Torque estimation of permanent magnet synchronous machine using improved voltage model flux estimator

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Abstract
A torque estimator for permanent magnet synchronous machines is presented based on flux estimation. The voltage model of the machine (Faraday's law) is employed to estimate the flux where pure integrators are applied to the electromotive force (EMF) of the machine. To solve the problem of DC drift due to pure integrators, a current-model flux estimator modified by a Kalman-like observer is temporarily switched on to reinitialize the pure integrators. The current-model estimator with its associated observer is activated only if the difference between the reference and estimated torque exceeds a predefined threshold. For the remainder of the operating time, only pure integrators are used to estimate flux. Furthermore, an online EMF offset detector is designed to compensate EMF offsets before integral action to avoid divergence of the estimated flux. The proposed approach allows smooth estimation at standstill compared with existing approaches that use filters. Moreover, the method allows space to be freed up at the central processing unit because it is not always highly occupied during motor operating times. The simulation and experimental results show the capabilities of the proposed torque/flux estimator.

1 INTRODUCTION

Permanent magnet synchronous machines (PMSMs) are popular in a wide range of industrial applications because of their robustness and lower maintenance requirements. This type of electric machine is also of interest to car companies, where development of electric vehicles (EVs) and hybrid electric vehicles (HEVs) continues to grow rapidly. For an EV/HEV drive system, a knowledge of motor torque is mandatory for precision of the control system, especially for security during car accidents when the electric feed of the traction motor must be stopped quickly. Avoiding the use of a torque meter in EV/HEV propulsion technologies makes the drive system more compact and more reliable. Thus, torque estimation has been accounted for over the last decade [1–7].

The electromagnetic torque of an electric motor can be obtained if the stator current and machine flux information is available. As the currents are measured, the torque can be estimated if the flux is well estimated. The well-known method for flux estimation is based on Faraday's law, where the flux is estimated by integration of electromotive force (EMF). This estimation is generally performed in the two axes of the stationary reference frame known as the $\alpha$ and $\beta$ axes. The advantages of this method are simplicity, low dependence on magnetic parameters and no requirement for rotor position [8]. The unknown initial magnetic condition of the estimator, divergence of estimated flux due to offsets in EMF and dependency on stator resistance are the main drawbacks of this method. To remove the first two inconveniences, replacing the integrators with low-pass filters (LPFs) was proposed [9]. The LPF can solve the problem of the unknown initial condition and DC offset problems of EMF, but attenuation for amplitude and shift of the phase of the estimated flux are encountered [10]. To solve the latter problem, adding a compensator at the output of the LPF to compensate the attenuation/shift problem [11–13] was proposed. By the way, the modified LPF method with a fixed-frequency LPF accompanying a compensator is not a good solution for a variable-speed drive where the motor frequency is always changing [14]. Thus, it was proposed in [15] that the LPF is removed from the main estimator and a proportional-integrator controller is added to eliminate the DC value and higher harmonics from the...
estimated flux. Recently, a band-pass filter known as a second-
order generalized integrator (SOGI) was proposed, instead of
the modified LPF integrator, for use in variable-speed drives [16, 
17]. The SOGI method proposes to pick up the fundamental
frequency of EMF before integration. Both the modified LPF
and the SOGI method are capable of removing the afore-
mentioned flux estimator problems, but new problems of
observability at standstill and complexity are encountered. To
solve the observability problem at standstill, third-order,
fourth-order and cascade generalized integrators [18, 19] were
proposed, which makes the estimator increasingly complex. 
Even with complex cascade SOGI topologies for flux esti-
mation, deviations in estimated flux can occur at very low
speeds based on the experimental results because of the very
high sensitivity of the filter to DC values at null speed.

The flux of a PMSM can also be estimated based on the
relation between the flux and currents; this is known as the
current model [20, 21]. This estimation is highly dependent
on the variation in machine inductances and permanent magnet
flux, was proposed [27]. The SOGI method was proposed in [27]
to solve the observability problem at standstill, third-order,
fourth-order and cascade generalized integrators [18, 19] were
proposed, which makes the estimator increasingly complex. 
Even with complex cascade SOGI topologies for flux esti-
mation, deviations in estimated flux can occur at very low
speeds based on the experimental results because of the very
high sensitivity of the filter to DC values at null speed.

