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Model-based flux weakening strategy for synchronous machines without additional regulators

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Abstract: This study proposes a model-based control strategy for flux weakening operation of a synchronous reluctance machine, with a methodology that is extendible to any synchronous machine. The strategy leverages in the presence of digital non-linear models that describe the relation between currents and flux linkages in the machine. Such models are usually needed for conventional maximum-torque-per-ampere control and sensorless control, but here they are exploited to achieve flux-weakening operation without the need of flux weakening regulators, ensuring a seamless transition between the operating regions of the machine. The external voltage regulation loop for flux weakening is thus eliminated and substituted by a combination of look-up tables and binary searches, which are executed within one digital control period and which generate the required current and voltage references that fulfill the drive limitations. The method can also be coupled with mechanisms to compensate for magnetic parameter inaccuracies, to achieve an accurate tracking of the reference torque. The proposed solution is simulated and validated in a laboratory test bench on an 11 kW synchronous reluctance machine.

1 Introduction

1.1 Background

Pulled by the increasing demand of efficiency, power density, and dynamic performances, the last decade has witnessed a steady increase in the exploitation of drives for synchronous machines for applications ranging from industry and automotive to renewable energy conversion. While drive solutions for surface permanent magnet synchronous machines (SPMSMs) are well established, some optimisation is still required for interior permanent magnet synchronous machines (IPMSMs), permanent magnet-assisted synchronous reluctance machines (PMA SynRMs) and synchronous reluctance machines (SynRMs), where the intrinsic non-linear magnetic properties of the machines and the availability of reluctance torque must be handled correctly to achieve the best efficiency and correct flux weakening operation.

A lot of problems can actually be solved with advanced (and possibly automatic) drive commissioning procedures which are able to determine the non-linear magnetic characteristics of the machine, in particular considering the effects of magnetic saturation and cross-magnetisation in the relation between the currents and the flux linkages. Several examples of these procedures are available in the literature, among others [1–3] are specifically devoted to the non-linear characteristics of synchronous machines. The results of such procedures are of foremost importance for the tuning of the regulators and the selection of the correct current references for minimal loss operation and flux weakening [4].

However, having good confidence in the knowledge of machine parameters does not lead immediately to optimised drive performances. Optimal algorithms that use the parameters in a correct way must follow. Among others, a class of problems with lively R&D activities relates to the optimal flux weakening strategies for IPMSMs, PMA SynRMs, and SynRMs. Interesting comparisons are available, as in [5] where field-oriented control (FOC) and direct torque control (DTC) with their related flux-weakening strategies are analysed. Two types of FOC strategies are evaluated: one based on an external voltage loop, and another based on a motor model and explicit expressions for the $d$-axis current. The analysis concludes, among other things, that the model-based flux weakening is parameter-dependent, and therefore more sensitive to parameter variations. Indeed, the previous work [6], with focus on a model-based flux weakening, indirectly reached similar conclusions by proposing the inclusion of the stator resistance and the mapping of magnetic saturation for the inductances in the current reference expressions in flux weakening.

Improvements to DTC-based flux weakening algorithms, to further improve their robustness, are still ongoing as shown in recent works [7, 8]. On the FOC side, although more prone to parameter sensitivity errors, the model-based flux weakening strategies are gaining interest because the schemes with an external voltage loop introduce a further regulator to tune, and the related dynamics that must not interfere with the current control loop. For this reason, interesting alternative attempts to re-define the control reference frame for achieving a seamless flux-weakening operation without regulators have been published [9–12]. A recent approach based on a predictive stator flux and load angle control is also available in [13].

1.2 Contribution

In this study, a model-based flux weakening approach for FOC is analysed. The algorithm exploits the knowledge of the non-linear relation between current and flux linkages. Such a relation could be obtained either by automatic commissioning procedures or through calibrated finite-element analysis and they come in the form of look-up tables (LUTs).

The maximum-torque-per-ampere (MTPA) operation and the flux weakening operation are seamlessly integrated by means of nested binary searches, which are executed within one control cycle. The binary searches are activated depending on the fulfillment of the drive limitations on nominal current and available voltage. This approach ensures that at each control cycle the current and voltage references fulfill the constraints without the need of any external voltage loop for flux weakening.

