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Optimal torque control of saturated synchronous motors: plug-and-play method

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Abstract—This paper deals with the optimal state reference calculation for synchronous motors having a magnetically salient rotor. A look-up table computation method for the maximum torque-per-ampere (MTPA) locus, maximum torque-per-volt (MTPV) limit, and field-weakening operation is presented. The proposed method can be used during the start-up of a drive, after the magnetic model identification. It is computationally efficient enough to be implemented directly in the embedded processor of the drive. When combined with an identification method for the magnetic model, the proposed method enables the plug-and-play start-up of an unknown motor. Furthermore, a conventional reference calculation scheme is improved by removing the need for one two-dimensional look-up table. A 6.7-kW synchronous reluctance motor (SyRM) drive is used for experimental validation.

Index Terms—Optimal control, synchronous motor drives, torque control.

I. INTRODUCTION

SYNCHRONOUS motors with a magnetically salient rotor—such as a synchronous reluctance motor (SyRM), a permanent-magnet (PM)-SyRM, and an interior PM synchronous motor—are well suited for hybrid or electric vehicles, heavy-duty working machines, and industrial applications [1]–[5]. These machines often exhibit highly nonlinear saturation characteristics, which should be properly taken into account in the control system in order to reach high performance. At the same time, a quick start-up procedure is desirable, especially in industrial applications.

Fig. 1 exemplifies a control system of a typical current-controlled drive. The motor is driven at the maximum torque-per-ampere (MTPA) locus, at the current limit, at the maximum torque-per-volt (MTPV) limit, or in the field-weakening region (which is the region bounded by the above-mentioned locus and limits), depending on the torque reference $T_{\text{ref}}$, the operating speed $\omega_m$, and the DC-bus voltage $u_{\text{dc}}$. This paper focuses on the calculation of the optimal state references for the fastest control loop, i.e., the current references $i_{d,\text{ref}}$ and $i_{q,\text{ref}}$ in the example control system shown in the figure.

In most existing control methods, the MTPA locus and the MTPV limit are either computed off-line based on the known magnetic saturation characteristics or measured with a suitable test bench. The resulting look-up tables are then implemented in the real-time control system. The off-line computation as well as test-bench measurements are typically time-consuming processes. Instead of using an MTPA look-up table, the MTPA locus could be tracked using signal injection [6], which, however, causes additional noise and losses. In some applications, an approximate MTPV limit could be searched for in an iterative manner by means of repeated acceleration tests [7].

Field-weakening methods can be broadly divided into feedback methods [8]–[12] and feedforward methods [13]–[20]. The feedback field-weakening methods apply the difference between the reference voltage and the maximum available voltage. These methods use the maximum voltage in the field-weakening operation, but they do not necessarily guarantee minimum losses. The voltage control loop has to be properly tuned and should have much lower bandwidth than the innermost current controller [12]. Furthermore, the MTPA locus and the MTPV limit are needed in the feedback methods as well.

The feedforward field-weakening methods provide the optimal references without delays. Due to the feedforward nature of these methods, the dynamics of the inner control loop remain intact and the noise content in the state references is minor. However, modeling inaccuracies may reduce the
available torque or increase the losses. Some feedforward field-weakening methods are based on analytical solutions of the intersection of a voltage ellipse and a torque hyperbola [13–15]. A disadvantage of these methods is that they do not take the magnetic saturation into account (and the saturation effects cannot be properly taken into account afterwards, since the saturation deforms the shape of the voltage ellipses and torque hyperbolas). Other feedforward methods are based on off-line computed look-up tables [16–20]. The magnetic saturation can be included properly in these methods, but the off-line data processing is difficult and time-consuming, even though some open-source post-processing algorithms are available [21]. Feedforward methods based on the finite-element method (FEM) rely on the knowledge of the motor geometry [16], [18]. Furthermore, the feedforward methods can be augmented with an additional voltage controller in order to be able to dynamically adjust the voltage margin [19], [20].

