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Flux-Linkage-Based Current Control of Saturated Synchronous Motors

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Abstract—Magnetic saturation characteristics of synchronous reluctance motors (SyRMs), with or without permanent magnets (PMs), are highly nonlinear. These nonlinear effects can be included in the current controller by changing its state variable from the stator current to the stator flux linkage using the known saturation characteristics. A direct discrete-time variant of the flux-linkage-based current controller is developed in a state-space framework. If the magnetics are modeled to be linear, the proposed control structure reduces to the standard current controller in this special case. Experimental results on a 6.7-kW SyRM drive demonstrate that the proposed flux-linkage-based controller enables a higher closed-loop bandwidth and is more robust against parameter errors, as compared to the standard current controller.

Index Terms—Current control, flux linkage, magnetic saturation, permanent-magnet synchronous motor, synchronous reluctance motor.

NOMENCLATURE

Vectors

Vectors in rotor coordinates are marked as follows.

i	Stator current $[i_d, i_q]^T$.
u	Stator voltage $[u_d, u_q]^T$.
ψ	Stator flux linkage $[\psi_d, \psi_q]^T$.

Vectors in stator coordinates are marked with the superscript s and their components are marked with the subscripts α and β , e.g., $i^s = [i_\alpha, i_\beta]^T$. Transposes are marked with the superscript T . Reference values are marked with the subscript ref .

Matrices

G_1	State-feedback gain matrix.
G_i	Integral gain matrix.
G_t	Reference-feedforward gain matrix.
I	Identity matrix $\begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$.
J	Orthogonal rotation matrix $\begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$.
K_1, K_2	State-feedback gain matrices.
K_i	Integral gain matrix.
K_t	Reference-feedforward gain matrix.
L_i	Incremental inductance matrix.
O	Zero matrix $\begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix}$.
Φ	State-transition matrix.

Other Symbols

R	Stator resistance.
s	Differential operator d/dt .
T_s	Sampling period.
u_{dc}	DC-bus voltage.
z	Forward-shift operator.
α	Closed-loop bandwidth.
β	Discrete-time pole $\exp(-\alpha T_s)$.
ϑ_m	Electrical angular position of the rotor.
ω_m	Electrical angular speed of the rotor.

I. INTRODUCTION

SYNCHRONOUS reluctance motors (SyRMs), with or without permanent magnets (PMs), have good flux-weakening capability and the high torque density. The high torque density of SyRMs and PM-SyRMs comes at a price of highly nonlinear magnetic saturation characteristics. A high-performance current controller is needed to optimally control these motors.

The focus of this paper is on the classical control methods equipped with a pulse-width modulator (PWM). The current controller operates typically in rotor coordinates [1]–[16], as illustrated in Fig. 1. In the case of highly saturated machines, such as SyRMs and PM-SyRMs, it is necessary to schedule the current controller gains to depend on the operating-point current. However, the transient performance may remain unsatisfactory despite significant tuning efforts, since the standard current controller cannot compensate for the dynamic saturation effects. Therefore, the closed-loop bandwidth has to be comparatively low in order to avoid large overshoots in the control response.

The measured current and the reference current can be easily mapped to the corresponding flux linkage variables using the known saturation characteristics. These flux linkages can then be used in the current controller, as originally proposed for a simple proportional controller in an early paper [17]. The flux-linkage-based current controller takes the saturation effects inherently into account, leading to a superior transient response under highly saturated conditions. If the magnetics are modeled to be linear, the flux-linkage-based current controller becomes mathematically equivalent to the standard current controller.

Apart from the magnetic saturation, discrete-time delays (sample-and-hold process and computational delay) may ruin the current control performance in high-speed or high-performance applications. Therefore, the current controller should preferably be designed directly in the discrete-time

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Fig. 1. Current-controlled motor drive, where the current controller operates in rotor coordinates. The vectors in stator coordinates are marked with the superscript s .

