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Speed Sensorless Control of a Surface-mounted Permanent Magnet Drive

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Abstract— Permanent magnet synchronous machines are a popular choice for electric vehicle propulsion, with both interior and surface magnet rotor designs in use. Position feedback is required for commutation, with minimal phase delay. Speed data should be able to track the vehicle speed at maximum acceleration. Elimination of the shaft position sensor reduces cost and improves reliability, but can be challenging for surface magnet machines at low speeds, due to their lack of saliency. However, bread-loaf and inset magnets give sufficient saliency to make signal injection position estimation schemes viable. Experimental results are presented for a commercial 3.27kW machine, using signal injection at low speeds and transitioning to a novel model reference adaptive system back emf estimation scheme at high speeds, which employs an iterative search algorithm to reduce position tracking delays. The rationale for selection of filter frequencies and tuning gains is explained.

Index Terms—Permanent Magnet motor, Speed estimation, MRAS, signal injection.

I. INTRODUCTION

Following commitments to full-electric (EV) and hybrid-electric (HEV) vehicles by countries including the UK, France and China [1-3], there is intensive development work on vehicle drive-trains for electric propulsion. The majority of commercially-available domestic EVs use synchronous permanent magnet machines, which may be either interior permanent magnet (IPM) or surface-mounted permanent magnet (SMPM) machines [4]. Rotor angular position is required for commutation and speed feedback is used for decoupling and speed control. Traditionally, position is measured by an optical incremental encoder or magnetic resolver. However, the former is difficult to use in the high vibration environment of a road vehicle, and the latter can be expensive [5]. Speed can then be found from rate of change of position, often using a phase-locked loop (PLL) or observer [5-8]. Techniques to estimate speed and position are reviewed in [5-8].

Speed is generally estimated from back emf [5-8], but at low speeds the back emf tends to zero, so a different estimation method is required. Two inductance-based effects may be used to identify angular position: saliency and partial saturation of the magnetising flux path [5-15]. Both techniques are suitable for IPM machines, where the

rotor interior magnets increase the reluctance of the direct-axis compared with the quadrature axis, and the airgap distance can be small, with more likelihood of saturation. In contrast, SMPMs with arc-shaped magnets have similar reluctance in the airgap and magnet, and the airgap is large, to allow space for the magnets, so there is little saturation. However, recent papers have shown that other magnet shapes such as bread-loaf and inset [16-17] can offer performance benefits for SMPM machines. These rotor designs, shown in Fig 1, give sufficient saliency for signal injection methods to be viable at low speeds. The injected signal may be a pulse [9-11], or a pulsating or rotating high frequency sinusoid [12-15]. In this research, pulsed signals were not considered, since they need high bandwidth sensing and data acquisition. The rotating sinusoid was preferred to the pulsating one, which has been identified as showing long convergence times and instability [15].

Fig. 1 PM rotor designs, a) arc, non-salient, b) bread-loaf and c) inset, salient.

At higher speeds, speed is generally estimated from back emf [4-5], either by direct measurement or from a voltage model of the circuit equations. The direct measurement of the third harmonic of back emf requires additional circuits with access the neutral point of the motor windings [18]. The voltage model is sensitive to measurement and parameter errors [8], so the model-referenced adaptive system (MRAS) method has been used for better accuracy and disturbance rejection, often combined with stator resistance tracking [19]. Generally, a proportional–integral (PI) controller is used for speed estimation, but this can be challenging to tune across the operating range [20]. A novel iterative MRAS scheme, presented in [21] for induction machines and [22] for wind generators, shows better dynamic response and improved robustness to drive parameter changes in the medium speed range. In this paper, the novel MRAS is adapted for the SMPM motor drive, and evaluated against the classical MRAS model, particularly for the transition from low to medium speed sensorless operation.

Section II outlines the conventional and novel MRAS schemes for medium speeds and shows how a signal

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injection technique for low speed [5-8] has been adapted for SMPM machines. Section III shows the test system and presents results of sensorless speed control for an off-the-shelf, low-speed SMPM machine.