We consider a plug-and-play start-up method for optimal reference calculation, originally presented in the conference version [22] of this paper. As illustrated in Fig. 1, the magnetic model of the motor is first identified and the control look-up tables are then computed. These two tasks are executed off-line during the start-up, preferably in the embedded processor of the drive. After the start-up has been completed, the control look-up tables are ready to be used in the real-time control. After defining the motor model in Section II, the main contributions are presented as follows:

- A conventional reference calculation scheme [17–20], applicable to feedforward field-weakening, is improved by removing the need for one two-dimensional look-up table in Section III.
- A look-up table computation method is proposed in Section IV. When combined with an identification method for the magnetic model, the proposed method enables the plug-and-play start-up of an unknown motor. Various identification methods for the magnetic model of SyRMs, PM-SyRMs, and interior PM synchronous motors are already available [23–25].
- The effects of model uncertainties on the current loci and the achievable torque are analyzed in Section V. The proposed method is also evaluated by means of simulations and experiments.

The current-controlled drive is used as an example, but the look-up table computation method (or its part) could also be used in connection with other control methods. If, e.g., the direct-flux vector control were used, only the MTPA locus and the MTPV limit would be needed [7].

II. MOTOR MODEL

A. Fundamental Equations

The motor model in rotor coordinates is considered, as shown in Fig. 2. The stator voltage equations are

$$\frac{d\psi_d}{dt} = u_d - R_i d + \omega_m \psi_q$$  \hspace{1cm} (1a)

$$\frac{d\psi_q}{dt} = u_q - R_i q - \omega_m \psi_d$$  \hspace{1cm} (1b)

where \(i_d\) and \(i_q\) are the current components, \(\psi_d\) and \(\psi_q\) are the flux linkage components, \(u_d\) and \(u_q\) are the voltage components, \(\omega_m\) is the electrical angular speed of the rotor, and \(R \) is the stator resistance. The current components

$$i_d = i_d(\psi_d, \psi_q)$$  \hspace{1cm} \(i_q = i_q(\psi_d, \psi_q)\)  \hspace{1cm} (2)

are generally nonlinear functions of the flux components. They are the inverse of the flux maps, often represented by two-dimensional look-up tables. Here, the modeling approach (2) is chosen, because it is more favorable towards representation in the algebraic form. Since the nonlinear inductor should not generate or dissipate electrical energy, the reciprocity condition [26]

$$\frac{\partial i_d}{\partial \psi_q} = \frac{\partial i_q}{\partial \psi_d}$$  \hspace{1cm} (3)

should hold. Typically, the core losses are either omitted or modeled separately using a core-loss resistor in the model.

The produced torque is

$$T = \frac{3p}{2} (\psi_d i_q - \psi_q i_d)$$  \hspace{1cm} (4)

where \(p\) is the number of pole pairs. If the functions (2) and the stator resistance are known, the machine is fully characterized both in the steady and transient states. For example, the MTPA locus can be resolved from (2) and (4). In the following, the current magnitude will be denoted by

$$i = \sqrt{i_d^2 + i_q^2}$$  \hspace{1cm} (5)

The similar notation is used for the voltage and flux linkage magnitudes as well.

B. Algebraic Magnetic Model

The constant current source \(i_f\) in the equivalent circuit in Fig. 2 represents the magnetomotive force (MMF) of the PMs [1]. The structure of the equivalent circuit is based on the assumption that the MMFs of the d-axis current and of the PMs are in series. Under this assumption, an algebraic magnetic model [24], originally developed for SyRMs, can be extended to the PM synchronous machines as

$$i_d = \left(a_{d0} + a_{d1}|\psi_d|^\alpha + \frac{a_{d2}}{\delta + 2}|\psi_d|^\gamma|\psi_q|^\delta+2\right)\psi_d - i_f$$  \hspace{1cm} (6a)

$$i_q = \left(a_{q0} + a_{q1}|\psi_q|^\beta + \frac{a_{q2}}{\gamma + 2}|\psi_d|^\alpha|\psi_q|^\gamma\right)\psi_q$$  \hspace{1cm} (6b)

