Position and Speed Sensorless Control of PMSM Drives Based on Adaptive Observer

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Abstract
This paper develops a model reference adaptive system (MRAS) based observer for sensorless vector control of permanent magnet synchronous motor (PMSM) drives. There are used two models in MRAS observer to estimate the linkage flux with relatively clean waveforms. The voltage model in stator reference is the reference model, and the current model in estimated rotor reference is the adaptive model with the rotor position as an adjustable parameter. An adaptive compensation filter depending on speed improves the stability of the pure integration voltage model. The nonlinear adaptation mechanism extracts the error of the flux linkage position and estimates the rotor speed and position. The observer structure, implementation aspects, simulation and experimental results are presented and discussed.

1. Introduction
The main advantages of PMSM are high torque density, high efficiency and small size [1], [2]. In servo drive applications, to control PMSM with sinusoidal flux distribution is requested a precise knowledge of the rotor position to synchronize stator currents with this position. Generally, sensors for mechanical signals as position and speed have the following drawbacks: increase the size, weight and cost; are sensitive to disturbance and temperature variations; and may lead to malfunction. Therefore, there is a high interest in sensorless control structures, i.e., without mechanical sensors, using observers [3].

MRAS based observer approaches which use the voltage model and the current model for sensorless control of induction motor drives are presented in recent papers [3-8]. One model is a reference model and the other is an adaptive model depending on an adaptive parameter. The main idea is that the outputs of these two estimators are compared in an adaptation mechanism to estimate the adaptive parameter that tunes the adaptive model in order to drive to zero the output error between these models. The difficulties that affect the stability are given by the noise in current measurements - especially at low speed no load, and the filter phase lags. Referring to MRAS based observers for sensorless control of PMSM drives, they are less discussed; a BLDC sensorless control using a disturbance observer and an adaptive velocity estimator is presented in [9]. The main goal of the present paper is to develop a MRAS based observer to estimate the flux, speed and position for sensorless vector control of PMSM drives using only the stator currents and the inverter dc-link voltage measurements.

The starting points of this paper are: i) the flux observer in wide speed range [10] based on the parallel connection of the voltage model and the current model, with rotor position measurement, having a flux error compensator in stator reference depending on speed to improve the stability of the voltage model pure integrator; and ii) the ideas on MRAS based observer approaches for induction motor. The paper structure is: MRAS based observer for PMSM - principles, flux observer, MRAS adaptation mechanism, implementation aspects, simulation and experimental results.
2. MRAS Based Observer

2.1. Principle of the MRAS Based Observer for PMSM

In order to estimate the flux linkage \( \lambda \), the electrical rotor speed \( \omega \), and the electrical rotor position \( \theta \) for PMSM drives, a MRAS based observer is used having the principle topology presented in fig. 1. Two estimated flux linkage vectors \( \lambda_v^\wedge \) and \( \lambda_c^\wedge \) are obtained from two models. The voltage model \( \text{Eu}^\wedge \) (1) in stator reference is the reference model with the inputs the stator voltage vector \( u \), and the stator current vector \( i \). It has a pure integrator character and it depends on the estimation of the stator phase resistance \( R \), and on the initial condition \( \lambda_v^0 \). The current model \( \text{Ei}^\wedge \) (2) in rotor reference is the adaptive model with the input the stator current vector \( i \), and with an adjustable parameter, i.e., the estimation of the electrical rotor position \( \theta^\wedge \) used in rotator operators. This model has not dynamic but it is strongly dependent on magnetic parameters: \( \lambda_{o0} \) - permanent magnet flux; \( L_{do}, L_{qo} \) - d, q axis inductance. An adaptation mechanism, with the inputs the flux \( \lambda_v^\wedge \) and \( \lambda_c^\wedge \) which have relatively clean waveforms, extracts the flux position error and it forces this error to zero using the rotor position estimation \( \theta^\wedge \) as an execution variable.

\[
\text{Eu}^\wedge: \quad \hat{\lambda}_v = u - Ri , \quad \lambda_v(0) = \lambda_v^0
\]

\[
\text{Ei}^\wedge: \quad \hat{\lambda}_c = (\lambda_{o0} + L_{do} i_d + j L_{qo} i_q) e^{j\theta}, \quad \lambda_d + j \lambda_q = (i_d + j i_q) e^{j\theta}
\]

Generally, the subscript “o” means estimated parameters, and the superscript “^\wedge” means estimated variables.