Fig. 2. Saturation characteristics of a 6.7-kW SyRM: (a) $\psi_d = \psi_d(i_d, i_q)$; (b) $\psi_q = \psi_q(i_d, i_q)$; (c) $\psi_d = \psi_d(i_d, 0)$. The saturation characteristics are obtained by fitting the measured data to the explicit model, given in the Appendix.

domain [9]–[15], if either a high fundamental frequency or a high closed-loop bandwidth (with respect to the sampling frequency) is required. Unfortunately, the saturation characteristics are more difficult to take into account in the direct discrete-time design, since the underlying discrete-time motor model relies on the assumption of constant inductances.

In this paper, we propose a flux-linkage-based current controller, based on the nonlinear change of state variables from the currents to the flux linkages, suitable for highly saturated synchronous motors. Both continuous-time and discrete-time variants are developed in a state-space framework. The saturation characteristics are described in a generic form, which makes the controller compatible with both look-up tables and explicit functions. The robustness against the errors in the magnetic saturation model is analyzed. The proposed controller is evaluated by means of experiments using a 6.7-kW SyRM drive. The preliminary version of this paper was presented in a conference [16].

Fig. 3. Model of a saturated synchronous motor in rotor coordinates.

II. MOTOR MODEL

A. Saturation Characteristics

The model of a saturated synchronous motor in rotor coordinates is considered. The effects of spatial harmonics and temperature changes are omitted. Generally, the stator flux linkage ψ is a nonlinear function of the stator current i , i.e.,

$$\psi = \psi(i) = \begin{bmatrix} \psi_d(i_d, i_q) \\ \psi_q(i_d, i_q) \end{bmatrix} \quad (1)$$

The reciprocity condition $\partial\psi_d/\partial i_q = \partial\psi_q/\partial i_d$ should hold, since the nonlinear inductor does not generate or dissipate energy. As an example, Fig. 2 illustrates the saturation characteristics of a SyRM, which exhibits also significant cross-saturation. In the special case of linear magnetics, the stator flux linkage is

$$\psi = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} i + \begin{bmatrix} \psi_f \\ 0 \end{bmatrix} \quad (2)$$

with the constant inductances L_d and L_q and constant PM-flux linkage ψ_f . It is worth noticing that the effect of PMs is inherently included in (1). Alternatively, the saturation characteristics can be defined using the inverse function of (1), i.e.,

$$i = i(\psi) = \begin{bmatrix} i_d(\psi_d, \psi_q) \\ i_q(\psi_d, \psi_q) \end{bmatrix} \quad (3)$$

In this paper, the saturation is modeled using generic saturation characteristics in the form of (1) or (3). The advantage of these forms is that they are directly compatible with both look-up tables and explicit functions. Furthermore, the definition of the chord-slope inductance, illustrated in Fig. 2(c), becomes ambiguous, if the machine is equipped with PMs.

B. Voltage Equation

The voltage equation in rotor coordinates is

$$\frac{d\psi}{dt} = u - Ri - \omega_m J \psi \quad (4)$$

where u is the stator voltage, R is the stator resistance, ω_m is the electrical angular speed of the rotor, and $J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$ is the orthogonal rotation matrix. The voltage equation (4) together with the saturation characteristics (3) forms a nonlinear state-space model of the motor, shown in Fig. 3.

Substituting (1) into (4) gives the voltage equation with the stator current as a state variable

$$L_i \frac{di}{dt} = u - Ri - \omega_m J \psi \quad (5)$$

Fig. 4. Current controllers: (a) current as a state variable; (b) flux linkage as a state variable.

where the incremental inductance matrix is

$$\begin{aligned} \mathbf{L}_i = \mathbf{L}_i(\mathbf{i}) &= \begin{bmatrix} \frac{\partial \psi_d(i_d, i_q)}{\partial i_d} & \frac{\partial \psi_d(i_d, i_q)}{\partial i_q} \\ \frac{\partial \psi_q(i_d, i_q)}{\partial i_d} & \frac{\partial \psi_q(i_d, i_q)}{\partial i_q} \end{bmatrix} \\ &= \begin{bmatrix} L_{di}(i_d, i_q) & L_{dqi}(i_d, i_q) \\ L_{dqi}(i_d, i_q) & L_{qi}(i_d, i_q) \end{bmatrix} \end{aligned} \quad (6)$$

The matrix is symmetric due to the reciprocity condition. Fig. 2(c) illustrates the definition of the incremental inductance. If the generic saturation characteristics are known, the definition of the incremental inductance is unambiguous in all operating point (unlike the definition of the chord-slope inductance).