where \(a_{d0}, a_{d1}, a_{q0}, a_{q1},\) and \(a_{dq}\) are non-negative coefficients and \(\alpha, \beta, \gamma,\) and \(\delta\) are non-negative exponents. The coefficient...
\( a_{d0} \) is the inverse of the unsaturated d-axis inductance and the coefficient \( a_{q0} \) is the inverse of 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 functions (6) fulfill the reciprocity condition (3). The model is invertible: for any given \( \psi_d \) and \( \psi_q \), the corresponding values of \( i_d \) and \( i_q \) can be obtained by numerically solving (6). The magnetic model (6) corresponds to the nonlinear inductances:

\[
L_d = \frac{1}{a_{d0} + a_{dd} |\psi_d|^\alpha + \frac{a_{dq}}{\gamma + 2} |\psi_d|^\gamma |\psi_q|^\delta + 2} \quad (7a)
\]

\[
L_q = \frac{1}{a_{q0} + a_{qq} |\psi_q|^\beta + \frac{a_{dq}}{\gamma + 2} |\psi_d|^\gamma |\psi_q|^\delta} \quad (7b)
\]

For SyRMs, a standstill method for identification of the magnetic model is available [24]. This identification method needs only the measurements of the phase currents and the DC-bus voltage, while neither motion sensor nor AC-side voltage sensors are needed. If the motor is equipped with PMs, the constant \( i_t \) can be calculated based on the nameplate data, measured using the no-load test, or identified at standstill [25]. In practice, the constant \( i_t \) depends on the temperature, but the dependency is omitted here. The accuracy of the magnetic model may be insufficient for flux-switching PM machines, in which the MMFs of the d-axis current and of the PMs are essentially in parallel [27]. In the case of PM-SyRMs, the model cannot capture the desaturation phenomenon of thin iron ribs [28], appearing at positive d-axis currents, while the accuracy in the feasible operating region at negative d-axis currents is typically sufficient.

In the following, a 6.7-kW SyRM and a 7.5-kW PM-SyRM will be used as example machines. Their rated values are given in Table I and model parameters in Table II. The parameters of the magnetic model for the SyRM are obtained by fitting the model to the FEM data. Fig. 3 shows the flux linkages of the SyRM as functions of the currents, calculated from (6). It can be seen that both axes saturate significantly. The magnetic model parameters of the 7.5-kW PM-SyRM are obtained by inverting the model to the measured data. The surfaces \( \psi_d \) and \( \psi_q \) take the self-axis and cross-saturation phenomena into account.

### III. Control System

Fig. 1 depicts the overall structure of the current-controlled drive system, which is used as an example in this paper. The reference calculation is explained in the following.

#### A. Feedforward Field-Weakening Scheme

Fig. 4 shows a feedforward field-weakening scheme including the MTPA locus and the current and MTPV limits.
The maximum voltage $u_{\text{max}} = k_u u_{\text{dc}} / \sqrt{3}$ is calculated from the measured DC-link voltage $u_{\text{dc}}$. Hence, any sudden variations in $u_{\text{dc}}$ are directly translated into the references. The factor $k_u$ defines the voltage margin. As can be realized from (8), some voltage reserve is necessary for the resistive voltage drops and for changing the flux linkages in transient conditions. If $k_u = 1$ is chosen, the overmodulation region may be entered even in the steady state (due to the resistive voltage drops), which causes the sixth harmonics in the stator voltage. To avoid entering the overmodulation region in the steady state, the factor $k_u < 1$ should be chosen. Alternatively, the resistive voltage drops could be compensated for by means of the measured current and the resistance estimate, cf. e.g. [7]. However, applying this feedback action in the reference calculation may introduce increased noise content and ringing phenomena. It is also worth noticing that the factor $k_u$ could be dynamically adjusted by means of an additional voltage controller [19], [20]. For simplicity, a constant value for $k_u$ is used in this paper.