The algorithm is also compatible with methods for a real-time compensation of the magnetic state variables (in particular, of the flux linkages), in order to reduce the parameter sensitivity of the
model-based solution. The approach inspired by [14] and further discussed in [15] is used as an example for the sensitivity evaluation.

The paper is divided as follows. Section 2 describes the basics of the proposed control approach, highlighting the interaction between its different blocks. Section 3 reports the experimental results obtained on an 11 kW SynRM prototype, discussing the behaviour of the proposed control approach under different operating conditions. Section 4 includes some final remarks, followed by the references.

2 Theory of the proposed approach

2.1 Overview of the algorithm

The proposed strategy shown in Fig. 1 modifies the conventional FOC for a synchronous machine (not necessarily a SynRM), by introducing additional elements in the conventional current proportional–integral (PI) regulation loop. The speed regulation loop is left untouched and will not be further discussed. A space vector notation in bold font is used for currents, voltages and flux linkages.

Focusing on the current regulation, the torque reference \( \tau^* \) (optionally adjusted for parameter mismatches with the method of Section 2.4) and the mechanical speed \( \omega_m \) are provided to a reference generator described in Section 2.5. The reference generator relies on a feed-forward machine model discussed in Section 2.2.

The reference generator provides current and voltage references in the \( dq \) reference frame, named \( i_{dq} \) and \( u_{dq}^* \). In case of no parameter mismatch, \( u_{dq}^* \) is the voltage required to obtain the desired \( i_{dq} \) and thus \( \tau^* \). Therefore, \( u_{dq}^* \) is considered as a feed-forward value to the space vector modulation.

In the presence of parameter mismatches, the sole voltage generation of \( u_{dq}^* \) will not produce the desired \( i_{dq} \). Therefore, current PI controllers in parallel to the feed-forward voltage \( u_{dq}^* \) are employed as a means of compensation. The output of the PI controllers, named \( c_{dq} \), ensures a zero steady-state error in the \( dq \) current regulation and it is optionally further used to compensate for parameter mismatches, as described in Sections 2.3 and 2.4.

Section 2.5 also describes the seamless flux weakening operation without external voltage loop performed by the reference generator.

2.2 Feed-forward machine model

The machine model contained in the reference generator of Fig. 1 is the key element for the seamless flux weakening operation and, optionally, for the parameter mismatch detection. The model is the digital implementation of the known voltage equations of a synchronous machine in the space vector notation

\[
\begin{align*}
\dot{u}_{dq} &= R_{S}i_{dq} + \frac{d}{dt}\lambda_{dq} + j\omega_{me}\lambda_{dp} \\
\dot{\lambda}_{dq} &= f(i_{dq}, \lambda_{pm})
\end{align*}
\]  

where \( u_{dq}, i_{dq} \) and \( \lambda_{dq} \) are the space vector voltages, currents, and flux linkages, \( R_s \) is the stator resistance, and \( \omega_{me} \) is the electromechanical speed related to the mechanical speed \( \omega_m \) by the number of pole pairs \( p \). The vector function \( f \) maps the \( dq \) currents and the flux linkage of the magnets \( \lambda_{pm} \) (if present) to the \( dq \) flux linkages. Note that the flux of the magnets is a space vector, typically aligned on the \( d \)-axis for SPMSMs and IPMSMs, but possibly aligned on the \( q \)-axis for PMaSynRMs, depending on the chosen \( dq \) frame definition. SynRMs have zero \( \lambda_{pm} \).

The non-linear function \( f \) accounts for the magnetic relation between currents and flux linkages, in particular, the magnetic saturation and cross-saturation effects. For a more detailed discussion of the theoretical foundations of the function \( f \), the reader is redirected to [3].

The machine model (1) generates the feed-forward voltage as shown in Fig. 2, where the symbol \( \hat{\cdot} \) denotes estimated quantities that are not directly measured. In particular, the output of the LUT implementing the function \( f \) is an estimated flux linkage \( \hat{\lambda}_{dq} \), thus not necessarily corresponding to the actual flux linkage \( \lambda_{dq} \). As discussed in Section 2.4, the value of \( \hat{\lambda}_{dq} \) is optionally brought to match the actual flux linkage value \( \lambda_{dq} \) by means of the contribution \( \Delta\lambda_{dq} \).