Fig. 1 : MRAS based observer - principle topology

There are two main problems to solve:

i) correction of the \( \text{Eu}^\wedge \) pure integrator in the flux observer;

ii) adaptation mechanism structure and parameter design.

2.2. Adaptive Flux Observer

A flux observer for PMSM in wide speed range using measured electrical rotor position \( \theta \) is proposed in [10]. This observer based on the ideas from [7] applied for induction motors, combines the advantages of the current model \( \text{Ei}^\wedge \) at low speed, and the voltage model \( \text{Eu}^\wedge \) at medium and high speed. The flux estimation \( \lambda_c^\wedge \) is obtained as the output of the \( \text{Ei}^\wedge \) estimator that works in an estimated rotor reference \( dq^\wedge(\theta^\wedge) \) with rotor position estimation \( \theta^\wedge \). To correct the offset error in \( \text{Eu}^\wedge \) pure integrator and to realize a smooth frequency transition between the two models depending on speed, the observer \( O \) (3) with a proportional and integral (PI) compensator \( K \) (4) which generates the correction \( u_c \) on the flux error vector \( \Delta \lambda \) was proposed.

\[
O: \quad \hat{\lambda}_v = u - R \dot{i} + u_c , \quad \dot{\lambda}_v(0) = \hat{\lambda}_v^0
\]

\[
K: \quad u_c = K_p \Delta \lambda + K_i \int \Delta \lambda \, dt, \quad \Delta \lambda = \hat{\lambda}_c - \hat{\lambda}_v
\]

\[
k_p = \omega_1 + \omega_2 , \quad k_i = \omega_1 \omega_2 , \quad \omega_2 = 3..10 \omega_1 , \quad \omega_2 = |\hat{\omega}_m|/2
\]
The observer design determines the parameters $k_p$ and $k_i$ of the diagonal matrix $K_p$ and $K_i$ (5) function on the desired observer bandwidth $\omega_1, \omega_2$ [10]. In the actual proposed MRAS based observer an adaptive bandwidth is used with $\omega_2 = |\omega_m|^2$. This condition was determined from practical experimental tests in the idea that, inclusive at low speed, the flux observer output $\lambda_v^\wedge$ weightly contains the voltage model $E\mathbf{u}$ information, i.e., the reference model. Indeed, in this case the speed $\omega$ is nearly of the observer bandwidth and therefore the flux observer output $\lambda_v^\wedge$ depends on both models. The model speed $\omega_m^\wedge$ is estimated from a reduced order model, first order lag (PT1), of the closed loop system that has as input the reference electrical speed $\omega^\ast$. This solution leads to a good observer global stability.

The stator voltage vector $\mathbf{u}(u_a, u_b)$ is computed from the block $S/u$ (6) instead of the stator voltage measurements. It uses only the measured inverter dc-link voltage $V_{dc}$ and the switching binary function $S_{abc}$. $S_{abc}$, $S_a$, $S_b$, $S_c \in \{0,1\}$ which is the final output to direct control of the voltage source inverter (VSI).

$$S/u: \quad \mathbf{u} = \frac{1}{3} V_{dc} (2 S_a - S_b - S_c) + j \frac{1}{\sqrt{3}} V_{dc} (S_b - S_c) \quad (6)$$

In summary, the vector structure of the proposed adaptive flux observer is given in fig. 2.