II. CONTROL AND ESTIMATION

A. Machine model and control

The SMPM machine equations in the rotor reference frame are [5]

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = R_s \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \omega_r \begin{bmatrix} 0 & -L_q \\ L_d & 0 \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ \omega_r \psi_m \end{bmatrix} \quad (1)$$

$$T_e = \frac{3}{2} p (\psi_m i_q + i_d i_q (L_d - L_q)) \quad (2)$$

where v_d , v_q are the stator voltage components; i_d , i_q are the stator current components, ψ_m is the flux linkage due to the permanent magnet, R_s is the stator resistance and L_d , L_q are the direct and quadrature self-inductances respectively, all expressed in the rotor reference frame. T_e is the electrical torque, ω_r is the electrical frequency and p is the pole pairs. Equations (1)-(2) can also be used to develop the field oriented control and maximum torque per amp control scheme presented in Fig.2, where the details can be found in [23] and ω_m , θ_m refer to mechanical speed and position.

Fig.2 Block diagram of speed control from [23].

B. MRAS speed estimation and position estimation

The conventional MRAS speed estimation compares estimated stator flux linkage from back emf (3), with stator flux linkage from magnet flux and current (4)-(5), where subscripts 'v' or 'i' are for the voltage and current models respectively. $\alpha\beta$ indicates components in the stationary reference frame.

$$\begin{bmatrix} \psi_{\alpha v} \\ \psi_{\beta v} \end{bmatrix} = \int \left(\begin{bmatrix} v_{\alpha} \\ v_{\beta} \end{bmatrix} - R_s \begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} \right) dt \quad (3)$$

$$\begin{bmatrix} \psi_{di} \\ \psi_{qi} \end{bmatrix} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \psi_m \\ 0 \end{bmatrix} \quad (4)$$

Equation (4) can be transformed to the stationary reference frame, using the inverse Park's transform (5), which depends on estimated rotor angle $\hat{\theta}_r$.

$$\begin{bmatrix} \psi_{\alpha i} \\ \psi_{\beta i} \end{bmatrix} = \begin{bmatrix} \cos \hat{\theta}_r & \sin \hat{\theta}_r \\ -\sin \hat{\theta}_r & \cos \hat{\theta}_r \end{bmatrix} \begin{bmatrix} \psi_{di} \\ \psi_{qi} \end{bmatrix} \quad (5)$$

In practice, (3) fails to converge for errors in initial

condition and measurement offsets [12], so was combined with a high pass filter with a cut-off frequency of 3 Hz. The cut-off was set as low as possible to allow investigation of the estimator's low speed performance. Voltages were estimated from reference values, using the measured DC link voltage and dead-time compensation [12].

Fig.3 Block diagram of (a) conventional MRAS and (b) novel MRAS. ($\hat{\omega}_m$ then replaces the ω_m in the control algorithm in Fig.2)

The magnitude of the cross product of the two estimated fluxes, ϵ , shown in (6), will go to zero if the rotor angle is correct.

$$\epsilon = \psi_{\alpha v} \psi_{\beta i} - \psi_{\beta v} \psi_{\alpha i} \quad (6)$$

In conventional MRAS, speed is estimated by a PI controller and integrated to get $\hat{\theta}_r$, which is fed back and used to drive the error, ϵ , to zero, as shown in Fig.3.(a). The convergence time depends on the controller gains, which were selected for tracking response at low speeds under load. Since (6) only uses the phase error between estimated fluxes, the magnitude error is often used for stator resistance compensation [19], but was not implemented in this study, since it was not the focus of the research.

For the novel MRAS, the angle estimation algorithm converges within a single PWM period, giving a single time-step delay in estimated orientation angle, and allowing faster tracking of speed variations.

The novel MRAS algorithm from [21] divides the 360° search space into j sectors and finds the sector with the smallest magnitude of ϵ . This sector is then subdivided and the search repeated iteratively. In this research, 8 sectors and 13 iterations gives an angle resolution of 0.005° for each search result. The flowchart of the search algorithm is presented in Fig.4, where $\Delta\theta_i$ is the i^{th} displacement angle calculated in (7) and iterated angle $\theta_{i,j}$ is found using (8). θ_{base} is initialised as the rotor angle from the previous time step, and $\epsilon_{initial}$ corresponds to a 45° orientation error. In practice, the search space can be reduced by initializing from the previous angle location.