III. CONTINUOUS-TIME CURRENT CONTROL DESIGN

Fig. 1 shows the block diagram of a current-controlled motor drive, where the current controller operates in rotor coordinates. The electrical angular position of the rotor is denoted by ϑ_m . In this section, the PWM is considered to be ideal, $\mathbf{u} = \mathbf{u}_{\text{ref}}$. The effects of the voltage saturation and of the time delays are considered later in Section IV. For simplicity, the current controllers are tuned assuming the stator resistance $R = 0$, i.e., the effect of the resistive voltage drop is compensated for by the integral control action.

A. Current as a State Variable

Fig. 4(a) shows the current control structure, where the magnetic saturation is taken into account based on (5). The voltage reference is given by

$$\mathbf{u}_{\text{ref}} = \omega_m \mathbf{J} \boldsymbol{\psi} + \mathbf{L}_i \boldsymbol{\xi} \quad (7)$$

where $\boldsymbol{\xi}$ is an auxiliary control variable, obtained from a linear controller to be designed subsequently. Assuming $\mathbf{u} = \mathbf{u}_{\text{ref}}$ and $R = 0$ and substituting (7) into (5) leads to

$$\frac{d\mathbf{i}}{dt} = \boldsymbol{\xi} \quad (8)$$

The closed-loop poles can now be easily placed via the linear part of the current controller. Here, a state-feedback controller with integral action and reference feedforward is used,

$$\boldsymbol{\xi} = \mathbf{G}_t \mathbf{i}_{\text{ref}} + \frac{\mathbf{G}_i}{s} (\mathbf{i}_{\text{ref}} - \mathbf{i}) - \mathbf{G}_1 \mathbf{i} \quad (9)$$

where $s = d/dt$ is the differential operator, $\mathbf{G}_t$ is the reference-feedforward gain, $\mathbf{G}_i$ is the integral gain, and $\mathbf{G}_1$ is the state-feedback gain. These gains can be chosen in various ways. As an example, choosing $\mathbf{G}_t = \alpha \mathbf{I}$, $\mathbf{G}_i = \alpha^2 \mathbf{I}$, and $\mathbf{G}_1 = 2\alpha \mathbf{I}$, where $\mathbf{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$ is the identity matrix, corresponds to the internal model control (IMC) design [1]. In this case, all the closed-loop poles are placed at $s = -\alpha$. The closed-loop reference-tracking dynamics are

$$\mathbf{i} = \frac{\alpha}{s + \alpha} \mathbf{i}_{\text{ref}} \quad (10)$$

where α is the bandwidth.

For implementing the control law (7), five nonlinear functions $\psi_d(i_d, i_q)$, $\psi_q(i_d, i_q)$, $L_{di}(i_d, i_q)$, $L_{qi}(i_d, i_q)$, and $L_{dqi}(i_d, i_q)$ should be implemented, typically with look-up tables, which complicates the control system. Furthermore, delays of the discrete-time implementation cause errors in the modeled incremental inductances, especially if the sampling frequency is low.