As seen in Fig. 4, the optimal MTPA flux magnitude $\psi_{\text{mtpa}}$ is read from a look-up table, whose input is the torque reference. The MTPA flux is limited based on the maximum flux $\psi_{\text{max}}$, yielding the optimal flux magnitude $\psi_{\text{ref}}$ under the voltage constraint. The torque reference $T_{\text{ref}}$ is limited by the torque $T_{\text{max}}$ corresponding to the combined MTPV and current limits, yielding the limited torque reference $T_{\text{lim}}$. An advantage of the scheme shown in Fig. 4 is that the optimal reference values are obtained without any delays. The scheme can also be used in connection with other control schemes, such as the direct-flux vector control.

**B. Current References**

If the current-controlled drive is used, the optimal flux reference and the limited torque reference have to be mapped to the corresponding current references. Fig. 5(a) shows the conventional current reference calculation method [17]–[20], based on the two two-dimensional look-up tables. Interpolation is used to get the values of $i_{d,\text{ref}}$ and $i_{q,\text{ref}}$ from the look-up tables.

Fig. 5(b) shows the proposed current reference calculation method. A single two-dimensional look-up table is used to determine $\psi_{d,\text{ref}}$. Then, the value of the q-axis flux $\psi_{q,\text{ref}}$ is obtained by means of the Pythagorean theorem. The flux-linkage references are mapped to the current references using (6). Since only one two-dimensional look-up table is needed, the memory requirements of the control system are less than in the conventional method. An interpolation procedure, applicable to the conventional method as well, is given in the Appendix.

**IV. LOOK-UP TABLE COMPUTATION**

Fig. 6 shows an overall diagram of the look-up table computation method, which is divided into four stages. In the following equations, $L_d \leq L_q$ is assumed. Further, the d-axis of the coordinate system is fixed to the direction of the PMs (or along the minimum inductance axis), without loss of generality. After the look-up table computation, the d- and q-axes of the SyRM are flipped to the standard SyRM representation, i.e., the d-axis along the maximum inductance axis. In this section, the notation and terminology is simplified such that we do not particularly refer to the reference quantities (e.g., $\psi$ is used instead of $\psi_{\text{ref}}$). It should be clear from the context how the resulting look-up tables are used in the real-time control system.
For each current magnitude \( \Delta i \) the optimization problem
\[
\psi_d, ref = \psi_d, ref \left( \psi_d, \psi_q \right) = \text{solve } \left( \begin{array}{c}
\psi_d (i_d, \psi_q) = i_d \\
\psi_q (i_d, \psi_q) = i_q
\end{array} \right)
\]
(11)

After solving (10), the optimal q-component \( i_{q,\text{mtpa}} \) is obtained from (10c).

The optimal flux magnitude is
\[
\psi_{\text{mtpa}} = \sqrt{\psi_{d,\text{mtpa}}^2 + \psi_{q,\text{mtpa}}^2}
\]
(12)

where the torque is expressed as a function of \( i_d \)
\[
T(i_d) = \frac{3p}{2} [\psi_d(i_d, \psi_q) \cdot i_q(i_d) - \psi_q(i_d, \psi_q) \cdot i_d]
\]
(10b)
\[
i_q(i_d) = \sqrt{i^2 - i_d^2}
\]
(10c)

and the search interval is \(-i \leq i_d \leq 0\). The flux components \( \psi_d \) and \( \psi_q \) corresponding to \( i_d \) and \( i_q \) are calculated by numerically inverting the algebraic magnetic model (6), i.e.,

As shown in Fig. 6, the procedure (10)–(12) is repeated in a for loop for each element \( i(l) \) of the list (9). Then, a look-up table for the control system, cf. Fig. 4, is created from the resulting lists \( \{\psi_{\text{mtpa}}(l)\} \) and \( \{T_{\text{mtpa}}(l)\} \). As illustrated in Fig. 6, the inputs to the MTPA computation stage are the number of points \( L \) to be computed, the maximum current \( i_{\text{max}} \), and the parameters of the magnetic model (6). MTPA loci are generally comparatively smooth and, typically, \( L \)
around 10–20 suffices. The maximum current $i_{\text{max}}$ is selected based on the motor and converter ratings.

Fig. 7(a) shows the computed MTPA look-up table for the 6.7-kW SyRM and Fig. 7(b) for the 7.5-kW PM-SyRM. In the control algorithm, the optimal flux reference magnitude $\psi_{\text{ref}}$ is obtained based on this look-up table, as shown in Fig. 4.