### 2.3. MRAS Adaptation Mechanism

Both $E\mathbf{i}_r$ and $E\mathbf{u}$ estimators give same information - the flux linkage vector $\lambda$. This is a unique vector in PMSM, but computed in two ways: $\lambda_v^\wedge$ and $\lambda_c^\wedge$ that have relatively clean waveforms. The main problem is to find a procedure to obtain the flux position error $\Delta \theta_\lambda$ between the two flux vector estimations. This error $\Delta \theta_\lambda$ will be used as corrective information to obtain the adaptive parameter $\hat{\theta}$ in the adaptive model $E\mathbf{i}_r$ in order to drive to zero the error $\Delta \theta_\lambda$. In other words, the phase of $\lambda_c^\wedge$ vector will be forced to converge to the phase of $\lambda_v^\wedge$ vector like a phase locked loop (PLL). The flux vector modules make no difference. In this context, let examine the trigonometric and algebraic vector expressions of $\lambda_v^\wedge$ and $\lambda_c^\wedge$, in $\alpha\beta$ stator reference given by (7). From the imaginary part expressions of the scalar product between the vector $\lambda_v^\wedge$ and the conjugate vector $\lambda_c^\wedge$ in (8) it is calculated the flux position error $\Delta \theta_\lambda$ (9). Note that the error $\Delta \theta_\lambda$ does not depend on speed; that is a remarkable result.

$$\hat{\lambda}_v = \lambda_v e^{j\theta_v} = \lambda_v^{\alpha\alpha} + j \lambda_v^{\alpha\beta}, \quad \hat{\lambda}_c = \lambda_c e^{j\theta_c} = \lambda_c^{\alpha\alpha} + j \lambda_c^{\alpha\beta} \quad (7)$$

$$\text{Im}(\hat{\lambda}_c^* \hat{\lambda}_c) = \lambda_v \lambda_c \sin \Delta \theta_\lambda = \lambda_v^{\alpha\beta} \lambda_c^{\alpha\alpha} - \lambda_v^{\alpha\alpha} \lambda_c^{\alpha\beta}, \quad \Delta \theta_\lambda = \theta_v - \theta_c \quad (8)$$

$$\Delta \theta_\lambda = \sin \Delta \theta_\lambda \equiv ( \lambda_v^{\alpha\beta} \lambda_c^{\alpha\alpha} - \lambda_v^{\alpha\alpha} \lambda_c^{\alpha\beta} ) / \lambda_{bo}^2 \quad (9)$$

A Lunberger type observer (10) estimates the rotor speed $\omega^\wedge$ and position $\theta^\wedge$ using as corrective term the flux position error $\Delta \theta_\lambda$. This observer shows like a PI compensator ($k_0, k_\omega$).

$$\begin{bmatrix} \dot{\theta}^\wedge \\ \dot{\omega} \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} \theta^\wedge \\ \omega \end{bmatrix} + \begin{bmatrix} k_\theta \\ k_\omega \end{bmatrix} \Delta \theta_\lambda^{\wedge}, \quad \dot{\omega}_1 = \dot{\theta}, \quad \begin{bmatrix} \hat{\theta}(0) \\ \hat{\omega}(0) \end{bmatrix} = \begin{bmatrix} \hat{\theta}_0 \\ \hat{\omega}_0 \end{bmatrix} \quad (10)$$
The structure of the MRAS adaptation mechanism is given in fig. 3. There are two possibilities for the speed estimation: \( \omega^f \) and \( \omega^e \) [11]. The \( \omega^e \) estimation is noiseless as the output from an integrator. The \( \omega^e \) estimation taken after the summation contains a term proportional with the error \( \Delta \theta_h \) and thus it is affected by noise. On the other hand, \( \omega^f \) has more accurate dynamic estimation.