The research objective is to use a very simple voltage-model flux
estimator for torque estimation by adding a modified current-
model estimator for a short percentage of the operation time. It
is shown that to eliminate the effect of the unknown initial
condition of the voltage-model estimator (DC drift in estimated
flux), it is not necessary to add a filter for overall time, which
would result in complexity as well as deviations in flux/torque
estimation at zero/low speed. The voltage model can be reinit-
ialised at some necessary points to return it to its original
condition. For this, it is proposed that only two pure integrators
are applied to EMF in the α and β axes to obtain flux informa-
However, the modified current-model observer proposed in [27]
is employed as an auxiliary estimator to correct the non-actual
initial condition of the voltage-model estimator at necessary
points. The latter estimator is activated only for a very low per-
centage of the operation time, if needed, to correct the initial
condition and is off for the remainder of the time. The estimator
is simple enough but without the problem of instability/
observability at zero speed, because it has neither a closed-loop
configuration nor filters. A very simple offset detector is also
designed for the estimator in order to avoid any offset of the
EMF before integrating.

2 | TORQUE/FLUX ESTIMATION

The electromagnetic torque of a PMSM based on flux and
currents represented in the stationary reference frame can be
estimated by (1)

\[ T_e = \frac{3}{2} p (\lambda_\alpha i_\beta - \lambda_\beta i_\alpha) \]  

(1)

where \( p \) are the pole pair numbers. Thevariables \( \lambda \) and \( i \)
represent, respectively, total flux and stator current. From (1)
and because the current measurements are available, the motor
torque is obtained if the total flux are estimated. Two well-
known methods of flux estimation based on the voltage and
current models are represented in this section. The voltage
model is associated with the most efficient filters, modified
LPF and SOGI, whereas the current model is improved by an
observer.

2.1 | Voltage-model flux estimation

The flux linkages of a PMSM in a stationary reference frame
can be estimated by (2)

\[
\begin{align*}
\lambda_\alpha &= \int (v_\alpha - R_i i_\alpha) \, dt + \lambda_\alpha(0) \\
\lambda_\beta &= \int (v_\beta - R_i i_\beta) \, dt + \lambda_\beta(0)
\end{align*}
\]  

(2)

where \( R_i \) is stator resistance. The variables \( v \) and \( e \) represent
stator voltage and back-EMF, respectively. The initial flux is
defined as \( \lambda(0) \) in (2). Based on the current–flux PMSM model,
the flux of the machine in a stationary reference frame can also
be presented as

\[
\begin{align*}
\lambda_\alpha &= \left[ L_\Sigma + L_\Delta \cos(2\theta) \right] i_\alpha + L_\Delta \sin(2\theta) i_\beta + \phi_{f0} \cos(\theta) \\
\lambda_\beta &= \left[ L_\Sigma - L_\Delta \cos(2\theta) \right] i_\beta + L_\Delta \sin(2\theta) i_\alpha + \phi_{f0} \sin(\theta)
\end{align*}
\]  

(3)

with \( L_\Sigma = (L_{\alpha0} + L_{\beta0})/2 \) and \( L_\Delta = (L_{\alpha0} - L_{\beta0})/2 \), and where
\( \phi_{f0}, L_{\alpha0} \) and \( L_{\beta0} \) are the constant magnetic parameters of the
model in nominal condition as permanent magnet flux and the $d$- and $q$-axis stator inductances, respectively. Based on (2), a non-real value for the initial flux condition induces a non-true offset for the estimated flux. Regarding (3), if the rotor position ($\theta$) is measured, this circumstance can occur at the starting point due to errors between the assumed nominal parameters ($\phi_{f0}$, $L_{d0}$ and $L_{q0}$) and real parameters at the starting point. It can also occur during motor function because of demagnetisation and magnetic saturation that can cause differences between the real and assumed magnetic parameters. Consequently, the offset created in the estimated flux induces nontrue oscillations in the estimated torque defined in (1). The second problem that can occur, due to the open-loop integrator of the estimator (2), is divergence of the estimated flux/torque in the presence of any offset in EMF. The offset can result from offsets in the measurement and/or non-linearity of the inverter. The following methods are proposed to solve these problems. The following provides a brief explanation of these.