**B. MTPV**

For creating the look-up table, a list of $M$ equally-spaced stator flux magnitudes is defined

$$\{\psi(m)\} = (m - 1)\Delta\psi, \quad m = 1, 2, \ldots M \quad (13)$$

where $\Delta\psi = \psi_{\text{mtpa}}(L)/(M - 1)$, i.e., the maximum flux magnitude $\psi_{\text{mtpa}}(L)$ is the result from the last step of the MTPA computation. For each flux magnitude $\psi$, the maximum torque $T_{\text{mtpv}}$ is obtained by solving

$$T_{\text{mtpv}} = \max_{\psi \in [0,\psi]} T(\psi_d) \quad (14a)$$

where the torque is expressed as

$$T(\psi_d) = \frac{3p}{2} [\psi_d \cdot i_q(\psi_d, \psi_q) - \psi_q(\psi_d) \cdot i_d(\psi_d, \psi_q)] \quad (14b)$$

$$\psi_q(\psi_d) = \sqrt{\psi^2 - \psi_d^2} \quad (14c)$$

The magnetic model (6) is directly used in (14b), i.e., no magnetic model inversion is needed in this stage. The Brent algorithm is used for solving the optimization problem (14).

As shown in Fig. 6, the problem (14) is solved for each element $\psi(m)$ of the list (13). Then, the look-up table for the control system is created using the resulting output list $\{T_{\text{mtpv}}(m)\}$. Fig. 7 shows the computed MTPV look-up tables for the two machines.

**C. Maximum Current Limit**

The already defined input list (13) of the flux magnitudes is considered. For each flux magnitude $\psi(m)$, the d-component $\psi_{d,\text{lim}}$ of the flux corresponding to the maximum current $i_{\text{max}}$ is solved

$$\psi_{d,\text{lim}} = \max_{\psi_d \in [\psi_{d,\text{mtpv}},\psi_{d,\max}]} \{i^2(\psi_d) = i_{\text{max}}^2\} \quad (15a)$$

where the square of the current magnitude is expressed as

$$i^2(\psi_d) = i_d^2(\psi_d, \psi_q) + i_q^2(\psi_d, \psi_q) \quad (15b)$$

$$\psi_q(\psi_d) = \sqrt{\psi^2 - \psi_d^2} \quad (15c)$$

The lower bound $\psi_{d,\text{mtpv}}$ in (15) is the d-component of the MTPV flux at each $\psi$ and the upper bound $\psi_{d,\max}$ is the d-component of the MTPA flux at the maximum current. After (15) has been solved, the corresponding torque $T_{\text{lim}}$ is obtained from (14b). The Brent algorithm is used to solve this bounded nonlinear problem.

As shown in Fig. 6, the problem (15) is solved for each element $\psi(m)$. The lower bounds $\{\psi_{d,\text{mtpv}}(m)\}$ needed in (15) have already been computed during the MTPV stage. The upper bound $\psi_{d,\max} = \psi_{d,\text{mtpa}}(L)$ is the d-component of the MTPA flux at the maximum current $i_{\text{max}}$ and it has also been computed. From the resulting output list $\{T_{\text{lim}}(m)\}$, a look-up table for the control system is created. Fig. 7 shows the computed current limits of two times the rated current for the SyRM and the PM-SyRM. The MTPV and current limits can be easily merged into one limit

$$T_{\text{max}} = \min (T_{\text{mtpv}}, T_{\text{lim}}) \quad (16)$$

**D. Two-Dimensional Reference Look-Up Table**

For given flux magnitude $\psi$ and torque reference $T_{\text{ref}}$, the d-component $\psi_{d,\text{ref}}$ is solved

$$\psi_{d,\text{ref}} = \max_{\psi_d \in [\psi_{d,\text{mtpv}}, \psi]} \{T_{\text{ref}} = T(\psi_d)\} \quad (17)$$

where the torque $T(\psi_d)$ is given by (14b). The lower bound $\psi_{d,\text{mtpv}}$ is the d-component of the MTPV flux at $\psi$. The Brent algorithm is used to solve (17).

