omega} - \omega) g(t). \quad (22)$$

Let $H = H^T > 0$ be a positive definite matrix such that $H\mathcal{A} + \mathcal{A}^T H < 0$ be negative definite. Such H exists since $\mathcal{A}$ is Hurwitz. Then the speed estimation mechanism is designed according to the following equation

$$\frac{d\hat{\omega}}{dt} = -\gamma e_z^T H g(t), \quad \gamma > 0. \quad (23)$$

The tuning model (21) and the adjustment equation (23) constitute the speed observer. The convegence conditions for the speed observer (21), (23) are stated in the following theorem.

Theorem 2 *Consider a dynamic model of an induction motor (1), (2) and the speed observer (21), (23). Let the stator voltage command v_s be a bounded continuous piecewise smooth function. Suppose that the identifiability condition*

$$|g(t)| \geq \delta > 0 \quad (24)$$

holds for all $t \geq t_0$ and some $\delta > 0$, $t_0 \geq 0$.
Then the speed identification occurs

$$\lim_{t\to\infty} \hat{\omega}(t) \to \omega.$$

All remarks to Theorem 1 and regarding the methods for reduction of the sensitivity to the noise mentioned in Sec.3.1 are also valid for the speed observer (21), (23).

The proof of Theorem 2 is based on the Lyapunov function candidate

$$V(e_z, \hat{\omega}) = \frac{1}{2} e_z^T H e_z + \frac{1}{2\gamma} |\hat{\omega} - \omega|^2.$$

The speed observer (21), (23) utilizes the reference model of the higher order and thus more information about the IM dynamics is used in the rotor speed adjustment. The choice of matrices C and H may influence the dynamics of the system. In some cases the correct choice of these matrices may significantly simplify the whole observer.

3.3. Rotor flux observer

Transformation of IM equations in Section 3.1 can be used to design a new rotor flux observer. The conventional rotor flux observer utilizes natural dynamics of the IM and involves integration of the flux dynamic equation (1).

Using IM model (11) the integration in the estimation of the rotor flux can be avoided. Having the rotor speed estimated the equation (11) provides an estimate of the derivative of the stator current. By substituting this estimate into the equation (1) the algebraic equation for the rotor flux is defined.

Combining equations (11) and (2) and moving all the terms except the term with the rotor flux to the right yield the expression

$$\frac{M}{\sigma L_s L_r}(-\frac{R_r}{L_r} I + n_p \omega J)\lambda_r = -a(t) - \omega b(t) - \epsilon - \\ -\frac{1}{\sigma L_s}(R_s + \frac{M^2 R_r}{L_r^2}) i_s + \frac{1}{\sigma L_s} v_s, \quad (25)$$

where $a(t)$ and $b(t)$ are defined in (12) and (13) and ϵ tends to zero exponentially.

Neglecting ϵ and resolving (25) with respect to the rotor flux the following observer is obtained

$$\hat{\lambda}_r = \frac{\sigma L_s L_r}{M(R_r^2/L_r^2 + n_p^2 \hat{\omega}^2)} \left(-\frac{R_r}{L_r} I - n_p \hat{\omega} J \right) \cdot \\ \left[-a(t) - \hat{\omega} b(t) - \frac{1}{\sigma L_s}(R_s + \frac{M^2 R_r}{L_r^2}) i_s + \frac{1}{\sigma L_s} v_s \right]. \quad (26)$$

If the only information needed from the rotor flux is its orientation, then it is not necessary to calculate the first multiplier in the equation (26).

4. Experimental results

The proposed speed observer has been utilized for the speed sensorless active engine damping. The block diagram of the experimental setup is given in Fig. 1. It includes an indirect injected (IDI) 1.8L Puma diesel engine, mounted on a cart, an 8 kW integrated starter generator (ISG), a clutch, and a water brake to model an external load. This configuration is representative of a hybrid electric vehicle powertrain in which the ISG is used to provide a start-stop operation, a power boost, and also to replace a standard passive flywheel function.

The DSP based controller **ACE** is used to control the inverter and communicate with the engine electronic controller (EEC), sensors and an external laptop computer. The controller development and C code generation are performed in the Xmath/SystemBuild graphical environment in the laptop computer. To make the computation process more efficient all processes are divided into two classes: slow and fast. The slow processes are implemented in the outer loop with the sampling frequency of 1 kHz and the fast ones are run with the frequency of 10 kHz in the inner loop. The fast processes include real-time implementation of the indirect field-oriented (IFO) control of the IM and updates of current and speed signals. The speed observer is implemented in the outer loop with frequency 1 kHz. The observer parameters are selected to be $\gamma = 2000$, $L = 1000$, $c = 100$ and they can be tuned in real time during ISG operation using a communication line between the laptop and **ACE**. For more details on the active damping control design and implementation the reader is referred to [14].

In the experiment shown in Figures 2-4 the proportional controller is used for the active speed damping. Figure 2 shows that the speed estimate tracks well the high frequency (around 30 Hz) engine speed pulsation . In Figure 3 the reference value for i_d component of the stator current equals to $120A$, and the i_q component is determined according to the reference torque generated by the controller.

Figure 4 shows the angular position of the rotor, where θ_{sensor} is the rotor angle obtained from the sensor, θ_{hat} is the estimate of the angle and θ_{ctr} is the value used in IFO control. The difference between θ_{hat} and θ_{ctr} is caused by the delay in the data acquisition system. The error between θ_{sensor} and θ_{ctr} is constant that means that the flux converges to the reference value and thus the field orientation occurs.