Fig.4 Flow chart of search algorithm for novel MRAS method [21].

$$\Delta\theta_i = \frac{45^\circ}{2^i}, \quad i=0, \dots, 13 \quad (7)$$

$$\theta_{i,j} = \theta_{base} + \Delta\theta_i(j-4), \quad j=1, \dots, 8 \quad (8)$$

Estimated speed is computed using a phase-locked loop (PLL) based on the estimated position, shown in Fig.3.(b), which has better noise rejection than the differentiation method in [22].

C. Signal injection and angle extraction

The signal injection method is used at low speeds, where the back emf based methods are no longer applicable. A rotating high frequency (HF) sinusoidal signal (9) is injected into the stationary ($\alpha\beta$) reference frame. The analysis below follows [6].

$$\begin{bmatrix} v_{h\alpha} \\ v_{h\beta} \end{bmatrix} = V_h \begin{bmatrix} -\sin\omega_h t \\ \cos\omega_h t \end{bmatrix} \quad (9)$$

where v_h is the voltage signal magnitude and ω_h is the signal frequency.

In the stationary reference frame the machine model in (1) is rearranged as

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = R_s \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \begin{bmatrix} \Sigma L + \Delta L \cos 2\theta_r & \Delta L \sin 2\theta_r \\ \Delta L \sin 2\theta_r & \Sigma L - \Delta L \cos 2\theta_r \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \omega_r \begin{bmatrix} -\sin 2\theta_r & \cos 2\theta_r \\ \cos 2\theta_r & \sin 2\theta_r \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \psi_m \begin{bmatrix} -\sin\theta_r \\ \cos\theta_r \end{bmatrix} \quad (10)$$

where $\Sigma L = \frac{1}{2}(L_d + L_q)$, $\Delta L = \frac{1}{2}(L_d - L_q)$ and θ_r is the rotor equivalent electrical angle.

For HF injection, the inductance terms in (10) dominate, so the IR drop can be neglected, and the current response from the injected voltage signal can be approximated as (11).

$$\begin{bmatrix} i_{h\alpha} \\ i_{h\beta} \end{bmatrix} = \frac{V_h}{\Sigma L^2 - \Delta L^2} \begin{bmatrix} \frac{\Sigma L}{\omega_h} \cos\omega_h t + \frac{\Delta L}{\omega_h} \cos(2\theta_r - \omega_h t) \\ \frac{\Sigma L}{\omega_h} \sin\omega_h t + \frac{\Delta L}{\omega_h} \sin(2\theta_r - \omega_h t) \end{bmatrix} \quad (11)$$

From (11), the ratio of the modulation term to the high frequency term is dependent on the degree of saliency ($\Delta L/\Sigma L$) so does not depend on the absolute value of the inductance. Demodulation using estimated angular position $\hat{\theta}_{rHF}$ results in error, ε_{HF} , as in (12)

$$\begin{aligned} \varepsilon_{HF} = & -\frac{V_h}{\Sigma L^2 - \Delta L^2} \frac{\Sigma L}{\omega_h} \sin(2\hat{\theta}_{rHF} + 2\omega_h t) \\ & + \frac{V_h}{\Sigma L^2 - \Delta L^2} \frac{\Delta L}{2\omega_r - \omega_h} \sin(2\theta_r - 2\hat{\theta}_{rHF}) \end{aligned} \quad (12)$$

After low-pass filtering to remove the $2\omega_h$ term, a term proportional to the position error ε_{fHF} can be extracted, as in (13), which assumes ($\Delta L \ll \Sigma L$) and ($\omega_r \ll \omega_h$). The $\frac{\Delta L}{\Sigma L^2}$ term in (13) means that a low-inductance machine with significant saliency, as used in the lab, may result in higher HF currents per volt injected than an IPM machine.

$$\varepsilon_{fHF} \approx -\frac{V_h}{\Sigma L^2} \frac{\Delta L}{\omega_h} (\theta_r - \hat{\theta}_{rHF}) = k \Delta \theta_{rHF} \quad (13)$$

The signal injection frequency was set at 800Hz with voltage magnitude of $40V_{peak}$ under the $100V_{dc}$ supply condition [24]. Two digital second order band-pass filters were used before the demodulation to separate the desired signal from the motor fundamental current. The cut-off frequencies were set at 750Hz and 850Hz to maximize the quality of HF current response and minimize the delay caused by the filters [24]. A 50Hz second order digital low-pass filter was used to remove the double frequency term.