In the steady state the flux linkage vector \( \lambda \) is synchronized to the rotor, and thus the rotor speed \( \omega^e \) converge to the real speed \( \omega \). However, in transient regimes when the torque demand is changed, the speed estimation will be affected because the flux linkage vector moves faster relative to the rotor to produce the new torque level [12]. In other words, the cause is that the electrical time constant that determines the electromagnetic torque dynamic is usually at least 10 times then the mechanical time constant which determines the speed dynamic. Thus possible oscillations may be occurred. A solution to attenuate this drawback is to use for the speed controller a filtered estimated speed \( \omega^e \) obtained from \( \omega^e \) by filtering through a PT1 \( (T_i) \) element. Note that the reverse speed procedure does not require special design structure for the speed sign computation, especially at the zero crossing speed.

2.4. Parameter Design of the MRAS Adaptation Mechanism

Taking into account the delay \( h \) in the sampling procedure, in fig. 4 is presented the equivalent control structure of the MRAS adaptation mechanism based on the implementation structure from fig. 3. The problem is reduced to design a PI controller \( (k_p, k_i) \) in correlation with desired criteria. The equivalent transfer function \( H_L(s) \) of the process fixed part in fig. 4 is:

\[
H_L(s) = \frac{A}{s(1+h s)} = \frac{1}{1 + hs}, \quad A = \lambda^2_{\omega_0} \tag{11}
\]

A design procedure that leads to good results is a generalization of the symmetric optimum criterion after Kessler [1], [13]. According to this procedure, the parameters \( k_p = k_o, k_i = k_o \) of the PI compensator \( H_c \) associated to the process \( H_L(s) \) (11) are selected to respect the following condition. The Bode plots (the frequency transfer functions plots for the magnitude and phase) of the open-loop transfer function \( H_O = H_L \ast H_c \) are placed symmetrically to the crossover frequency \( \omega_f \) that is the geometric mean of the two \( H_O \) characteristic frequencies. In this case, the design procedure uses a single parameter \( m \) that is selected with respect one of the following requirements: the phase reserve, the frequency bandwidth, or the damping factor, i.e., \( \xi = (m-1)/2 \). For every adopted value of the parameter \( m \), the resulted tuning parameters ensure the maximum possible system phase reserve. The crossover frequency \( \omega_c \) and the PI controller parameters \( k_o, k_i \) are given by the expressions:

\[
k_p = \frac{1}{m h A}, \quad k_i = \frac{1}{m^2 h^2 A} \tag{12}
\]

In the particular case of the application from fig. 3, there are: \( \omega_{\text{max}} = 400 \text{ rad/s}, \lambda_0 = 0,092 \text{ Wb} \) and it selects \( h = 0,2 \text{ ms} \). In order to obtain a small overshoot it is chosen \( \xi = 2 \) thus \( m = 5 \). According to the relations (12) it results: \( \omega_0 = 1000 \text{ rad/s}, k_p = 120 \text{ 000}, k_i = 53 \text{ 000} \). The time constant of the PI compensator is \( T_i = k_p / k_i = m^2 h \), i.e., \( T_i = 5 \text{ ms} \). For the PT1 filter is chosen \( T_f > 10 \times h \), i.e., \( T_f = 15 \text{ ms} \).

![Fig. 3: Structure of the MRAS adaptation mechanism](image-url)

![Fig. 4: Equivalent structure of the MRAS adaptation mechanism](image-url)
2.5. Sensorless Control Structure

An example for a sensorless vector control structure of PMSM drives using the proposed MRAS based observer to estimate the speed and position is presented in fig. 5. This structure contains the following main modules: \( R_0 \) - speed controller anti-windup PI type \((k_{p o}, T_{i o}, k_{s o}, I_{l o})\), with a PT1 filter \((T_{f o})\) on the speed reference; \( \theta^o \)-simplified rotation operator \((i_{d o} = 0)\) with the estimated rotor position orientation \( \theta^o \); phase current controllers bi-position hysteresis type; MRAS-Obs - proposed MRAS based observer (fig. 2 and fig. 3); VSI - voltage source inverter, and PMSM.

