A Very Simple Strategy for High Quality Performance of AC Machines Using Model Predictive Control

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Abstract—This paper presents a new and very simple strategy for torque and flux control of AC machines. The method is based on Model Predictive Control and uses one cost function for the torque and a separate cost function for the flux. This strategy introduces a drastic simplification, achieving a very fast dynamic behavior in the controlled machines. Experimental results obtained with an induction machine confirm the drive’s very good performance.

Index Terms—Predictive Control, Drives, Power Electronics

I. INTRODUCTION

The control of electrical machines has been one of the most classical and challenging problems of electrical engineering. With the explosive development observed in electromobility in the last decade, the control of electrical machines is of highest interest for industry today.

Two strategies are widely accepted as standard solutions for high performance AC drives: Field Oriented Control (FOC) and Direct Torque Control (DTC). FOC was invented in 1972 [1], [2] and DTC was invented in 1986 [3], [4], these strategies were developed more than 30 years ago, at a time where modern microprocessors were not available. Microprocessors have since been used to improve the performance of these strategies without introducing significant changes in the basic concepts of the theories.

However, the tremendous calculation power available today at high speeds and reduced costs makes it possible to develop different control strategies. In effect, Model Predictive Control is one of these modern control strategies that use microprocessors’ calculation power differently in the field of power electronics [5]–[15]. Up to now, the Finite Control Set Model Predictive Control (FCS-MPC) for torque and flux of AC machines has been done mainly using a single cost function with a weighting factor to give more importance to one of these control objectives [16]–[18].

The calculation of the weighting factor has been one of this control strategy’s important challenges. In most cases, the weighting factor is obtained by a trial and error process that is not easy or elegant, nor is it acceptable for many users [13]–[15], [19]–[23].

This paper presents a new strategy for predictive torque and flux control of AC machines that does not use weighting factors. This strategy is called Sequential Model Predictive Control (SMPC), and it uses a sequential structure with a single cost function for each control objective in the system. The first stage controls the torque, and the second stage is dedicated to controlling the flux. The resulting strategy solves in a very simple and logical way, all the problems and difficulties related to the calculation of the weighting factors.

The following sections of the paper will present the mathematical models for the machine and the inverter, the prediction equations, the control strategy and the experimental results obtained with an induction machine.

II. MATHEMATICAL MODELS

A. The Power Inverter

The inverter used in this work is the 2-level Voltage Source Inverter (2L-VSI). Fig. 1 shows the power circuit of the 2L-VSI. This inverter is the simplest and most mature power inverter technology; it has only two power switches for each output leg that work complementarily, but it generates a large harmonic content. However, as the focus of this work is the control strategy, this simple inverter is used.

Fig. 2 shows the possible voltage vectors generated by the 2L-VSI. There are eight possible voltage vectors described in
This article has been accepted for publication in a future issue of this journal, but has not been fully edited. Content may change prior to final publication. Citation information: DOI 10.1109/TPEL.2018.2812833, IEEE Transactions on Power Electronics

B. Model of the Induction Machine

To generate the mathematical model of the induction machine (IM) the stator flux $\Psi_s$ and stator current $i_s$ are taken as state variables. The dynamic equations of IM can be expressed in stationary frame as follow [24], [25]:

$$v_s = R_s i_s + \frac{d\Psi_s}{dt}$$  \hspace{1cm} (4)

$$0 = R_r i_r + \frac{d\Psi_r}{dt} - j\frac{\omega}{p} \Psi_r$$  \hspace{1cm} (5)

$$\Psi_s = L_s i_s + L_m i_r$$  \hspace{1cm} (6)

$$\Psi_r = L_m i_s + L_r i_r$$  \hspace{1cm} (7)

$$T = \frac{3}{2}p|\Psi_s \otimes i_s|$$  \hspace{1cm} (8)

$$J \frac{d\omega}{dt} = T - T_L$$  \hspace{1cm} (9)

where $v_s$ is the voltage vector, $\omega$ denotes the rotor angular speed, $p$ is the pair of poles, and $R_s$ and $R_r$ are the stator and rotor resistance, respectively. $L_s$, $L_r$ and $L_m$ are the stator, rotor and mutual inductance, respectively. Finally, $T$ and $T_L$ are the electrical torque and load torque, respectively.

### III. Equations for Prediction

For prediction of torque and flux [8], [14], [20], estimation of the stator flux $\Psi_s$ and the rotor flux $\Psi_r$ are required at the present sampling time $k$.