B. Flux Linkage as a State Variable

Fig. 4(b) shows the current control structure, where the flux linkage is chosen as a state variable. Similarly to (9), a state-feedback controller with integral action and reference feedforward is used,

$$\mathbf{u}_{\text{ref}} = \mathbf{K}_t \boldsymbol{\psi}_{\text{ref}} + \frac{\mathbf{K}_i}{s} (\boldsymbol{\psi}_{\text{ref}} - \boldsymbol{\psi}) - \mathbf{K}_1 \boldsymbol{\psi} \quad (11)$$

where $\mathbf{K}_t$ is the reference-feedforward gain, $\mathbf{K}_i$ is the integral gain, and $\mathbf{K}_1$ is the state-feedback gain.¹ Two typical selections for these gains are the IMC design [1]

$$\mathbf{K}_t = \alpha \mathbf{I} \quad \mathbf{K}_i = \alpha^2 \mathbf{I} \quad \mathbf{K}_1 = 2\alpha \mathbf{I} - \omega_m \mathbf{J} \quad (12)$$

and the complex vector design [4]

$$\mathbf{K}_t = \alpha \mathbf{I} \quad \mathbf{K}_i = \alpha \mathbf{I} (\alpha \mathbf{I} + \omega_m \mathbf{J}) \quad \mathbf{K}_1 = 2\alpha \mathbf{I} \quad (13)$$

Both these designs lead to the first-order closed-loop system

$$\boldsymbol{\psi} = \frac{\alpha}{s + \alpha} \boldsymbol{\psi}_{\text{ref}} \quad (14)$$

However, the disturbance rejection characteristics of the two designs are different, leading also to different sensitivities to parameter errors [12]. The step response corresponding to (14) is shown in Fig. 5.

¹The controller (11) does not see the magnetic saliency of the motor due to the mapping of the current to the flux linkage. Therefore, the controller (11) and the resulting flux linkage dynamics could be described using complex vectors and complex gains, instead of real vectors and gain matrices.

Fig. 5. Step responses of the continuous-time system (14) and the discrete-time system (23), where $\beta = \exp(-\alpha T_s)$. Setting $\beta = 0$ would give the dead-beat response. The green dashed line shows the response of the discrete-time system without the computational delay.

In the steady state, $\psi = \psi_{\text{ref}}$ holds due to the integral action. The same saturation model $\psi = \psi(i)$ is used to map both the reference current and the actual current to the corresponding flux linkages. Therefore, $i = i_{\text{ref}}$ holds in the steady state, even with inaccurate saturation model. The control structure in Fig. 4(b) needs only two look-up tables, $\psi_d(i_d, i_q)$ and $\psi_q(i_d, i_q)$, and it inherently takes the effects of the incremental inductances into account. It is also worth noticing that the two controllers in Fig. 4 are mathematically equivalent, if the magnetic saturation is omitted and if the same pole locations are chosen.

IV. DIRECT DISCRETE-TIME CONTROL DESIGN

For designing the current controller directly in the discrete-time domain, a hold-equivalent discrete-time model of the motor is needed. Furthermore, the limited inverter voltage has to be properly taken into account in the current controller, independently of the selected design approach.

A. Hold-Equivalent Motor Model

Fig. 6 shows the switching-cycle-averaged model of a PWM inverter in stator coordinates. The model consists of the voltage saturation (corresponding to the voltage hexagon), the computational delay of one sampling period, and the zero-order hold (ZOH). The effects of the voltage saturation are omitted here, but they are considered in the latter subsection.

In accordance with Fig. 6, the voltage $\mathbf{u}^s(t)$ in stator coordinates is assumed to be constant between consecutive sampling instants [12]. Furthermore, the stator resistance $R = 0$ is assumed. Under these assumptions, the exact hold-equivalent discrete-time model in rotor coordinates can be derived from (4), leading to

$$\psi(k+1) = \Phi \psi(k) + T_s \Phi \mathbf{u}(k) \quad (15)$$

where T_s is the sampling period. The state-transition matrix is

$$\Phi = \exp(-T_s \omega_m \mathbf{J}) \quad (16)$$

The resulting motor model is shown in Fig. 7. It can be seen that the saturation characteristics appear only in the output

Fig. 6. Model of a PWM inverter in stator coordinates. The model consists of the voltage saturation, the computational delay z^{-1} , and the ZOH.

Fig. 7. Hold-equivalent motor model in rotor coordinates. The state-transition matrix is $\Phi = \exp(-T_s \omega_m \mathbf{J})$.

equation due to the assumption $R = 0$. Therefore, a generic saturation model $i = i(\psi)$ can be used.