For creating the look-up table, (17) can be solved in two nested loops. As an input to one loop, the list (13) of the flux magnitudes $\{\psi(m)\}$ is used. In the other loop, the already calculated MTPV torque values $\{T_{\text{mtpv}}(m)\}$ can be used as an input, i.e. $\{T_{\text{ref}}(m)\} = \{T_{\text{mtpv}}(m)\}$. This selection not only defines the maximum torque which can be generated (under the MTPV limit) but also explicitly gives the lower bound $\psi_{d,\text{mtpv}}$ for each $\psi_{d,\text{ref}}$. If the maximum current is fixed, $\{T_{\text{ref}}(n)\} = \{T_{\text{max}}(m)\}$ can be used instead. The look-up table for the control system is created from the resulting table $\{\psi_{d,\text{ref}}(m, n)\}$. Fig. 8 shows the two-dimensional look-up tables for the two machines. The $i_d = 0$ line shown in Fig.
Fig. 9. MTPA locus, MTPV limit, and current limit for the 6.7-kW SyRM: (a) \textit{i}_d-i_q plane; (b) \textit{\psi}_d-\textit{\psi}_q plane. The dashed lines show the loci, when the magnetic saturation is not taken into account (inductances correspond to the rated operating point). The black dashed line corresponds to the constant current circle.

Fig. 10. MTPA locus, MTPV limit, and current limit for the 7.5-kW PM-SyRM: (a) \textit{i}_d-i_q plane; (b) \textit{\psi}_d-\textit{\psi}_q plane. The dashed lines show the loci, when the magnetic saturation is not taken into account (inductances correspond to the rated operating point). The black dashed line corresponds to the constant current circle.

8(b) does not need to be computed, but it is shown just for illustration purposes.

E. Generalization for Machines With \textit{L}_d > \textit{L}_q

Some special PM machines can have \textit{L}_d > \textit{L}_q [31]. The presented algorithm can be easily modified to include these machines as well. Only a few modifications in the ranges are needed: \textit{i}_d \in [-\textit{i}, \textit{i}] in (10); \textit{\psi}_d \in [-\textit{\psi}, \textit{\psi}] in (14); and \textit{\psi}_d \in [-\textit{\psi}_{d,\text{MTPV}}, \textit{\psi}] in (15). In this general case, the computation time of the algorithm will increase.

V. RESULTS

The computed look-up tables and the reference calculation scheme corresponding to Figs. 4 and 5(b) are evaluated by means of analysis, simulations, and experiments. The parameters used for the look-up table computation are: \( L = 10 \); \( M = 150 \); and \( i_{\text{max}} = 2 \) p.u. The parameters of the magnetic model (6) given in Table II are also needed. The computation time for the look-up tables is less than 35 s in an Android mobile phone. We expect that the computation time is slightly longer in a typical digital-signal processor applied in frequency converters.

A. Analysis

The effect of the magnetic saturation on the optimal references is analyzed. Fig. 9 shows the current references in the \textit{i}_d-\textit{i}_q plane and the flux references in the \textit{\psi}_d-\textit{\psi}_q plane for the 6.7-kW SyRM. The solid lines correspond to the computed optimal values, while the magnetic saturation is omitted in the case of the dashed lines. The effect of the saturation is clearly visible: using constant inductances would result in non-optimal operating points. Fig. 10 shows the corresponding loci for the 7.5-kW PM-SyRM. The saturation affects mainly the MTPA locus in this case.