Furthermore, it is assumed that the stator resistance value can be estimated to track its changes due to temperature variations. It is suggested to use injection methods as in [16] to avoid a bidirectional dependence on the stator resistance estimation and magnetic parameter estimation.

A final remark on the derivative operation is shown in Fig. 2. Due to the digital nature of the control, the implementation of the derivative operator must be carefully performed to avoid noisy signals. Typical implementations that include a derivative operation followed by low-pass filtering (in this work, a cut-off frequency of 50 Hz was used) are an established way to circumvent the problem and are not further discussed in this study.
2.4 Optional flux and torque compensator

Happen only if the machine model is severely wrong. It was already mentioned in Section 2.1 that the flux linkage and therefore, the flux estimation error follows the methods [14, 15], to which the reader is redirected for a more detailed theoretical analysis.

Although not shown in Fig. 3, a conventional anti-windup algorithm was implemented in the PI controllers. Since the anti-windup could be activated only during transients, it does not influence the steady-state information on the magnetic mismatches. They are only possibly activated when the current PI regulators are asked to produce a significant amount of voltage, which may happen only if the machine model is severely wrong.

2.5 Reference generator

The complexity of the proposed method resides in the reference generator. A flow chart diagram of its operation is shown Fig. 5. The reference \( r^* \) is sent to a first LUT which provides a provisional d-axis current reference \( i^*_d \), chosen according to the MTPA principle. The provisional \( i^*_d \) is sent to a second LUT which accepts \( r^* \) as a second input. The output is a provisional q-axis current reference \( i^*_q \). The second LUT is not limited only to the MTPA trajectory but describes the complete set of torque curves reversed to obtain the q-axis current in output. Note that the second LUT is a mathematical manipulation of the function \( f \) used in the machine model of Section 2.2 since the torque is obtained as a cross-product between the currents and the flux linkages. In this perspective, a precise commissioning of the function \( f \) ensures an accurate calculation of the second LUT and of the first MTPA LUT as a special current trajectory case.

Both provisional current references, as well as \( o_{me} \) and \( \Delta \lambda_{dq} \), are sent to the machine model of Fig. 2. Before using the generated

\[
\Delta \lambda_{dq,i} \text{ is controlled to zero by a PI regulator, generating the flux linkage compensation } \Delta \lambda_{dq} \text{ of Figs. 1 and 2. It is worth noting that, being valid at steady-state, the term } \Delta \lambda_{dq} \text{ contributes to generating the voltage reference with a bandwidth that should be tuned to be lower than the current loop one, so that the two loops will not unnecessarily interfere.}
\]

The dynamics of the current PI regulator is meant to prevail without being affected. In rough terms, considering a constant speed (the calculation can be replicated for different speeds), the open loop transfer function from \( \Delta \lambda_{dq,i} \) to \( \lambda_{dq} \) should have a gain crossover frequency not higher than the bandwidth of the current PI regulation.

The division by \( o_{me} \) in (3) requires a minimum allowed speed to avoid a division by zero.

Experimental observations showed that a lower limit of few rad/s for \( o_{me} \) returns satisfactory results. Nevertheless, due to this limitation, a correct flux linkage compensation at standstill is not possible.

The discontinuities in the signal \( \Delta \lambda_{dq} \) introduced by the speed limitation are eliminated using the low-pass filter \( G_f(s) \). Its cut-off frequency can be set slightly larger than the flux compensator loop bandwidth, without significantly altering its dynamics.

Along with the flux linkage compensation, a torque compensation is calculated by means of the following equation:

\[
\Delta \tau = \frac{3}{2} \tau \Delta \lambda_{dq} \times i_{dq}. \quad (4)
\]

The calculation of the torque compensation is required to avoid a mismatch between the torque reference \( \tau^* \) produced by the speed regulator and the actual torque produced by the machine. This torque compensation is zero when the flux linkage compensation term is equal to zero.