### 2.1.1 Modified low-pass filter

By adding a high-pass filter to the output of the integrators in (2), the problems of offset and divergence of the estimated flux is eliminated. In other words, the pure integrators are replaced with an LPF with a cut-off frequency, $\omega_c$, defined in Laplace domain as $1/(s + \omega_c)$. To compensate for the attenuation/shift of sine wave estimated flux in steady state, a compensator defined in phasor domain as $(\jmath \omega_c + \alpha_c)/(\jmath \omega_c)$ is introduced as shown in Figure 1 [14]. The method is efficient enough for medium/high speeds when rotor speed is nearly constant. For variable-speed drives, the frequency of filter ($\omega_c$) must be adapted dynamically with the stator frequency, which is not possible with this proposed topology. Furthermore, transient operations including zero-speed crossing are not guaranteed with this method. Similar approaches are presented in [11] that use nearly the same idea and can be categorised as modified LPF with the same problems mentioned previously.

### 2.1.2 Second-order generalized integrator

Recently, a second-order filter known as the SOGI has been proposed for use instead of the modified LPF method for variable-speed drives. The filter is considered a band-pass filter that picks up the fundamental frequency of EMF before the integral action. The scheme-block of the SOGI method applied for flux estimation is shown in Figure 2. The transfer function of the SOGI filter is defined as $k\alpha/(\beta^2 + k\alpha^2 + \alpha^2)$ in the Laplace domain, where $k$ is the damping factor and $\alpha$ is the natural frequency of the filter [34]. The method is rather complex with more calculation time because it uses two integrators for each axis (four integrators in total). Furthermore, estimation based on the SOGI method is very sensitive at zero speed, which causes deviations in flux estimation at standstill [19]. Cascade stage SOGIs were proposed to reduce the sensitivity of the estimator, which made the estimator more complex. In [17], eight integrators based on the cascade SOGI were proposed to estimate the machine flux in the $a$ and $b$ axes.

In Figure 3, the deviations in estimated flux based on the modified LPF and SOGI methods are shown, where a sample simulation test is considered for a reversal-speed profile.

### 2.2 Current-model flux estimation

As mentioned in (3), total flux can also be estimated by the current–flux PMSM model simplified in a synchronous rotating reference frame as

$$\begin{aligned}
\lambda_d &= \phi_{f0} + \varphi_d = \phi_{f0} + L_{d0}i_d \\
\lambda_q &= \varphi_q = L_{q0}i_q
\end{aligned}$$

where $\varphi_d$ and $\varphi_q$ are stator fluxes generated by stator currents. The parameters $\phi_{f0}$, $L_{d0}$ and $L_{q0}$ may not be available or may change during operation due to phenomena such as magnetic saturation and demagnetisation. Thus, exact identification of these parameters as well as their observation during motor function is mandatory for the estimator (4). Assuming a difference between the parameters of the estimator (4) and those of the real machine, it can be represented as [27]

$$\begin{aligned}
\hat{\lambda}_d &= \phi_{f0} + L_{d0}(i_d - g_d) \\
\hat{\lambda}_q &= L_{q0}(i_q - g_q)
\end{aligned}$$

with

![Figure 1: Modified low-pass filter flux estimator](image1)

![Figure 2: Second-order generalized integrator flux estimator](image2)
Flux estimation based on voltage model with modified low-pass filter and second-order generalized integrator methods for a reversal-speed profile including zero-speed crossing.

\[ g_d = (\Delta L_d^{-1}) \phi_d - \frac{\Delta \phi_f d}{L_d} \]  \[ g_q = (\Delta L_q^{-1}) \phi_q \]  

and

\[ \Delta L_d^{-1} = \frac{1}{L_{d0m}} - \frac{1}{L_{d0}} \]  \[ \Delta L_q^{-1} = \frac{1}{L_{q0m}} - \frac{1}{L_{q0}} \]  \[ \Delta \phi_f = \phi_{\phi0m} - \phi_{\phi0} \]  

where, \( L_{d0m}, L_{q0m} \) and \( \phi_{\phi0m} \) are the exact values of the stator inductances and the real value of permanent magnet flux for the machine, respectively, while \( \Delta L_d^{-1}, \Delta L_q^{-1} \) and \( \Delta \phi_f \) represent the deviations between the real and first analytical evaluations of those parameters.