![Sensorless control structure of PMSM drives using MRAS observer](image)

Fig. 5 : Sensorless control structure of PMSM drives using MRAS observer

3. Simulation Results

The case study considers a PMSM with ferrite, 130-SFP-2.3 type made by Electrotehnica Bucharest, with the rated parameters: \( T_{o} = 2.3 \text{ Nm}, L_{in} = 3 \text{ A}, V_{d c} = 100 \text{ V}, \Omega_o = 100 \text{ rad/s}, p = 4, \lambda_{o} = 0.092 \text{ Wb}, L_{o} = 0.02 \text{ H}, L_{d o} = 0.012 \text{ H}, R_o = 1.8 \text{ Ohm}, J_o = 0.005 \text{ kgm}^2, B_o = 0.001 \text{ Nm/rad/s}, \text{finv} = 5 \text{ kHz}.

The speed controller anti-windup PI type, is designed by respect the symmetric optimum criterion after Kessler, with filtered speed reference. Its parameters are: \( k_{p o} = 0.5, T_{i o} = 60 \text{ ms}, T_{f o} = 90 \text{ ms}, k_{s o} = 10, L_{in} = 7 A \). The PI compensator parameters \( k_p, k_i \) of the flux observer from fig. 2 have: \( \omega_o = |\omega_{0q}|/2, \omega_q = |\omega_{0q}|/15 \). The speed \( \omega_{0q} \) of the PT1 reduced model is limited to 2 rad/s, and the time constant \( T_m = 150 \text{ ms}, \) for a settling time \( t_s = 200 \text{ ms} \). The adaptation mechanism parameters, from fig. 3, are designed in the paragraph 2.4, and they are: \( k_b = 120 \text{ 000} \) and \( k_{io} = 53 \text{ 000} \) \((T_{io} = 5 \text{ ms})\); \( T_f = 15 \text{ ms} \).

Digital simulations and experimental tests present transient and steady state performances, and they prove the robustness of the proposed sensorless drive system with MRAS based observer - especially the \( \Delta \theta \) convergence, for low and high speed, with large transient electromagnetic torque, and with real parameter variations. Considering a temperature variation of 30°C, the phase resistance \( R \) increases with 10%, and the permanent magnet flux \( \lambda_0 \) decreases with 5%. The saturation phenomena decrease the inductance \( L_{q} \), but the dependence \( L_{q}(i_q) \) can be determined experimentally and implemented in the control system [2]. From these remarks, the PMSM parameter variations taking into account in the detuned case are: \( R = \{1.1, 0.9\} R_o \), and \( \lambda = \{0.95, 1.05\} \lambda_o \).

The digital simulation uses Matlab-Simulink package. The numerical integration method is a simplest one, i.e., Euler method, with 200 \( \mu \text{s} \) sampling rate, in correspondence with the inverter switching frequency. In tuned and detuned parameter cases, the severe step responses in transient regimes are: start up, reverse speed and loaded torque, both at low speed and high speed, to prove the theoretical considerations - especially the \( \Delta \theta \) convergence. The simulation time is 0 to 1.5 s when the following step inputs are applied: at \( t_0 = 0 \text{ s}, \omega_0^* = \{-2, 100\} \text{ rad/s}; \) at \( t_1 = 0.4 \text{ s}, \omega_0^* = -\omega_0^*; \) at \( t_2 = 1 \text{ s}, T_0^* = \{0.5, 2\} \text{ Nm} \). Figs. 6 and 7 show transient regimes in the first detuned case: \( \lambda_o = 0.95 \lambda_o, R = 1.1 R_o \). The interested transient responses are: real mechanical speed \( \omega \), estimated mechanical speed \( \omega^\wedge \), position error \( \Delta \theta = \theta - \theta^o \), electromagnetic torque \( T_r \), and estimated flux vector component \( \lambda_r \) \((\lambda_{rmax}, \lambda_{rmin})\).