Fig. 3 Current PI controllers with compensating terms

Fig. 4 Flux and torque compensate

Fig. 5 Flow chart diagram of the reference generator
voltage reference (optionally compensated for parameter mismatches by \(\Delta i_d\)) as a feed-forward, its magnitude is immediately checked against the maximum allowed voltage \(U_{\text{lim}}\).

If the voltage limitation condition is not met, the nested binary search named ‘voltage binary search’ in Fig. 5 is activated to find a new current reference that would meet the voltage requirements. A new provisional value of \(i_d\) is calculated instead of the MTPA one, and provided to the second LUT for the generation of a new value of \(U_d\). The two new current references are used to recalculate the provisional voltage reference magnitude, which is checked again against \(U_{\text{lim}}\). The whole binary search process, which is executed within one control cycle, continues until the calculated current references lead to a voltage reference magnitude that fulfils the \(U_{\text{lim}}\) constraint.

The nature of the reference \(i_d\) obtained by the binary search depends on the machine type. In the case of a SynRM, the new \(i_d\) is found in the current range between zero and the MTPA value \(i_d\text{MTPA}\). The binary search successively divides the initial current range into smaller halves in order to find the reference that guarantees the voltage condition. The algorithm stops when the current interval length at the \(n\)th iteration is smaller than an imposed value \(\delta i_d\).

Once the voltage limitation is satisfied, however, another binary search for the validation of the current limit \(I_{\text{lim}}\) must be performed, as a consequence of the potential flux linkage mismatch. This is caused by the potential mismatch between \(r^{*}\) and the actual torque when the LUTs are not describing the correct torque curves. Although the compensating term \(\Delta r\) ensures a correct torque tracking, the potential current references might be outside the nominal current limitation. Therefore, the nested binary search on the current limitation ensures that the torque reference \(r^{*}\) guarantees the fulfilment of the current limit.

Both nested binary searches are executed within the same control cycle, thus avoiding the need of any external voltage regulation loop and providing current and voltage references with instantaneous flux weakening capabilities. Their execution times is linked to the maximum number of search iterations, which in the case of the voltage binary search is obtained with

\[
N_v = \left[\log_2 \frac{I_{dMTPA}}{\Delta i_d}\right],
\]

where \(I_{dMTPA}\) is the maximum \(d\)-axis reference current and \(\delta i_d\) is the maximum allowed current error during the voltage binary search. A similar formula can be used for the maximum number of iterations \(N_i\) of the current binary search, with \(r^{*}\) as the maximum torque reference and \(\delta r\) as the maximum torque error during the current binary search. The total maximum number of iterations of the nested searches is equal to \(N_vN_i\), which should be considered as a worst-case scenario when estimating the total execution time of the proposed algorithm.

It is anticipated that in the SynRM experimental setup described in Section 3.1, the nested binary searches and the rest of the control were well executed within every control cycle. The selected parameters were \(\delta i_d = 0.01\) A and \(\delta r = 0.01\) Nm, while \(I_{dMTPA}\) and \(r^{*}\) were considered equal to the nominal motor current and torque, respectively. The resulting total maximum number of iterations is 121.

### 3 Experimental validation

#### 3.1 Setup

The experimental test bench is shown in Fig. 6. The 11kW SynRM prototype under test, whose parameters are reported in Table 1, is shown on the right side. The loading machine on the left side (an 11 kW Baldor machine) is back-to-back connected with the SynRM, with a torque meter in between them. The loading machine is controlled via an off-the-shelf ABB ACS850 converter. The SynRM, instead, is connected only to the power unit of an ABB ACS850, while its control board has been replaced by a custom interface and connected to an OPAL-RT Technologies OP5600 system (the black box on the upper-left side of the figure). The phase currents and the direct current (DC)-bus voltage are measured with a custom measurement box and connected to an OP5600 A/D board. The OP5600 digital I/Os communicate to the ACS850 power unit through the custom interface.