A precision estimation by (5) is achieved if the unknown variables \( g_d \) and \( g_q \) are observed using motor voltages and currents. Considering these unknown terms in the dynamic equation of the stator currents of a PMSM, the estimation is obtained as follows:

\[ \frac{d i_d}{d t} = \frac{-R_s L_{d0}}{L_{d0}} i_d + \frac{\omega_s L_{q0}}{L_{d0}} q_d + \frac{\omega_s L_{q0}}{L_{q0}} q_d + \frac{v_d}{L_d} + \frac{d g_d}{d t} \]  \[ \frac{d i_q}{d t} = \frac{-R_s L_{q0}}{L_{q0}} i_q + \frac{\omega_s L_{d0}}{L_{q0}} q_d + \frac{\omega_s L_{d0}}{L_{q0}} q_d + \frac{v_q}{L_q} + \frac{d g_q}{d t}. \]  

By accounting for unknown variables \( g_d \) and \( g_q \) as piecewise constant functions, an observer can be defined as

\[ \dot{x} = A(\omega) x + v(\omega, v_d, v_q) - K(C x - y) \]  \[ \dot{y} = C x \]  

where \( K \) is the gain of the observer, with

\[ x = \begin{pmatrix} i_d \\ i_q \\ g_d \\ g_q \end{pmatrix}, A(\omega) = \begin{pmatrix} -\frac{R_s}{L_d} & \frac{\omega_s L_{q0}}{L_d} & 0 & \frac{\omega_s L_{q0}}{L_d} \\ 0 & -\frac{R_s}{L_{q0}} & \frac{\omega_s L_{d0}}{L_{q0}} & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{pmatrix} \]  \[ y = \begin{pmatrix} i_d \\ i_q \end{pmatrix}, v = \begin{pmatrix} \frac{v_d}{L_d} \\ \frac{v_q - \omega_s \phi_f}{L_{d0}} \end{pmatrix}, C = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{pmatrix}. \]  

By observing the variables \( g_d \) and \( g_q \) from (9) (10) (11) and considering them as inputs for the estimator (5), the total flux of the machine is available even with magnetic parameter uncertainties due to magnetic saturation and demagnetisation. Finally, the estimated flux by (5) can be transformed into the stationary reference frame to be represented in the same frame as the estimator (2).

It has been proved in [27] that there is no explosion for the estimated states even at zero speed when the gain of the observer is calculated by the Riccati equation as follows:

\[ \frac{d P}{d t} = -A^T P - PA - P Q P + C^T R^{-1} C \]  \[ K = P^{-1} C^T R^{-1}. \]  

where the \( Q \) and \( R \) matrices are symmetric definite positive matrices and are the weighting matrices to tune. Indeed, observer gain is a function of rotor speed where it goes to zero at standstill. However, the observer is complex due to the need to solve the Riccati equation in matrix form with the same system rank. This deterministic version of the Kalman filter is
also called the Kalman-like observer. For other types of observers such as Luenberger or high gain, a problem with deviation at zero speed is expected similar to that of the voltage-model estimator with filters.

3 | PROPOSED FLUX ESTIMATOR

3.1 | Combination of voltage and current models

The pure voltage-model flux estimator without filters and compensators provides good estimation for a wide range of rotor speeds, including zero speed, if the rotor position is available and the problems of unknown initial conditions and EMF offset do not exist. The current-model estimator is accurate but has the drawbacks of observer complexity and the need to tune the observer parameters for different speeds. To capture the advantages of both estimators, it is proposed that a combination of voltage and current model estimator is used as the main estimator, while the current model estimator is only activated for a very short percentage of operation time, if needed. The model of the proposed estimator is represented as

\[
\begin{align*}
\widetilde{\lambda}_{a-volt} &= \int (\psi_a - R_a) dt, \quad \widetilde{\lambda}_{a-curr}(0) = \widetilde{\lambda}_{a-curr}(0) \\
\widetilde{\lambda}_{\beta-volt} &= \int (\psi_\beta - R_\beta) dt, \quad \widetilde{\lambda}_{\beta-curr}(0) = \widetilde{\lambda}_{\beta-curr}(0)
\end{align*}
\]  
(13)

It should be noted that the integrators used in (13) are resettable with the capability to be reinitialised during the function. Any magnetic uncertainties at the starting point of the estimation and magnetic parameter deviations during motor function due to magnetic saturation and demagnetisation are detected by the observer to modify the flux estimates of the current model. These are then considered the new initial conditions for the integrators in the voltage-model estimator. After each initialisation, the current model with observer is switched off until its next use to minimise execution time with an aim to avoid heating the processor. To turn the observer-based current model on and off, a new command for the estimator is necessary. The reference torque is proposed as a control system signal to compare with the estimated torque obtained based on the estimated flux. Indeed, a detector is designed to compare the absolute values of the reference and estimated torques by considering a predefined threshold, \( \epsilon \), to be the allowable difference. For upper differences compared with \( \epsilon \), it is assumed that a change in the magnetic condition of the motor has occurred that requires reinitialisation of the estimator. To implement this rule, a Boolean operator named ‘RES’ is introduced to reset and reinitialise the voltage-model integrators. The ‘RES’ operator is defined as

\[
RES = \begin{cases} 
1 & |\widehat{T}_e - T_e| > \epsilon \\
0 & |\widehat{T}_e - T_e| \leq \epsilon.
\end{cases}
\]  
(14)