The rotor flux can be calculated using the equivalent equation of the rotor dynamics of an IM in rotating reference frame aligned with the rotor winding, which gives:

$$\Psi_r + \tau_r \frac{d\Psi_r}{dt} = L_m i_s$$  \hspace{1cm} (10)

where $\tau_r = L_r/R_r$ is the rotor time constant. Using the backward-Euler discretization and considering $T_s$ as the sampling time, the discrete-time equation for the rotor flux estimation is as follows:

$$\Psi_r^{k+1} = L_m T_s i_s^{k+1} + \left(1 - \frac{T_s}{\tau_r}\right) \Psi_r^k$$  \hspace{1cm} (11)

The stator flux can be estimated by the equation:

$$\Psi_s^k = \frac{L_m}{L_r} \Psi_r^k + \left(1 - \frac{L_m^2}{L_s L_r}\right) i_s^k$$  \hspace{1cm} (12)

Now, the stator flux prediction is obtained by the forward-Euler discretization:

$$\Psi_s^{k+1} = \Psi_s^k + T_s v_s^k - T_s R_s i_s^k$$  \hspace{1cm} (13)

The stator current prediction is also obtained by the forward-Euler discretization:

$$i_s^{k+1} = C_1 i_s^k + C_2 \Psi_s^k + \frac{T_s}{L_s} v_s^k$$  \hspace{1cm} (14)

where $R_{\sigma} = (R_s + (L_m/L_r)^2 R_r)$ corresponds to the equivalent resistance, $C_1 = (1 - (R_{\sigma} T_s/L_{\sigma}))$, $L_{\sigma} = \sigma L_s$ is the leakage inductance of the machine and $C_2 = (L_m/L_r) T_s/L_{\sigma}((1/\tau_r) - j\omega)$.  

### Table I: Possible switching states of 3 phase 2L-VSI

<table>
<thead>
<tr>
<th>Switching State</th>
<th>Voltage Vector</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_A$</td>
<td>$S_B$</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
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<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

Table I, and vectors $v_0$ and $v_7$ are the null voltage vectors ($v_\alpha = 0; v_\beta = 0$).

The mathematical equations that describe the 2L-VSI are:

$$v_a = S_a \frac{V_{DC}}{2}$$  \hspace{1cm} (1)

$$v_b = S_b \frac{V_{DC}}{2}$$  \hspace{1cm} (2)

$$v_c = S_c \frac{V_{DC}}{2}$$  \hspace{1cm} (3)

The voltage in $\alpha - \beta$ frame can be written as:

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = \frac{2}{3} V_{DC} \begin{bmatrix} 1 & -0.5 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix}$$

Fig. 2: Vectors of the 3-phase 2L-VSI.
Finally, the torque prediction depends on the stator flux and stator current predictions and can be written as follow:

$$T^{k+1} = \frac{3}{2}b|\Psi^*_s|^{k+1} \otimes i_s^{k+1}$$  \hfill (15)

IV. THE CONTROL STRATEGY

The proposed control strategy, called Sequential Model Predictive Control (SMPC), uses a cascade structure to control more than one control objective. The strategy uses a sequence of cost functions to control each control objective. Instead of using a single cost function with several control objectives related by a weighting factor, the problem is solved by using different cost functions, each of which is dedicated to controlling a single control objective.

It should be noted that in the implementation of the predictive control strategy, the delay in the application of the optimal vector must be considered because the measurement, the data processing, and the optimization algorithm are not instantaneous. To compensate for this delay, the control variables should be predicted for the future instant $k+2$. This delay compensation strategy is well documented in [26].

The block diagram of the SMPC strategy is presented in Figure 3. The error between the reference speed ($\omega^*$) and the measured speed ($\omega$) is introduced to a Proportional-Integral (PI) controller, which delivers the reference Torque ($T^*$) to be generated by the machine.

The cost function for the torque control ($g_1$) is given by:

$$g_1 = (T^* - T^{k+2})^2$$  \hfill (16)

where $T^{k+2}$ is the predicted torque, given by:

$$T^{k+2} = \frac{3}{2}b|\Psi^*_s|^{k+2} \otimes i_s^{k+2}$$  \hfill (17)

This cost function is represented by block 2 of the block diagram in Fig. 3. In addition, $g_1$ is calculated for all seven different voltage vectors generated by the inverter.