The OP5600 is equipped with a quad-core Intel DSP processor at 2.4 GHz (only one core was active for the control) and a Virtex 6 field programmable gate array (FPGA). The selected pulse-width modulation (PWM) switching frequency is 8 kHz, thus allowing 125\(\mu\)s for the control algorithm execution, which was found to be ~18.5\(\mu\)s during operation in the worst case. A conventional FOC with MTPA reference selection and without flux weakening functionality was approximately executed in 17\(\mu\)s in the worst case. A compensation of the PWM delay is always active to reduce the lag of electrical quantities.

#### 3.2 Measurements with correct magnetic model

Figs. 7 and 8 show the drive control behaviour during speed and load torque transients when the correct magnetic model is used in the LUT of Fig. 2. Initially, a speed reference step from 500 to 1000 rpm is applied in the presence of a load torque \(T_N\) equal to the 23.5% of the nominal torque \(T_N\). Then, after the steady state is reached, \(T_N\) is changed to 82.5% of \(T_N\) with a step. The flux weakening operation was forced by reducing the maximum allowed voltage \(U_{\text{lim}}\) to 90 V.

At the beginning of the test, both voltage and current magnitudes are well below their limits \(U_{\text{lim}}\) and \(I_{\text{lim}}\) as shown in Figs. 7c and 7d, respectively.

During the speed transient, the currents and voltages rise as expected. When the voltage limitation is hit because of the load step change (Fig. 7c), the currents are modified to enter flux weakening operation. Both \(i_d\) and \(i_q\) are correctly reduced to

![Fig. 6 Experimental setup](image)

<table>
<thead>
<tr>
<th>Table 1 11 kW SynRM parameters</th>
<th>Parameter value</th>
</tr>
</thead>
<tbody>
<tr>
<td>nominal voltage</td>
<td>400 V</td>
</tr>
<tr>
<td>nominal current</td>
<td>18 A</td>
</tr>
<tr>
<td>nominal speed</td>
<td>6000 rpm</td>
</tr>
<tr>
<td>nominal torque</td>
<td>17 Nm</td>
</tr>
<tr>
<td>pole pairs</td>
<td>2</td>
</tr>
<tr>
<td>DC bus voltage</td>
<td>563 V</td>
</tr>
<tr>
<td>stator resistance</td>
<td>0.72 Ω</td>
</tr>
<tr>
<td>inertia (SynRM)</td>
<td>0.00351 kgm²</td>
</tr>
<tr>
<td>inertia (complete back-to-back connection)</td>
<td>0.034 kgm²</td>
</tr>
</tbody>
</table>
weaken the magnetic field blue and fulfil the voltage limitation (Figs. 8a and b), while \( \dot{i}_q \) is increased (Fig. 8c).

Since a correct magnetic model is used, the contribution of \( \Delta \lambda_{d}q \) is negligible \( \sim 3\% \) of \( \lambda_{d}q \), see Figs. 8b and d, and so for \( \Delta \tau \) in Fig. 7b – and solely related to digital control imperfections, like an imperfect compensation of PWM delays and insulated-gate bipolar transistor (IGBT) dead times.

### 3.3 Measurements with magnetic model mismatch

A second test with the same speed and load torque transients used in Section 3.2 was performed in the case of a magnetic model mismatch. The results are shown in Figs. 9 and 10.

In this test, the LUTs used by the feed-forward machine model and the reference generator were grossly modified, by removing the cross-saturation effect and by overestimating the flux linkage curves by 10% with respect to the largest curves of \( \lambda_{d} \) and \( \lambda_{q} \) as a function of the currents. The two curves used in the mismatched function are shown in Fig. 11 as dashed lines, along with the original SynRM curves used for the previous test (only three curves of the cross-saturation effect are shown).

Similar to the results of Section 3.2, the drive copes correctly with the voltage and current limitations. However, in this case, the contribution of \( \Delta \lambda_{d}q \) shown in Fig. 10 is in the range of 20% of the actual \( \lambda_{d}q \), actively contributing to the correct estimation of \( \lambda_{d}q \). As a result, the values of \( \lambda_{d}q \) shown in Fig. 10b match the ones obtained in Fig. 8b where the correct magnetic model was used.