Speed was estimated from motor torque using a standard full-order Luenberger observer [25], rather than a PLL, in order to improve its dynamic response, as shown in Fig.5. The observer for speed and position information is designed as in (14) with a bandwidth of 10Hz after testing [24].

$$\begin{bmatrix} \dot{\hat{\theta}}_m \\ \dot{\hat{\omega}}_m \\ \dot{\hat{T}}_L \end{bmatrix} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & -\frac{\hat{B}}{J} & -\frac{1}{J} \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \hat{\theta}_m \\ \hat{\omega}_m \\ \hat{T}_L \end{bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{J} \\ 0 \end{bmatrix} T_e + \begin{bmatrix} l_1 \\ l_2 \\ l_3 \end{bmatrix} (\theta_m - \hat{\theta}_m) \quad (14)$$

where the observer correction term $\theta_m - \hat{\theta}_m = \Delta\theta_{rHF}/p$ is taken from the output of the signal injection stage.

Fig.5 Block diagram of signal injection estimation [24].

III. EXPERIMENTAL RESULTS

A. Test System

The experimental system is shown in Fig.6 and a full list of motor and controller parameters is presented in TABLE I. The novel MRAS scheme calculates the Park's transform 104 times per 100 μ s PWM cycle. This is feasible for the TMS320F28379D microcontroller with built-in maths co-processor, but would not be practical on a standard microcontroller.

Fig.6 Photo of the test rig.

TABLE I
MACHINE AND CONTROL PARAMETERS

Parameter	Value (Units)	Parameter	Value (Units)
<i>Motor parameters</i>		<i>Speed controller</i>	
Rated speed	3000 rpm	Bandwidth ¹	0.66 Hz
Rated voltage	400 V	Damping ¹	1.1
Pole pairs	3	<i>Conventional MRAS estimator</i>	
Rated Torque	7.8 Nm	K_p	10000 rpmV ⁻² s ⁻²
Rated power	3.27 kW	K_i	400000 rpmV ⁻² s ⁻³
L_d	6.12 mH	<i>Novel MRAS PLL</i>	
L_q	9.04 mH	Bandwidth ¹	2 Hz
R_s	0.8 Ω	Damping ¹	1.1
K_e	98 V/krpm	<i>Encoder PLL</i>	
K_t	1.6 NmA ⁻¹	Bandwidth ¹	10 Hz
<i>Inverter parameters</i>		Damping ¹	0.5
PWM frequency	10 kHz	<i>Signal injection</i>	
Dead time	2 μ s	V_h	40 V
DC link voltage	100 V	ω_h	800 Hz
Rated current	5.6 Arms	Bandwidth	10 Hz
<i>Current Controller</i>		Lowpass filter	50 Hz
Bandwidth	300 Hz	Bandpass filter	750-850Hz
<i>MRAS voltage estimator</i>		Highpass filter	3Hz

¹Bandwidth is stated as the natural frequency and damping factor is taken from the characteristic equation.

The TMDX IDDK379D inverter is supplied with a DC bus voltage of 100V from an external power supply, giving up to 35Vrms using sine-triangle PWM. Current is

measured with LM55 Hall effect current sensors, and torque through an RWT-320 in-line transducer. A 4096 line incremental encoder is used to derive speed and position measurements for comparison with the estimators. Controller values are written to an analogue output and displayed on an oscilloscope and exported numerically through the SCI interface to LabVIEW.

The test motor is a Unimotor 115UMD300CACAA SMPM machine, with a quoted constant synchronous reactance. The internal rotor geometry is not known, but measurements of inductance under locked rotor conditions for different rotor angles showed that it has a saliency ratio of 1.48, whereas for an IPM machine it is typically about 2. A Siemens induction machine and Control Techniques inverter are used to set the load torque.

Fig.7 Open-loop estimated speed with no load (top) and with loaded (bottom, at 45% of rated load) condition.

B. Test Conditions

Steady-state results shown in Fig.7 using the encoder feedback for control, demonstrate high noise for some estimation methods. However, the drive could still operate smoothly. It can be seen that the noise decreases when the drive is operating under load. Estimator noise can be reduced by increasing the high-pass filter cut-off

for flux estimation from the voltage model, or reducing the bandwidth of the estimator, but values were chosen to test the low-speed performance of these estimators.