Due to the finite computational time, the digital control system typically has one sampling-period time delay, i.e., $\mathbf{u}^s(k) = \mathbf{u}_{\text{ref}}^s(k-1)$. For control design, the computational delay can be easily included in the plant model as follows

$$\begin{bmatrix} \psi(k+1) \\ \mathbf{u}(k+1) \end{bmatrix} = \begin{bmatrix} \Phi & T_s \Phi \\ \mathbf{O} & \mathbf{O} \end{bmatrix} \begin{bmatrix} \psi(k) \\ \mathbf{u}(k) \end{bmatrix} + \begin{bmatrix} \mathbf{O} \\ \Phi \end{bmatrix} \mathbf{u}_{\text{ref}}(k) \quad (17)$$

where $\mathbf{O} = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix}$ is the zero matrix. Due to this computational delay, the order of the discrete-time plant model is higher than the continuous-time model. An advantage of the discrete-time control design is that the delay can be easily taken into account in the controller, as discussed in the following.

B. Control Design

Similarly to the continuous-time design, the stator currents are mapped to the flux linkages. Furthermore, a state-feedback controller with integral action and reference feedforward is applied [12], [18]. The control algorithm is

$$\begin{aligned} \mathbf{u}_i(k+1) &= \mathbf{u}_i(k) + T_s \mathbf{K}_i [\psi_{\text{ref}}(k) - \psi(k)] \\ \mathbf{u}_{\text{ref}}(k) &= \mathbf{K}_t \psi_{\text{ref}}(k) - \mathbf{K}_1 \psi(k) \\ &\quad - \mathbf{K}_2 \mathbf{u}_{\text{ref}}(k-1) + \mathbf{u}_i(k) \end{aligned} \quad (18)$$

where $\mathbf{u}_i$ is the integral state, $\mathbf{K}_i$ is the integral gain, $\mathbf{K}_t$ is the reference-feedforward gain, and $\mathbf{K}_1$ and $\mathbf{K}_2$ are the state-feedback gains. Fig. 8 shows the corresponding block diagram, where also the anti-windup mechanism is included. It can be seen that the discrete-time control design is very similar to the continuous-time design. The computational delay is omitted in the continuous-time design, while it is taken into account in the discrete-time design by means of $\mathbf{K}_2$.

Based on (17) and (18), the closed-loop reference-tracking dynamics can be expressed as

$$\psi(k) = (z^3 \mathbf{I} + z^2 \mathbf{A}_2 + z \mathbf{A}_1 + \mathbf{A}_0)^{-1} (z \mathbf{B}_1 + \mathbf{B}_0) \psi_{\text{ref}}(k) \quad (19)$$

where z is the forward-shift operator and $\mathbf{A}_0$, $\mathbf{A}_1$, $\mathbf{A}_2$, $\mathbf{B}_0$, and $\mathbf{B}_1$ are the coefficient matrices. The gain matrices can be

Fig. 8. Discrete-time current controller. The voltage saturation and the anti-windup mechanism are also shown. The limited voltage reference $\bar{u}_{\text{ref}}$ can be calculated as a part of the current control algorithm or it can be provided by the PWM algorithm. In the linear modulation range, $\bar{u}_{\text{ref}} = u_{\text{ref}}$.

solved as functions of the coefficient matrices

$$\begin{aligned} K_t &= \frac{\Phi^{-2} B_1}{T_s} & K_i &= \frac{\Phi^{-2} (I + A_0 + A_1 + A_2)}{T_s^2} \\ K_1 &= \frac{I + \Phi^{-2} (I + \Phi + A_1 + A_2 + A_2 \Phi)}{T_s} \\ K_2 &= I + \Phi + \Phi^{-2} A_2 \Phi^2 \end{aligned} \quad (20)$$