Fig. 11(a) shows the effect of the magnetic saturation on the maximum torque production for the 6.7-kW SyRM. It can be seen that the torque production capability drops significantly, if the magnetic saturation is omitted in the reference calculation and the inductances in the rated operating point are used instead. Furthermore, the cross-saturation also has a significant effect on the torque production. This result indicates that the cross-saturation effect should be included in the magnetic model in the case of SyRMs. Fig. 11(b) shows the maximum torque versus speed characteristics for different current limits. If needed, the reference calculation scheme in
Fig. 12. Simulation results showing acceleration from zero to 2 p.u. with two different voltage margins: (a) $k_u = 0.8$; (b) $k_u = 1$. The first subplot shows the reference speed $\omega_{m,\text{ref}}$ and the actual speed $\omega_m$. The second subplot shows the reference torque $T_{\text{ref}}$ and the estimated torque $\hat{T}$. The third subplot shows the reference and measured current components. The last subplot shows the magnitude $u_{\text{ref}}$ of the reference voltage, the maximum available voltage $u_{\text{dc}}/\sqrt{3}$ in the linear modulation range, and the maximum voltage $u_{\text{max}} = k_u u_{\text{dc}}/\sqrt{3}$ in the steady state.

Fig. 4 can be modified easily to use multiple current limits or even a dynamic current limit, as needed in some industrial applications. Three to four one-dimensional look-up tables could be used for different current limits and then the results between these limits could be interpolated as required.

The effect of the core losses on the torque production capability is analyzed for the 6.7-kW SyRM. For this analysis, the constant core-loss resistance of 18 p.u., corresponding to [32], has been included in the motor model. Since the core losses are omitted in the proposed reference calculation method, the achieved maximum torque is 6% less than the real maximum torque at the speed of 1.25 p.u. The loss of the achievable maximum torque increases with the speed. Effects of the core losses on the efficiency have been analyzed in [19], [32].

**B. Simulations**

Simulations and experiments were performed on the 6.7-kW SyRM drive. The magnetic saturation in the motor model and the controller is modeled using the magnetic model in (6). The load torque is modeled as the viscous friction. The total moment of inertia is 0.03 kgm$^2$. The space-vector pulse-width modulator and the discrete-time current controller [33] are used. The sampling and switching frequency are 5 kHz. The speed controller bandwidth is 10 Hz.

Fig. 12(a) shows the acceleration test when the voltage margin is defined by $k_u = 0.8$. The motor is accelerated from zero to the speed of 2 p.u. It can be seen that the measured currents follow their references well. The magnitude $u_{\text{ref}}$ of the voltage reference, the limit $u_{\text{dc}}/\sqrt{3}$ of the linear-modulation range, and the desired maximum steady-state voltage $u_{\text{max}} = k_u u_{\text{dc}}/\sqrt{3}$ are also shown. It can be seen that the overmodulation range is not entered in this case.

Fig. 12(b) shows the same test with $k_u = 1$. It can be seen that the overmodulation range is mostly used during the acceleration. Furthermore, the current $i_q$ deviates from its reference during the acceleration. This behavior is expected since only part of the overmodulation voltage reserve is available for changing the currents in transient operation, cf. (8). However, due to the higher value of the stator flux linkage with $k_u = 1$, the acceleration time is shorter than with $k_u = 0.8$.

A suitable choice for the voltage margin factor $k_u$ depends on the application. In these examples, the speed reference was changed stepwise. However, the speed reference signal is typically ramped, which decreases the need for the voltage margin. As discussed in Section III, the feedforward field-weakening method could be augmented with a voltage controller providing a dynamic voltage margin, which allows combining the maximum utilization of the DC-bus voltage in the steady state with proper tracking of the current references in transients.
Fig. 13. Experimental results showing acceleration from zero to 2 p.u. with two different voltage margins: (a) $k_u = 0.8$; (b) $k_u = 1$.

Fig. 14. Experimental results showing torque reference steps at two different speeds: (a) $\omega_m = 0.6$ p.u.; (b) $\omega_m = 1.2$ p.u.

C. Experiments

The reference calculation scheme corresponding to Figs. 4 and 5(b) was experimentally evaluated together with the computed look-up tables. The controller was implemented on a dSPACE DS1006 processor board. The rotor speed $\omega_m$ is measured using an incremental encoder. The stator currents and the DC-link voltage are measured.

Fig. 13 shows the experimental results corresponding to the simulation results in Fig. 12. It can be seen that the experimental results and the simulation results match very well. There is a slight overshoot in the speed response at the end of the acceleration due to imperfect tuning of the speed controller, further causing the torque reference to change its sign momentarily.