The torque compensation in Fig. 9b is also in the order of 30% of the actual torque, bringing the torque reference \( \tau^* \) to the same level experienced in Fig. 7b during the previous experiment. This

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**Fig. 7** Drive operation with correct magnetic model
(a) Mechanical speed, (b) Torque references and compensation, (c) Voltage reference magnitude and voltage limitation, (d) Current magnitude, reference magnitude and current limitation.

**Fig. 8** Drive operation with correct magnetic model
(a) d-axis current and its reference, (b) Estimated flux linkage in d and q axes, (c) q-axis current and its reference, (d) Flux linkage compensation in d and q axes.
steady-state level was confirmed by the torque meter reading at the end of the second transient.

3.4 MTPA/constant power/maximum-torque-per-voltage (MTPV) transition tests

The flux weakening capabilities of the proposed control structure are further tested with the help of Fig. 12a. An operating point $p_1$ on the MTPA curve is chosen, with a required torque curve $\tau$. The voltage limitation is represented in Fig. 12a by means of the voltage ellipse $\nu$ (part of it is displayed), which varies as a function of the voltage limit $U_{lim}$ and the motor speed. When $p_1$ is outside the voltage ellipse $\nu$ as in Fig. 12a, the proposed algorithm moves the operating point in order to guarantee the voltage limitation condition, which in Fig. 12a is the indicated by $p_2$, where the curve $\tau$ crosses the voltage ellipse $\nu$.

The described condition is in the constant power region because it still allows the use of the nominal current $I_{lim}$ since the voltage ellipse $\nu$ intersects the current limit circle in one point of the first quadrant. As the speed increases and $U_{lim}$ is kept constant, the voltage ellipse shrinks until it is completely within the current limit circle, as shown for example by the voltage ellipse $\nu'$ in Fig. 12a. In this condition, known as MTPV, the maximum allowed current is smaller than the limit value $I_{lim}$ and the flux weakening strategy must respect this new constraint.

In the following, the experimental evaluation of the proposed strategy in the different flux weakening regions is discussed. The voltage limit $U_{lim}$ was further reduced to 60 V in order to reach the MTPV region with safe speed values. The drive was

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**Fig. 9** Drive operation with magnetic model mismatch
(a) Mechanical speed, (b) Torque references and compensation, (c) Voltage reference magnitude and voltage limitation, (d) Current magnitude, reference magnitude and current limitation

**Fig. 10** Drive operation with magnetic model mismatch
(a) $d$-axis current and its reference, (b) Estimated flux linkage in $d$ and $q$ axes, (c) $q$-axis current and its reference, (d) Flux linkage compensation in $d$ and $q$ axes
forced to deliver its maximum allowed torque by setting a constant speed reference, and by increasing the load torque with very small steps in a quasi-static fashion until the actual motor speed was not matching the speed reference any longer. The last torque value fulfilling the given speed reference was recorded.

This procedure was repeated for 15 different speed references, from 80 to 220 rad/s with steps of 10 rad/s. The different 15 operating points are marked with the circles displayed in Fig. 12b. The first four points lie on the current limit circle, while the following points, obtained at higher speeds, are located in the MTPV region. The transition between the two flux weakening regions is smooth and as the speed increases, the operating point trajectories proceed towards the origin of the axes because of the absence of permanent magnets.

The experimental results are compared with the theoretical ones, calculated offline and marked with a cross in Fig. 12b. These were obtained by computing the voltage ellipses for each of the experimental speed values, and by extracting the intersecting point between each ellipse and its maximum allowed torque curve. The theoretical and experimental trajectories show a good match, validating the flux weakening method.

4 Conclusions

The study describes a model-based regulation algorithm for synchronous machines and in particular for SynRMs, where the non-linear magnetic model information is exploited to provide instantaneous flux weakening capabilities with optional real-time compensation of magnetic parameters mismatch. The instantaneous flux weakening capability is achieved by combining LUTs and nested binary searches of current and voltage references, executed until the references fulfill the drive constraints on maximum current and voltage. The nested binary searches are executed within one control cycle without any external voltage regulation loop.

The algorithm has been successfully tested on an 11 kW SynRM prototype with and without a correct magnetic model of the machine. Different operating conditions were tested too, proving a smooth transition between the MTPA/constant power/MTPV operating regions and thus confirming the validity of the proposed approach.

5 References