The two voltage vectors that generate the smallest values for $g_1$ (that is, the smallest error) are selected for the next control step, which corresponds to the minimization of the flux error. This action is performed by the cost function $g_2$, which corresponds to the flux error, defined by:

$$g_2 = (\Psi^* - \Psi_s^{k+2})^2$$  \hfill (18)

where $\Psi_s^{k+2}$ is the predicted flux, given by:

$$\Psi_s^{k+2} = \Psi_s^{k+1} + T_s v_s^{k+1} - T_s R_s i_s^{k+1}$$  \hfill (19)

This cost function is evaluated for each of the two voltage vectors selected by the previous step of torque control. This operation is represented by block 3 in Fig. 3.

Finally, the voltage vector that minimizes $g_2$ is selected and delivered to the load.

In Fig. 3, block 4 represents the power circuit of the inverter, block 5 represents equations (11) and (12) for flux estimation and block 6 represents equations (19) and (17) for flux and torque prediction.

Fig. 4 presents the flow diagram of the control strategy. The strategy starts measuring stator current ($i_s$) and speed at sampling interval ($k$), what is observed in step 1 of Fig. 4.

In step 2 the voltage vector calculated in the previous sampling interval is applied.

Step 3 estimates stator flux and rotor flux at sampling interval $k$.

Step 4 calculates $g_1$ for all seven voltage vectors.

Step 5 selects the two vectors with the smallest value for $g_1$.

Step 6 calculates $g_2$ for the two voltage vectors selected in the previous step.

Finally, step 7 selects the voltage vector that minimizes $g_2$ to be applied at the next sampling interval.

V. EXPERIMENTAL VALIDATION

A. The Test Bench

The test-bench consists of two 2.2 kW squirrel-cage induction motors, the load side and main motors. The load side machine is driven by a Danfoss VLT FC-302 3.0 kW inverter. The main motor is driven by a modified SERVOSTAR620 14 kVA inverter that provides full control of the IGBT gates.

A self-made 1.4 GHz real-time computer system is used. The rotor position is measured by a 1024-point per revolution incremental encoder. The sampling frequency is 16kHz. The average switching frequency is around 3.3kHz.

Table II shows the parameters of the test bench and Fig. 5 shows the equipment used in the laboratory.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC link voltage $V_{DC}$</td>
<td>582V</td>
</tr>
<tr>
<td>$R_s$</td>
<td>2.68Ω</td>
</tr>
<tr>
<td>$R_r$</td>
<td>2.13Ω</td>
</tr>
<tr>
<td>$L_m$</td>
<td>275.1mH</td>
</tr>
<tr>
<td>$L_s$</td>
<td>283.4mH</td>
</tr>
<tr>
<td>$L_r$</td>
<td>283.4mH</td>
</tr>
<tr>
<td>$p$</td>
<td>1</td>
</tr>
<tr>
<td>$\omega_{nom}$</td>
<td>2772.0RPM</td>
</tr>
<tr>
<td>$T_{nom}$</td>
<td>7.5 Nm</td>
</tr>
<tr>
<td>$J$</td>
<td>0.005kg/m²</td>
</tr>
</tbody>
</table>

B. Results

Fig. 6 shows the drive’s dynamic response in a speed reversal of ±2772 RPM. The variables recorded are: speed ($\omega$), torque (T) and stator current ($i_s$). During this operation the amplitude of the stator flux is kept constant. It can be observed that the stator current has a fast increase in its amplitude, generating a fast change in the torque. The speed shows a smooth transition from 2772 RPM to -2772 RPM.

Fig. 7 shows the steady state behavior of the drive. The variables in this figure are stator current ($i_s$), torque (T), stator flux ($\Psi_{st}$) and stator voltage ($v_s$). All variables show the typical waveforms delivered by a two-level inverter.
Fig. 3: Block diagram of Sequential Model Predictive Control (SMPC) of a 2L-VSI.

Fig. 4: Flow diagram of Sequential Model Predictive Control (SMPC) of a 2L-VSI.

Nevertheless, it is possible to see that stator flux shows a DC drift towards negative direction, as the flux is estimated based on the original measurement, i.e., the phase currents (a and b) in our test-bench, the measurements are not perfect in accuracy, errors will happen, which will introduce small bias at the end of this estimated flux. Some analysis has already been published in e.g. cite, and a potential solution can be that, using a full order estimator to get rid of this flux bias.

Fig. 5: Experimental test bench.