Using these expressions, the poles and transmission zeros of (19) can be arbitrarily placed. Due to the time delay, $A_0 = O$ is preferably selected. The discrete-time counterpart of the IMC design is given by

$$A_1 = \beta^2 I \quad A_2 = -2\beta I \quad B_1 = (1 - \beta) I \quad (21)$$

and the complex vector design is given by

$$A_1 = \beta^2 \Phi \quad A_2 = -\beta(I + \Phi) \quad B_1 = (1 - \beta) I \quad (22)$$

where $\beta = \exp(-\alpha T_s)$ is the exact mapping in the discrete domain of the intended real pole of the system. In both cases, the closed-loop transfer-operator matrix (19) reduces to

$$\psi(k) = \frac{1 - \beta}{z(z - \beta)} \psi_{\text{ref}}(k) \quad (23)$$

The step response corresponding to (23) is shown in Fig. 5.

C. Voltage Saturation and Anti-Windup

Fig. 9 illustrates the maximum available voltage, which corresponds to the border of the voltage hexagon. In the first sector, the maximum voltage magnitude is [8]

$$u_{\text{max}} = \frac{u_{\text{dc}}}{\sqrt{3} \sin(2\pi/3 - \vartheta_u)} \quad (24)$$

where $\vartheta_u = [0, \pi/3]$ is the angle of the reference voltage u_{ref}^s . This equation can be easily applied in other sectors as well. The realizable voltage reference can be calculated as

$$\bar{u}_{\text{ref}} = \begin{cases} u_{\text{ref}}, & \text{if } \|u_{\text{ref}}\| \leq u_{\text{max}} \\ \frac{u_{\text{ref}}}{\|u_{\text{ref}}\|} u_{\text{max}}, & \text{if } \|u_{\text{ref}}\| > u_{\text{max}} \end{cases} \quad (25)$$

If this actuator saturation is not properly taken into account in the controller, the integral term accumulates the control error

Fig. 9. Voltage hexagon of a two-level PWM inverter in stator coordinates.

TABLE I
DATA OF THE 6.7-kW SYRM

Rated values		
Phase voltage (peak value)	$\sqrt{2/3} \cdot 370$ V	1.00 p.u.
Current (peak value)	$\sqrt{2} \cdot 15.5$ A	1.00 p.u.
Frequency	105.8 Hz	1.00 p.u.
Speed	3 175 r/min	1.00 p.u.
Torque	20.1 Nm	0.67 p.u.
Rated motor parameters		
d-axis inductance L_d	46 mH	2.20 p.u.
q-axis inductance L_q	6.8 mH	0.33 p.u.
Stator resistance R	0.55 Ω	0.04 p.u.

during $\|u_{\text{ref}}\| > u_{\text{max}}$, leading to undesired overshoots in the controlled current. Fig. 8 shows an integrator anti-windup mechanism, which is based on the realizable voltage reference $\bar{u}_{\text{ref}}$ [1], [13], [18], [19]. It can be seen that this mechanism has no effect in the linear modulation range, where $\bar{u}_{\text{ref}} = u_{\text{ref}}$ holds. The realizable voltage reference can be either calculated in the current controller using (24) and (25) or it can be obtained from the PWM.

It is also worth mentioning that the switching-cycle-averaged voltage at the k th time instant is $u(k) = \Phi \bar{u}_{\text{ref}}(k-1)$, according to (17). This signal is typically needed in the flux observer.

V. EXPERIMENTAL RESULTS

The current controllers are evaluated by means of experiments. The controllers were implemented on a dSPACE MicroLabBox DS1202 processor board. The studied motor is a transverse-laminated 6.7-kW four-pole SyRM, whose rated data are given in Table I and the saturation characteristics are shown in Figs. 2(a) and 2(b). A single-update PWM is used and the sampling (switching) frequency is 5 kHz. The stator currents and the DC-link voltage are measured in synchronism with the PWM. The angular position of the rotor is measured using a resolver. A servo induction machine is used to regulate the speed in the constant-speed tests.