The speed responses are fast, without steady state errors, torque responses are also fast and \( \Delta \theta \) error converges to zero in every case. Figs. 8a and 8b show the \( \Delta \theta \) error transient in the second detuned case: \( \lambda_o = 1.05 \lambda_o, R = 0.9 R_o \), and in the ideal tuning case. The worst case, when \( \Delta \theta \) is larger, is the detuned case at reverse high speed. At low speed, \( \Delta \theta \) error is not so large, but the convergence time is longer.
In conclusion, the simulation results prove the theoretical assumptions and the robustness of sensorless system with the proposed MRAS observer to real parameter variations, in 2-100 rad/s speed range.
4. Experimental Results

4.1 Hardware Structure

The hardware configuration set-up of the experimental sensorless control system is presented in fig. 9.

The process coupler board ADA-1100 has the following resources that are used in this application: analog to digital converter (ADC), programmable timer, interrupts controller, and programmable I/O ports. Five multiplexed sample and hold 12 bit ADC channels, 24 µsec conversion time per channel are used for acquisition. The acquisition signals are: $i_a$, $i_b$ - phase currents and $V_{dc}$ - dc-link voltage, for control; $\sin \theta$, $\cos \theta$ from the resolver position transducer (PT) interface in order to calculate $\theta$ and $\omega$, in the control system using a PLL based observer, to monitoring and to compare these signals with the estimated one. There are two interrupt requests to the interrupt controller I8259A, i.e., the ADC end of conversion (EOC), and from the programmable timer I82C54 one channel that gives the sample rate $h = 200$ µsec. The output logic signals are 3 bits: $S_a$, $S_b$, $S_c$ to direct control the inverter.

The adapter interface has the goal to amplify and to filter the analog input signals, and to adapt the level of the digital output signals. Anti-aliasing analog low pass filters - first order lag type, with time constant equal with inverter switching period $h$ are used.

The voltage source inverter three phase bridge with bipolar transistors from one axis module VAMS-m type made by Electrotehnica Bucharest is used. From technical reasons $V_{dc} = 50V$. The maxim switching frequency of the inverter PWM method is 5 kHz. Two stator phase currents ($i_a$, $i_b$) are taken from sensors with galvanic separation. For monitoring only, a position transducer resolver type is used.

The program for the process coupler board uses small memory model and it has two variables that can be accessed by the C programs: a pointer to an acquisition string that contains acquisition data; a flag that indicates outrunning of sample rate (OSR). There are resource initializations for I/O ports, timer, ADC; and subroutines to install/uninstall interrupts from EOC_ADC and timer. The interrupt service routine for ADC reads the conversion results, writes this in the acquisition string, selects the next channel, starts conversion while the channel counter < 5. The interrupt service routine for timer programs the timer, selects first channel, starts conversion and sets OSR flag.

The program for digital control algorithms, written in C language, is compiled for execution on PC to optimize the speed computation. All functions of the real time control system are implemented running in 200 µsec, including also the digital filters for acquisition signals and a PLL observer to estimate the $\theta$ and $\omega$ from the $\sin \theta$, $\cos \theta$ resolver signals. There is a software watch dog timer that analyzes the OSR flag at the end of the computation. Digital recursive equations of these algorithms are obtained using usual numerical integration methods, e.g., rectangle rule and/or trapezoidal rule (Tustin), but a specific
method in not very important because the sample rate is higher comparative with the small electrical time constant. However, when measured signals are used in digital algorithms, the delay rectangle rule is preferred because the nature of the sampling procedure himself introduces a sampling period delay. The compensation for the inverter dead time was determined to be necessary for further accuracy improvements, especially at low speeds. A ring buffer list, with programmable size memorizes, loads at each sampling time three desired variables that are used later at the end of real time running.

Graphical display interface, written in C, off-line working, extracts the three variables from the ring list. There is the possibility to display the selected variables on two diagrams: upper diagram for first variable and down diagram for second or third variables or both in time. In other regime can be obtain a state plot diagram of two variables. There are facilities to modified the display scales and to obtain a zoom. The acquisition windows can be one shot at the start run command, or the last window before the stop command. This virtual instrument, friendly user, is very useful in experiments to integrate the control system, to tune algorithms, to show permanent and transient regimes in real time applications.