VI. CONCEPTUAL ASSESSMENT WITH DTC

The proposed strategy is different to Direct Torque Control (DTC) and standard Model Predictive Control. The main features of DTC are:

1) Two hysteresis are used to control Torque and Flux.
2) The engineer/user must know the effect that each voltage vector will have on the behavior of Torque and Flux to decide which voltage will be delivered to the load.
Fig. 6: Experimental results for speed reversal of ±2772 RPM: (a) Rotor speed ($\omega$); (b) Torque (T); (c) Stator current ($i_a$).

Fig. 7: Experimental results for steady state: (a) Stator current ($i_a$); (b) Torque (T); (c) Stator Flux ($\psi_a$); (d) Stator voltage ($v_a$).

3) The position of the stator flux in the complex plane must be identified by the control to select the right direction of the lookup table.

None of these important and necessary features are needed or considered using our proposed strategy, making it much simpler than DTC.

Fig. 9 shows the block diagram of DTC and the standard MPC. It is possible to see the difference between the standard MPC and the proposed control strategy.

VII. COMMENTS AND CONCLUSIONS

This paper has presented a new and very simple strategy for high performance control of an induction machine called Sequential Model Predictive Control (SMPC).

The method uses the approach of Model Predictive Control and is based on the fundamental equations of the machine and of the inverter.

SMPC calculates the variables of the system in a sequential way using a single cost function for each control objective. Moreover, this work demonstrates that it is not necessary to use weighting factors to control torque and flux when using predictive control.

Experimental results confirm that the strategy effectively controls torque and flux. This simple strategy eliminates the problem of calculating any weighting factor.

MPC is conceptually different from established strategies for high performance control of AC machines. It uses the capabilities of modern microprocessors and the discrete analysis of
the system to be controlled (inverter and machine) in a simple way.

Finally, these results confirm that this strategy is a very attractive and promising alternative for high performance AC drives.

REFERENCES


Zhenbin Zhang (S’13–M’16), was born in Shandong, China. He received the B.S. degree in electrical engineering and automation from Harbin Engineering University, Harbin, China, in 2008, and the Ph.D. degree (summa cum laude) from the Institute for Electrical Drive Systems and Power Electronics, Technical University of Munich, Munich, Germany, in 2016. From 2008 to 2011, he studied control theory and engineering with Shandong University, Jinan, China. From 2016 to 2017, he was a Research Fellow and the team leader for the "Modern Control Strategies for Electrical Drives" Group with the Institute for Electrical Drive Systems and Power Electronics, Technische Universitaet Muenchen. Since 2017, he has been a Full Professor with Shandong University. His research interests include power electronics and electrical drives, sustainable energy systems, and smart grids. Dr. Zhang was a recipient of the VDE Award, Germany, in 2017.

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Ralph M. Kennel was born in 1955 at Kaiserslautern (Germany). In 1979 he got his diploma degree and in 1984 his Dr.-Ing. (Ph.D.) degree from the University of Kaiserslautern. From 1983 to 1999 he worked on several positions with Robert BOSCH GmbH (Germany). Until 1997 he was responsible for the development of servo drives. Dr. Kennel was one of the main supporters of VECON and SERCOS interface, two multi-company development projects for a microcontroller and a digital interface especially dedicated to servo drives. Furthermore he took actively part in the definition and release of new standards with respect to CE marking for servo drives. Between 1997 and 1999 Dr. Kennel was responsible for “Advanced and Product Development of Fractional Horsepower Motors” in automotive applications. His main activity was preparing the introduction of brushless drive concepts to the automotive market. From 1994 to 1999 Dr. Kennel was appointed Visiting Professor at the University of Newcastle-upon-Tyne (England, UK). From 1999 - 2008 he was Professor for Electrical Machines and Drives at Wuppertal University (Germany). Since 2008 he is Professor for Electrical Drive systems and Power Electronics at Technische Universitaet Muenchen (Germany). His main interests today are: Sensorless control of AC drives, predictive control of power electronics and Hardware-in-the-Loop systems. Dr. Kennel is a Senior Member of IEEE, a Fellow of IET (former IEE) and a Chartered Engineer in the UK. Within IEEE he is Treasurer of the Germany Section as well as Distinguished Lecturer of the Power Electronics Society (IEEE-PELS). Dr. Kennel has received in 2013 the Harry Owen Distinguished Service Award from IEEE-PELS as well as the EPE Association Distinguished Service Award in 2015. Dr. Kennel was appointed “Extraordinary Professor” by the University of Stellenbosch (South Africa) from 2016 to 2019 and as “Visiting Professor” at the Haixi Institute by the Chinese Academy of Sciences from 2016 to 2021. There he was appointed as “Jiaxi Lu Overseas Guest Professor” in 2017.