The complex vector design is more robust against parameter errors than the IMC design [12]. Therefore, the gains corresponding to the complex vector design are used in the following. The desired closed-loop bandwidth is $\alpha = 2\pi \cdot 500$ rad/s. It is worth noticing that this bandwidth of 500 Hz is comparatively high for the controller operating at the sampling frequency of 5 kHz and controlling a motor with high saliency.

(a) Continuous-time design.

(b) Discrete-time design.

Fig. 10. Experimental results at $\omega_m = 1.5$ p.u. with the saturation model included in the controller: (a) continuous-time design; (b) discrete-time design. The measured current components i_d and i_q and their references are shown.

A. High-Speed Test

The effects due to the sample-and-hold process and the computational delay are significant when the fundamental frequency or the closed-loop bandwidth is high compared to the sampling frequency. Therefore, the following two controllers are compared by means of a high-speed test:

- 1) The continuous-time design of Fig. 4(b), discretized using the forward Euler approximation.
- 2) The proposed discrete-time design of Fig. 8.

In both these controllers, the magnetic saturation model corresponding to Figs. 2(a) and 2(b) is used.

Fig. 10 shows the experimental results when the motor under test is controlled in the current-control mode and its speed is maintained at $\omega_m = 1.5$ p.u. ($2\pi \cdot 159$ rad/s). The current references $i_{d,ref}$ and $i_{q,ref}$ are changed stepwise. Fig. 10(a) shows the result for the continuous-time design and Fig. 10(b) for the discrete-time design. The continuous-time design is unstable. In the discrete-time design, the current components follow their references. There is practically no cross-coupling and no overshoot in the control response.

If the fundamental frequency and the closed-loop bandwidth were much lower than the values used in Fig. 10, the continuous-time and discrete-time designs would give similar results. Since the proposed discrete-time design allows higher maximum speeds and the faster closed-loop dynamics than the continuous-time design, only the discrete-time design is considered in the following.

B. Zero-Speed Test

The effect of the magnetic saturation model on the control performance is studied by means of a zero-speed test. The

(a) With the constant inductances.

(b) With the saturation model.

Fig. 11. Experimental results for the discrete-time design at $\omega_m = 0$: (a) with the constant inductances; (b) with the saturation model.

Fig. 12. Experimental results showing acceleration from 0 to 2-p.u. speed for the discrete-time design with the saturation model. The first subplot shows the reference speed $\omega_{m,ref}$ and the measured speed ω_m . The second subplot shows the flux components ψ_d and ψ_q . The last subplot shows the measured current components i_d and i_q .

proposed discrete-time design of Fig. 8 is parametrized in the following two ways:

- 1) The constant inductances, corresponding to the rated operating point, are used. This parametrization corresponds to the conventional discrete-time current controller struc-

Fig. 13. Saturation models $\psi_q = \psi_q(i_q)$ at $i_d = 0.35$ p.u. with parameter errors.

ture [12].

- 2) The magnetic saturation model shown in Figs. 2(a) and 2(b) is used.

Fig. 11 shows the measurement results when the speed is maintained at zero, by locking the rotor. The current references $i_{d,ref}$ and $i_{q,ref}$ are changed stepwise. In Fig. 11(a), the constant inductances are used in the controller. Significant cross-coupling and ripple can be seen in the current components after the step changes in the current references. The response could be improved by significantly lowering the closed-loop bandwidth. Fig. 11(b) shows the result with the saturation model. The control response is much better, and the cross-coupling and ripple is very small.

C. Acceleration Test

The control scheme shown in Fig. 1 is augmented with a speed controller, which provides the torque reference based on the speed reference $\omega_{m,ref}$ and the measured speed ω_m . The optimal current references are calculated based on the torque reference and the measured DC-link voltage [20]. The motor is accelerated from zero to the speed of 2 p.u. with the maximum available torque, when the current limit is 1.5 p.u.