5. Conclusions

• The main results of this paper is the design of the adaptive speed and rotor flux observers of IM. The design is based on the special form of the IM model where the stator current and rotor flux are separated. The proposed speed observers are simple and, according to the simulation and experimental results, effective. They work better at high speeds and have a singular mode when the electromagnetic field is not rotating.

• A novel flux observer is described that is based on the algebraic relation between the measured and filtered voltage and current signals.As compared with the conventional rotor flux observers the one proposed in the paper does not integrate the rotor speed estimate and thus works better in the low speed region, where accurate speed estimates are not possible.

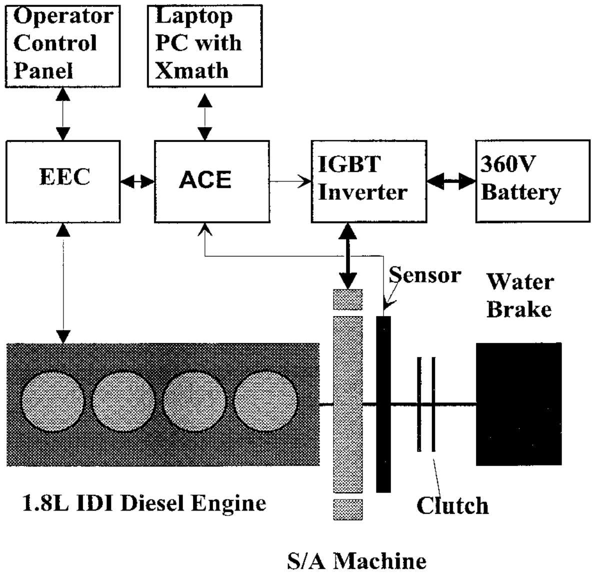

Figure 1: Experimental setup.

• The sensitivity analysis reveals that the sensorless controller utilizing the proposed speed observer shows increased robustness compared to IFO scheme with respect to variations of the rotor resistance. That may result in increased efficiency of an electric drive under changing thermal operation conditions.

• The experimental results demonstrate the practicality of the proposed approach with the example of active engine damping at idle using ISG. The speed estimate tracks well the speed signal with pulsation around $30Hz$ and field orientation holds. Further experiments are planned to investigate the sensitivity of the method to motor parameter variations and to compensate for the voltage signal errors and inverter dead-zone time.

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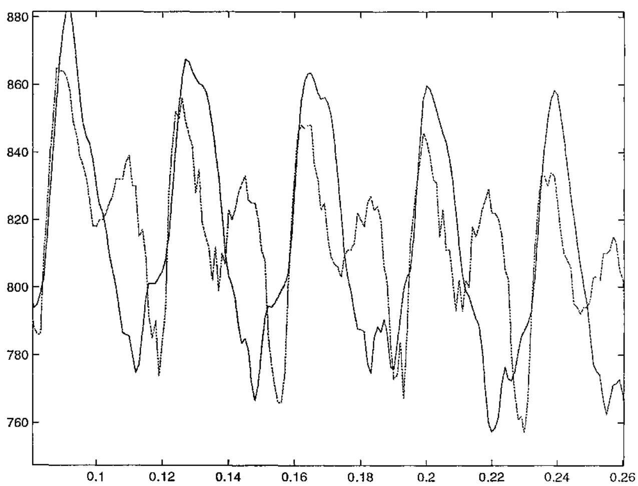

Figure 2: Rotor speed (red) and its estimate (blue), enlarged. Experiment #1.

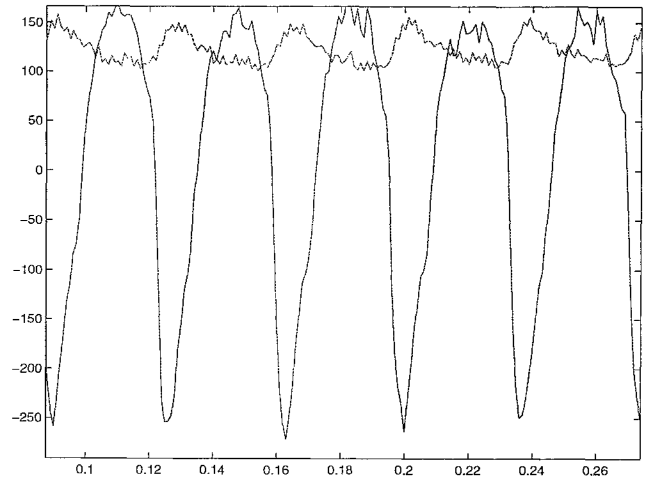

Figure 3: Stator currents i_d (red) and i_q (blue).

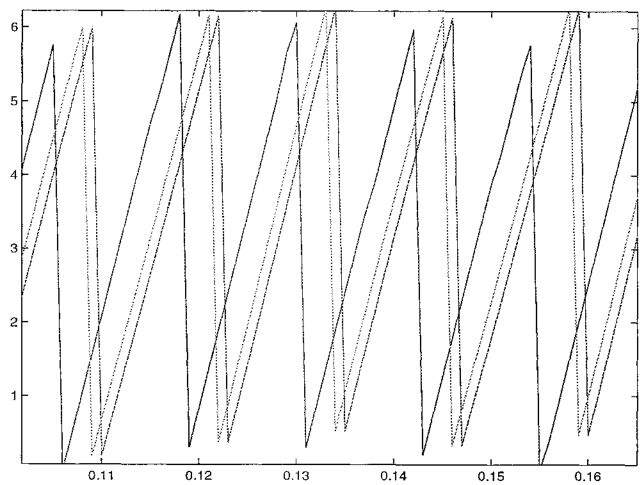

Figure 4: Rotor position angles: θ_{sensor} (blue), θ_{hat} (green), θ_{ctr} (red). Experiment #1.