4.3. Experimental Test Results

The experimental tests present transient and steady state performances of the proposed sensorless drive system with MRAS based observer in severe conditions, i.e., reverse step speed and step speed no load regimes. Extensive experimental tests have been done, but only representative results are presented. Fig. 10 shows transient diagrams in the reverse step speed regime. For $\omega^* = \pm 25 \text{ rad/s}$, the speed estimation $\omega^\wedge$ is very good, with fast transient response and small overshoot, without steady state error.

![Fig. 10: Transients in reverse step speed regime: the estimated mechanical speed $\omega^\wedge$ and position error $\Delta \theta$ for $\omega^* = \pm 25 \text{ rad/s}$ and $\pm 10 \text{ rad/s}$.]
There are short instability trends only in the zero speed crossing region because the delay in estimation correlated with the fast reverse speed requirements. The position error $\Delta \theta = \theta - \theta^*$ has acceptable value, i.e., $\pm 0.3 \text{ rad}$ in transient, and $\pm 0.1 \text{ rad}$ in steady state. For $\omega^* = 10 \text{ rad/s}$, the $\omega^*$ transient around zero speed crossing is much better with a smooth transition, but the speed ripple rises.

Fig. 11 presents the transient responses for the rotor position from the resolver transducer $\cos \theta_r$, in comparison with the estimated position $\cos \theta^*$, and the flux $\lambda_{\psi \alpha^*}$, $\lambda_{\xi \alpha^*}$ in transient and steady state regimes for a $\omega^* = 50 \text{ rad/s}$ step speed. There are confirmed the correct function of the flux observer with MRAS position based observer. The experiments prove that the $\text{Eu}^*$ offset presented in startup region is driving to zero in 1-2 electrical rotations by the PI compensator of the flux observer. The position $\theta^*$ and flux ripples are better at high speed when the error in $\text{Eu}^*$ integrator became negligible. The instability trends below 10 $\text{rad/s}$ may be explains by the current ripple and inaccurate inverter voltage control. At low speed, the stator voltage becomes lower and the effect of voltage drops across switching devices and power feeders, the temperature variation of stator resistance, and the inverter dead time become significant, and it affects especially the $\text{Eu}^*$ estimator.

The experimental test results prove the theoretical assumptions on MRAS based observer for sensorless control of PMSM drives in 60-1000 rpm range with a good asymptotic stability and good transient and steady state responses.
5. Conclusions

The main conclusions and contributions of this paper are the following:

1. The proposed structure of MRAS based observer to estimate $\lambda^e$, $\omega^e$, $\theta^e$ in sensorless vector control of PMSM drives exploits the relatively clean flux waveforms and consists on two main parts:
   i) A flux linkage adaptive observer that combines the voltage model $E_u^c$ in stator reference - as reference model, and the current model $E_i^c$ in estimated rotor reference - as adaptive model with the rotor position as adjustable parameter. An adaptive compensation filter PI type, depending on speed, improves the stability of the pure integration voltage model. The compensator bandwidth is speed adaptive obtained from a reduce order model of the closed loop system having $\omega^e$ as input;
   ii) A nonlinear adaptation mechanism that extracts the flux linkage position error $\Delta \theta_\lambda = \theta_\lambda^e - \theta_\lambda^a$, i.e., no speed dependent, and it estimates the rotor speed and position to drive $\Delta \theta_1$ faster to zero.

2. The reverse speed procedure does not require special design problems like speed sign computation near zero speed comparatively with other sensorless observers.

3. Extensive simulation results with the Euler method and a relative no fast sampling rate 0.2 ms, show very well performances of the proposed sensorless control system with MRAS based observer in ±20-1000 rpm speed range.

4. The severe experimental tests prove the feasibility of the proposed sensorless control system in ±60-1000 rpm speed range, with a good asymptotic stability and good transient and steady state responses, including reverse step speed regime.

References