The current-model estimator is activated and deactivated when the operator ‘RES’ is set to ‘1’ and ‘0’, respectively. To avoid multiple reinitialisations, it is proposed that a very small delay in the output of the RES operator is added as follows:

\[
RES(t) = RES(t - \tau)
\]  
(15)

where \( \tau \) is the time delay for the reinitialising action.

3.1.1 | Threshold (\( \epsilon \)) and time delay (\( \tau \)) selection

Both parameters \( \epsilon \) and \( \tau \) are defined based on experimental conditions and are tuned by the user. For \( \epsilon \), it is proposed that this parameter is set as a percentage of reference torque that defines an allowable tolerance between the exact and reference values of machine torque. The effect of maximum acceptable demagnetisation or magnetic saturation on tolerance must be accounted for to achieve good tuning of the threshold. The user can set the delay, \( \tau \), to between 1 and 10 times the sampling time to account for system safety in cases where the objective of torque estimation is vehicle security.

Another advantage of this method is that the current-model estimator is operating as a backup estimator. Indeed, if the operator ‘RES’ wrongly goes to 1 due to a problem in the control system and consequently the reference torque, the estimator is reinitialised for the entire operation time, which means the current-model estimator is now entirely activated without any interruption for the estimator.

For the proposed method, it is important that the usage time percentage of the observer is calculated during operation. For this, a new symbol, \( OT \), is introduced as

\[
OT(\%) = \frac{\sum t_i}{t_{tot}} \times 100
\]  
(16)

where \( \sum t_i \) is the sum of the intervals that the observer remains in ‘ON’ mode for the total time of operation of motor \( t_{tot} \). The remainder of the time, the fluxes are estimated only by the two integrators introduced in (2). Furthermore, the current model with observer still requires parameter tuning (Q and R) for the whole speed range, particularly for very low speeds, whereas the proposed algorithm provides some relaxation on this point.

3.2 | Electromotive force offset compensator

The problem of divergence due to offsets in EMF can be avoided by adding a simple online offset estimator. The
estimator can be designed based on the mean value of the EMF. The mean EMF value is calculated by

\[
\begin{align*}
    e_{\alpha, \text{mean}} &= \frac{1}{T} \int_{T}^{+T} e_{\alpha} dt \\
    e_{\beta, \text{mean}} &= \frac{1}{T} \int_{T}^{+T} e_{\beta} dt
\end{align*}
\]

(17)

where \( T \) is the period of stator currents and is considered an input variable parameter for the offset detector. The obtained offset from (17) is subtracted from the EMF signal before using it in (13).

A general scheme-block of the proposed method based on the combination of the voltage and current models for torque estimation of PMSMs is illustrated in Figure 4. As shown, the voltage-model estimator functions as the main estimator to estimate the stator flux required for torque estimation. At the same time and as needed (from the criteria of (14)), the current-model flux estimator function as an auxiliary flux estimator that is turned on to correct the initial condition of the main estimator.

4 | RESULTS AND DISCUSSION

To illustrate the effectiveness of the proposed estimator, the results of several simulation and experimental tests are represented in this section. The tests were performed for a 3 kW PMSM using classical vector control. The nominal parameters of the motor are shown in Table 1.

4.1 | Simulation results

To show the drawbacks of the unknown initial flux conditions and offset of EMF, two simulation tests at a constant speed of 1000 rpm are considered. The respective results of the two tests are shown in Figures 5 and 6. For both tests, two step changes, for the reference torque from 0 to 9 Nm and from 9 to 0 Nm, are considered. Based on the results shown in Figure 5, there is DC drift in the estimated flux (Figure 5b) due to differences between the initial condition of the estimator and the real machine. Consequently, a non-true oscillatory term is induced in the estimated torque (Figure 5a). For the second test, a very small offset of 0.1 V is added to the EMF calculated by commanded voltages and measured currents. To see only the effect of EMF offset on the results, the second estimator based on the current model is activated only at one point after starting to modify the initial condition. The results shown in Figure 6a,b indicate that the estimated flux and torque are diverged because of the offset in EMF.