Fig.8 Open-loop response for different estimation methods under no load condition (first two) and 45% of rated load condition (last two), for speed steps starting from 50rpm (first and third) and 150rpm (second and fourth) respectively.

C. Open-loop Speed Response

The different estimation results are compared in Fig.8 for open-loop estimation, when the encoder measurement is used for speed control. It should be noted that the signal injection test was tested separately, so may have experienced slightly different load behaviour.

It is apparent that all three estimators can follow the speed variation and track the encoder measurement. The novel MRAS shows less noise on the speed signal than the other two methods. The classical MRAS has higher noise under light load conditions, as it was tuned for dynamic response under load. In general, the open-loop responses of all three different estimation methods behave as well as the encoder measurement.

D. Sensorless Estimation for No-load Condition

The sensorless operation for the SMPM motor under different estimation methods are shown in Fig.9. The novel MRAS and classical MRAS lose synchronization in the overshoot region and no longer provide accurate tracking of the rotor speed at 50rpm, so only the signal injection method is shown for this condition. The estimated speed responses all show similar results to the encoder at 150rpm so can all be used at this slightly higher speed.

Fig.9 Sensorless operation for different estimation methods under no load condition, starting from 50 rpm (top) and 150rpm (bottom).

E. Sensorless Estimation for Loaded Condition

Fig.10 shows the SMPM motor sensorless operation under 45% of rated load. The classical MRAS, novel MRAS and signal injection estimation maintain speed control at 150rpm. However both MRAS methods show higher overshoot and faster response than with the encoder feedback. Again, only the signal injection method could be implemented at 50rpm of low speed region and it still performs well.

Speed tracking with the classical MRAS estimator was sensitive to PI gains. This estimator could track the speed reference with $k_p=700 \text{ rpmV}^{-2}\text{s}^{-2}$ and $k_i=5000 \text{ rpmV}^{-2}\text{s}^{-3}$ under light condition; however, $k_p=10000 \text{ rpmV}^{-2}\text{s}^{-2}$ and $k_i=400000 \text{ rpmV}^{-2}\text{s}^{-3}$ were needed for 45% rated load. This means that noise rejection and dynamic response could be improved by gain scheduling across the operating range. The more robust performance of the novel MRAS estimator is attributed to the fast estimation of rotor angle, allowing a lower cut-off frequency in the speed estimation. In contrast, the classical MRAS filter acts on both the speed and the angular position.

Fig.10 Sensorless operation under 45% rated load for different estimation methods, starting from 50 rpm (top) and 150rpm (bottom).

F. Sensorless Estimation for Switch Over Between Signal Injection Method and Novel MRAS method

Since the novel MRAS performs better than the classical MRAS method in terms of noise level and dynamics, the drive is now been tested under sensorless

operation to speed up to 250rpm from 50rpm, with switch over between the signal injection method at the lower speed and the novel method at the higher speed at same condition of 45% load. Fig.11 illustrates that the drive could operate smoothly under sensorless control, with no vibration during switch over between the different methods. In contrast, the classical MRAS method showed significantly higher speed variations, when combined with the signal injection.

Fig.11 Sensorless operation under 45% of rated load condition with switch over between signal injection and novel MRAS methods, together with encoder measurement.

IV. CONCLUSIONS

In this paper, sensorless speed control has been implemented successfully for an off-the-shelf SMPM machine, both for low speed estimation, and transitioning to medium speeds. A novel, iterative MRAS algorithm is shown to perform better than the standard MRAS method in the medium speed region, in terms of the trade-off between noise rejection and speed tracking under load and was also easier to tune. However, the novel scheme has significantly higher computational requirements.

In the low speed region, the signal injection method has been shown to work on a surface magnet PM machine with moderate saliency. However, the small average and differential inductance values compared with an IPM, make the implementation more challenging, resulting in significant audible noise and a reduction in speed loop bandwidth, due to increased filtering. Nevertheless it has been demonstrated as a viable technique, and the next step is to test on a high-speed integrated vehicle drive with a saliency ratio of 1.25.

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