The continuous-time design (with or without the saturation model) and the discrete-time design with constant inductances are unstable at this high bandwidth of $\alpha = 2\pi \cdot 500$ rad/s. Fig. 12 shows experimental results for the proposed discrete-time design with the saturation model. It can be seen that the measured current components follow their references very well and there is no cross-coupling between the current components. The maximum available voltage of the inverter is used during the acceleration, but the anti-windup mechanism prevents the overshoot in the controlled current components.

D. Effects of Parameter Errors

The robustness of the proposed discrete-time controller against parameter errors is evaluated using the zero-speed test. The errors are intentionally introduced in the saturation model used in the controller. The effect of the erroneous estimate $\hat{L}_{q0}$ for the unsaturated q-axis inductance L_{q0} on the control performance is studied. Two cases are considered:

(a) $\hat{L}_{q0} = 0.5L_{q0}$.

(b) $\hat{L}_{q0} = 2L_{q0}$.

Fig. 14. Experimental results for the discrete-time design at $\omega_m = 0$ with parameter errors in the saturation model: (a) $\hat{L}_{q0} = 0.5L_{q0}$; (b) $\hat{L}_{q0} = 2L_{q0}$.

$\hat{L}_{q0} = 0.5L_{q0}$ and $\hat{L}_{q0} = 2L_{q0}$. Fig. 13 illustrates the corresponding saturation models for ψ_q as a function of i_q at $i_d = 0.35$ p.u.

In the experiments, the speed is maintained at zero by locking the rotor. Fig. 14(a) shows the results for $\hat{L}_{q0} = 0.5L_{q0}$. It can be seen that the response is slightly slower than in the nominal case shown in Fig. 11(b). This result is as expected since the underestimated inductance estimate reduces the effective gains of the controller. Fig. 14(b) shows the results for $\hat{L}_{q0} = 2L_{q0}$. As expected, the control response becomes faster than the nominal response shown in Fig. 11(b). Furthermore, some overshoots can also be seen in the measured currents, which is due to the deviation of the actual closed-loop poles from the nominal ones. Overall, it can be seen that the errors in the parameter estimates have a minor effect on the control performance. Even if not shown here, similar conclusions could be drawn for the errors in the d-axis saturation model.

VI. CONCLUSIONS

In this paper, we proposed current control structures for taking the magnetic saturation effects into account. A simple structure is achieved via a nonlinear change of a controller state variable, i.e., the stator current is mapped to the stator flux linkage using the known saturation characteristics. A discrete-time equivalent of the proposed structure is also presented. If the discrete-time effects and the saturation effects are properly taken into account, higher closed-loop bandwidths and lower sampling frequencies can be achieved. Furthermore, the controller designed directly in the discrete-time domain is more robust against parameter errors than the controller designed in

the continuous-time domain. The presented controllers have been experimentally evaluated using a 6.7-kW SyRM drive.

APPENDIX SATURATION MODEL

The saturation characteristics of the 6.7-kW SyRM are modeled as [21]

$$i_d = \left(a_{d0} + a_{dd}|\psi_d|^S + \frac{a_{dq}}{V+2}|\psi_d|^U|\psi_q|^{V+2} \right) \psi_d \quad (26)$$

$$i_q = \left(a_{q0} + a_{qq}|\psi_q|^T + \frac{a_{dq}}{U+2}|\psi_d|^{U+2}|\psi_q|^V \right) \psi_q \quad (27)$$

where the coefficients are $a_{d0} = 0.36$ p.u., $a_{dd} = 0.15$ p.u., $a_{q0} = 1.08$ p.u., $a_{qq} = 6.20$ p.u., and $a_{dq} = 2.18$ p.u. and the exponents are $S = 5$, $T = 1$, $U = 1$, and $V = 0$. The coefficient $a_{d0} = 1/L_{d0}$, where L_{d0} is the unsaturated d-axis inductance. The coefficient $a_{q0} = 1/L_{q0}$, where L_{q0} is the unsaturated q-axis inductance. The coefficients a_{dd} and a_{qq} take the self-axis saturation characteristics into account, while a_{dq} takes the cross-saturation into account. The saturation model parameters are obtained by fitting the measured data to the magnetic model (26) and (27).

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