4.1.1 | Robustness tests

The last test is repeated a final time by initialising the initial conditions as well as subtracting the EMF offsets during the

| Symbol | Quantity | Value |
|--------|----------|-------|
| \( L_d \) | \( d \)-axis linear inductance | 3.5 mH |
| \( L_q \) | \( q \)-axis linear inductance | 5 mH |
| \( \psi_f \) | Permanent magnet flux | 0.144 Wb |
| \( p \) | Number of pair poles | 3 |
| \( T_f \) | Nominal torque | 9 Nm |
| \( (30/\pi)\omega_{ns} \) | Nominal speed | 3000 rpm |
| \( R_s \) | Stator resistance | 0.5 \( \Omega \) |

**TABLE 1** Permanent magnet synchronous machine parameters

**FIGURE 4** Scheme-block of the proposed method
motion based on the proposed method. Four parameters, \( Q \), \( R \), \( \epsilon \) and \( \tau \), are considered for the observation process and described in Table 2.

![Figure 5](image-url) Simulation results: (a) real, applied and estimated torque without reinitialised estimator, (b) real and estimated total flux in \( \alpha \)-axis without reinitialised estimator

![Figure 6](image-url) Simulation results: (a) real, applied and estimated torque without compensating electromotive force (EMF) offset, (b) real and estimated total flux in \( \alpha \)-axis without compensating EMF offset

This time with a reversal-speed profile started from 1000 rpm at 0 s, crossing zero speed at 1 s and finally arriving at -1000 rpm at 2 s. The results of the test are shown in Figure 9, where there is a very good agreement between estimated flux/torque and the real values without any explosion of the estimated variables at the zero-speed crossing (see Figure 9a,c). It can be seen that pure integrators with appropriate (actual) initial conditions estimate flux well, and deviations at zero speed do not occur, because the filters are not used. As illustrated in Figure 9b, parameter \( OT \) is calculated as 1.15% for the test, which proves that only two simple, pure integrators are working as the flux estimators for a high percentage of the operation time. It should be noted that the estimated torque shown in Figure 9a presents the motor torque provided (electromagnetic torque) when fixed by the reference torque. Thus, there are no changes for the provided torque during reversal speed. Surely, some change is expected in the torque applied to the motor shaft due to viscous and inertial effects. The latter test is performed a final time with the speed held at zero for 0.4 s to show the performance of the estimator at standstill. The results of this test are presented in Figure 10. As can be seen in the results, the estimator is still effective, and the real flux/torques are the same as those that were estimated, because the current-model observer has already corrected the initial condition. As proved in [27], the gain of the observer (10) at standstill goes to zero, and the observer stays in stable condition. However, it should be noted that the observer system (9) is no longer observable at standstill, and if a new change occurs for the initial condition at standstill, the current-model observer cannot correct the initial condition up to the new arriving speed.

| Table 2: Observer parameters |
|-----------------------------|
| Symbol | Quantity | Value |
| \( Q \) | Tuning matrix gain | \( \text{diag}[1 \ 1 \ 100 \ 100] \) |
| \( R \) | Tuning matrix gain | \( \text{diag}[0.01 \ 0.01] \) |
| \( \epsilon \) | Threshold | 5% of reference torque |
| \( \tau \) | Time delay | 0.001 s |

![Figure 7](image-url) Simulation results at a constant speed of 1000 rpm: (a) real, applied and estimated torque with reinitialised estimator and compensating electromotive force (EMF) offset, (b) intervals when the observer is ON, (c) real and estimated total flux in \( \alpha \)-axis with reinitialised estimator and compensating EMF offset
The last simulation test is dedicated to showing the robustness of the proposed method during motor operation and when partial demagnetisation has occurred. The results of the test are illustrated in Figure 11. It is assumed that a 5% demagnetisation occurs at time 0.25 s during motor operation at 1000 rpm with constant demanded torque. In this circumstance, it is expected that higher stator currents are necessary to maintain the requested torque. The presented results show that the detector activates at time 0.25 s to reinitialise the estimator based on the new magnet flux. It can be seen that the observer is activated for a very short time (almost 0.04 s) to reinitialise the integrator and then switched off. The results confirm that the torque is well estimated by the proposed method even after demagnetisation. The result of non-initialisation of the estimator is shown as a green line in the same figure to visually compare them and show the importance of online reinitialisation of the voltage-model estimator after motor demagnetisation based on the proposed method. Because the total flux of the machine is generated by the rotor (magnet) as well as the stator flux, the same situation is expected during magnetic saturation where changes are caused in stator inductances defined within the synchronous rotating reference frame. By the way, the effect of magnetic saturation on torque for the tested machine does not exceed the predefined threshold, so the results are not presented.