Fig. 14 shows the results for the constant speed tests at two different speeds. The load drive of the test bench regulates the speed and the drive under test is driven in the torque-control mode. Fig. 14(a) shows the results when the speed is regulated at 0.6 p.u. The torque reference is stepped from zero to 150% of the rated torque with increments of 25%. Fig. 14(b) shows the results when the speed is regulated at 1.2 p.u. and the torque reference is stepped from zero to the rated torque with increments of 25%. The current dynamics in this kind of constant speed tests are governed solely by the closed-loop
current control dynamics, as can be realized from Figs. 1, 4, and 5(b). The reference calculation scheme maps the torque reference to the optimal current references in a feedforward manner without any dynamics.

VI. CONCLUSIONS

A look-up table computation algorithm for the MTPA locus, MTPV limit, and feedforward field-weakening operation is presented. The algorithm can be used during the drive start-up, after the magnetic model identification. It is computationally efficient enough to be implemented directly in the embedded processor of the drive. Alternatively, the look-up tables could be computed remotely using the algorithm and then uploaded to the drive, if the drive is connected to a cloud server or to a mobile phone. A conventional real-time reference calculation scheme was also improved by removing the need for one two-dimensional look-up table. The importance of including the magnetic saturation in the look-up table computation was highlighted. Using constant inductances would result in non-optimal operating points, as the saturation deforms the voltage ellipses and torque hyperbolas. The proposed method properly takes the magnetic saturation into account. The computed look-up tables and the reference calculation scheme were evaluated using experiments on a 6.7-kW SyRM drive.

APPENDIX
INTERPOLATION

Generally, four points are needed for the interpolation in two dimensions. However, no data is available for the two-dimensional look-up table beyond the MTPV limit. Hence, when operating in the vicinity of the MTPV limit, only three points are available for the interpolation algorithm. Therefore, there are two different modes in the interpolation algorithm, as described in the following.

Suppose that the value of a function \( f \) is available at four points \((x_1, y_1), (x_1, y_2), (x_2, y_1), \) and \((x_2, y_2)\) shown in Fig. 15. The value of the function at \((x, y)\) can be approximated using bilinear interpolation

\[
f(x, y) = \frac{[x_2 - x] \begin{bmatrix} x_1 & y_1 & 1 \end{bmatrix} \begin{bmatrix} f(x_1, y_1) & f(x_1, y_2) & y_2 - y \end{bmatrix} - [x - x_1] \begin{bmatrix} y_1 & y_2 & 1 \end{bmatrix} \begin{bmatrix} f(x_1, y_1) & f(x_2, y_1) & y_2 - y_1 \end{bmatrix}}{(x_2 - x_1)(y_2 - y_1)} \tag{18}
\]

If the value at one point, e.g. at \((x_2, y_1)\), is not available, the values at the remaining three points can be used for approximating the value at \((x, y)\) by means of a plane equation

\[
f(x, y) = ax + by + c \tag{19}
\]

where

\[
\begin{bmatrix} a \\ b \\ c \end{bmatrix} = \begin{bmatrix} x_1 & y_1 & 1 \\ x_1 & y_2 & 1 \\ x_2 & y_1 & 1 \end{bmatrix}^{-1} \begin{bmatrix} f(x_1, y_1) \\ f(x_1, y_2) \\ f(x_2, y_2) \end{bmatrix} \tag{20}
\]

For interpolating \(\psi_d\) at the point \((\psi, T)\), the four points surrounding \((\psi, T)\) are read from the look-up table and (18) is then applied. If only three points are available due to the vicinity of the MTPV limit, (19) is applied instead of (18). The value of the q-axis flux component \(\psi_q\) could be calculated from the final interpolated result \(\psi_{d1}\) using the Pythagorean theorem, but it might create chattering in the calculated reference. Instead, the q-axis flux component is first calculated using the Pythagorean theorem at all the three or four points used in the interpolation algorithm for \(\psi_{d1}\). To get the interpolated \(\psi_{q1}\), either (18) or (19) is used, depending on the number of the points.

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REFERENCES


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