4.2 | Experimental results

An experimental setup composed of an interior PMSM (IPMSM) rated at 3 kW supplied by a three-phase voltage source inverter is arranged for the experimental tests. A 400 V DC voltage is applied to the inverter with insulated gate bipolar transistors as power switches. A PMSM motor is mechanically coupled with the shaft of the IPMSM to generate the required load torque. A torque meter is installed between their shafts. The switching frequency is set to 10 kHz using the classical sampled natural pulse-width modulation technique generated by a dSPACE DS1103. The digital board of dSPACE receives the stator currents and the DC link voltage data with a 10 kHz frequency and the measured torque data with a 2 kHz frequency. An encoder is also used to measure the position. The rotor speed data is calculated by a Kalman filter (to avoid using a pure derivator due to the presence of noise) applied to the measured position. A photograph of the test bench is shown as Figure 12. To compare the estimated torque results with those measured by the torque meter, the effects of motor inertia at transient speeds as well as motor friction at constant speed are subtracted from the estimated electromagnetic torque. For this, a simple acceleration estimator based on a Kalman filter is used. Then, the estimated torque to be compared with the measured torque is calculated by
$T_L(shaft) = T_e - J_m \ddot{\alpha} - B_m \omega_m$ \hspace{1cm} (18)

where $J_m$, $B_m$, $\omega_m$ and $\alpha$ are the moment of inertia, friction coefficient, mechanical rotor speed and acceleration of the motor, respectively.

First, the same experimental tests as those of the simulation are considered. Figure 13 illustrates the electromagnetic and load torques of the PMSM estimated by (1), where zero is considered the initial condition. The rotor speed is set to 1000 rpm while step change requested torques are applied to the machine. For this test, the offsets in EMF are not known but are compensated by (17). Similar to the simulation results, an oscillatory term is added to the mean torque due to the non-zero initial magnetic condition of the real motor. In Figure 13, the electromagnetic and requested torques are compared (Figure 13a), while the estimated torque on the motor shaft by (18) is compared with that measured by the torque meter (Figure 13b). Figure 14 represents the flux/torque results from another test with the same speed and torque profile with and without offset compensation. For this test, at the starting point, the integrators are reinitialised one time by the estimated flux from the second current-model estimator. The results show that the estimated torque (see Figure 14a) and estimated flux (see Figure 14b) diverge due to the presence of offset in the EMF. Based on the divergence rate of the estimated flux and estimated torque, it can be found that the real offsets, under conditions of the test, are not the same for different torques. However, they show no divergence after compensation, which proves the effectiveness of the online offset compensator. For the third experimental test, again with the same speed and torque profile, the second current-model estimator is temporarily activated based on the reset operator decision. Furthermore, the offset compensator is applied to the estimator online. The experimental results are shown in Figure 15 and confirm good agreement between the measured and estimated torque in the presence of unknown magnetic conditions and the offset of EMF. As shown in Figure 15b, the second current-model estimator with its associated observer is activated for only 1.8% of the operation time. In the real operation of an EV/HEV system, $OT$ is expected to have a very low value because the total time $t_{ot}$ is significantly higher.

**FIGURE 12** Experimental setup

**FIGURE 13** Experimental results at a constant speed of 1000 rpm:
(a) applied and estimated electromagnetic torque without reinitialised estimator and with compensating electromotive force (EMF) offset, (b) measured and estimated load torque without reinitialise estimator and with compensating EMF offset

**FIGURE 14** Experimental results at a constant speed of 1000 rpm:
(a) applied and estimated electromagnetic torque with reinitialise estimator and without compensating electromotive force (EMF) offset (b) estimated flux with reinitialised estimator and with/without compensating EMF offset

**FIGURE 15** Experimental results at a constant speed of 1000 rpm:
(a) measured and estimated load torque with reinitialised estimator and compensating electromotive force offset (b) intervals when the observer is ON
than the total time of the experimentation test. To see the performance of the estimator for higher speeds, nearly the same test is repeated at a constant speed of 2000 rpm. The results of this test are shown in Figure 16. As can be seen in the results, the estimator provides a good estimate of the torque with a reduced current-model interval (OT = 0.25%).

Secondly, a reversal-speed profile is considered, as shown in Figure 17a, to evaluate the estimation results under speed-reversal conditions. For this test, the results obtained by the proposed method are compared with those obtained by the pure integrator estimator, modified LPF estimator and SOGI estimator. The estimated flux and estimated torque are shown in Figures 17b and 18, respectively. Obviously, like constant-speed tests, the results of the pure integrator are not acceptable for reversal speed due to the DC drift as shown in Figures 17b and 18b. For the LPF estimator, a cut-off frequency of 10 Hz is considered. Good responses appeared neither at the starting point nor at the zero-speed crossing based on the results shown in Figures 17b and 18c. However, this method is simple and efficient enough for medium/high constant-speed operation, with higher values for the cut-off frequency, sharper estimation at the starting point, and worse results at zero crossing. Inversely, the lower filter frequencies result in slower estimation at the starting point but with better estimation at zero crossing. The SOGI estimator is tuned as \( k = \sqrt{2} \) and \( \omega' = \text{stator angular frequency}. \) A good estimation is shown in Figures 17b and 18d at the starting point and higher speeds, while the deviation still occurs at the zero-speed crossing. The estimation results based on the proposed method confirm good estimation at the starting point as well as the zero-speed crossing (see Figures 17b and 18a). As illustrated in Figure 17c, the parallel current-model estimator is turned on at critical points such as starting and zero-speed crossing, where the calculated OT is only 0.77%.

As with simulation tests, it is interesting to see the performance of the proposed estimator at standstill versus the other types of estimators based on the voltage model. The last experimental test is dedicated to the variable-speed profile including 5 s standstill in the presence of load torque. The speed profile is shown in Figure 19a, where the motor rotates at 1000 rpm (0–0.3 s), decelerates to 0 rpm (0.3–2.8 s), remains at zero speed for 5 s (2.8–7.8 s), accelerates in reverse direction (7.8–10.3 s) and finally rotates at 1000 rpm. The results of estimated flux and torque for the pure integrator, LPF, SOGI and proposed methods are shown in Figures 19b and 20. Based on these results, the estimated torque and measured torque are similar even at standstill because the initial conditions have been corrected by the current model. As mentioned

**Figure 16** Experimental results at a constant speed of 2000 rpm: (a) measured and estimated load torque with reinitialised estimator and compensating electromotive force offset (b) intervals when the observer is on

**Figure 17** Experimental results for a reversal-speed profile: (a) measured rotor speed (b) estimated \( \alpha \)-axis flux by the pure integrator, low-pass filter, second-order generalized integrator and proposed methods, (c), intervals when the observer is on

**Figure 18** Experimental results for a reversal-speed profile: (a) measured and estimated load torque by proposed method (b) measured and estimated load torque by pure integrator method (c) measured and estimated load torque by low-pass filter method (d) measured and estimated load torque by second-order generalized integrator method
in the explanation of the standstill simulation test, the observer presented in (9) has corrected the initial condition before standstill. Due to the lack of observability at zero speed for (9), if a change in initial condition occurs at zero speed due to changes in the magnetic parameters, the reinitialisation at standstill cannot provide the actual initial values, and an error will be created during zero-speed operation [27]. However, the error will disappear after a new initialisation when a non-zero speed is about to occur.

At the conclusion of the simulation and experimental results, the proposed estimator can observe the flux and torque of PMSMs for a wide range of speeds with quick response at starting points and without deviations at zero-speed crossings, while it mostly works with just two pure integrators.

5 | CONCLUSION

The combination of two estimators based on the voltage- and current-model observers of the PMSM was proposed for the flux and the torque estimation of a motor. The voltage-model estimator was proposed as the main estimator, while the current-model estimator would be temporarily activated based on the need of the voltage-model estimator to reinitialise its pure integrators. Thus, this eliminates the problem of DC drift in the estimated flux and consequently non-true oscillations on estimated torque. To avoid the problem of divergence in the estimated flux/torque, it was also proposed that the mean value of the EMF is subtracted before entry into the voltage-model estimator.

For the proposed method, the need for rotor position is mandatory. To develop a sensorless flux/torque estimator based on the proposed approach, we have started work in our laboratory to mix the proposed approach with high-frequency injection (HFI) methods where rotor position and speed are calculated based on HFI.

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