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## Latest Advances of Model Predictive Control in Electrical Drives

### *Part I: Basic Concepts and Advanced Strategies*

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# Latest Advances of Model Predictive Control in Electrical Drives. Part I: Basic Concepts and Advanced Strategies

Jose Rodriguez, *Life Fellow, IEEE*, Cristian Garcia, *Member, IEEE*, Andres Mora, *Member, IEEE*, Freddy Flores-Bahamonde, *Member, IEEE*, Pablo Acuna, *Member, IEEE*, Mateja Novak, *Member, IEEE*, Yongchang Zhang, *Senior Member, IEEE*, Luca Tarisciotti, *Member, IEEE*, S. Alireza Davari, *Senior Member, IEEE*, Zhenbin Zhang, *Senior Member, IEEE*, Fengxiang Wang, *Senior Member, IEEE*, Margarita Norambuena, *Member, IEEE*, Tomislav Dragicevic, *Senior Member, IEEE*, Frede Blaabjerg, *Fellow, IEEE*, Tobias Geyer, *Senior Member, IEEE*, Ralph Kennel, *Senior Member, IEEE*, Davood A. Khaburi, Mohamed Abdelrahem, *Senior Member, IEEE*, Zhen Zhang, *Senior Member, IEEE*, Nenad Mijatovic, *Member, IEEE*, Ricardo P. Aguilera, *Senior Member, IEEE*.

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J. Rodriguez, F. Flores-Bahamonde and L. Tarisciotti are with the Department of Engineering Sciences, Universidad Andres Bello, Santiago, 30332 Chile (e-mail: jose.rodriguez; freddy.flores; luca.tarisciotti@unab.cl).

C. Garcia and P. Acuna are with the Department of Electrical Engineering, Universidad de Talca, Curico, 3340000 Chile (e-mail: cristian.garcia; pablo.acuna@utalca.cl).

A. Mora and M. Norambuena are with the Department of Electrical Engineering, Universidad Tecnica Federico Santa Maria, Valparaiso, 30332 Chile (e-mail: andres.mora; margarita.norambuena@usm.cl).

M. Novak and F. Blaabjerg are with the AAU Energy, Aalborg University, Aalborg 9220, Denmark (e-mail: nov; fbl@energy.aau.dk).

Y. Zhang is with School of Electrical and Electronic Engineering, North China Electric Power University, Beijing 102206, China (e-mail: yozhang@ieec.org).

A. Davari is with the Faculty of Electrical Engineering, Shahid Rajae Teacher Training University, Tehran 16788 15811, Iran (e-mail: davari@sru.ac.ir).

Z. Zhang is with the School of Electrical Engineering, Shandong University, Jinan 250061, China (e-mail: zbz@sdu.edu.cn)

F. Wang is with Quanzhou Institute of Equipment Manufacturing, Haixi Institutes, Chinese Academy of Sciences, Jinjiang, 362200, China (e-mail: fengxiang.wang@fjirms.ac.cn).

D. A. Khaburi is with the Department of Electrical Engineering, Iran University of Science and Technology, Tehran 1311416846, Iran (e-mail: khaburi@iust.ac.ir).

T. Dragicevic and N. Mijatovic are with the Department of Electrical Engineering, Technical University of Denmark, 2800 Lyngby, Denmark (e-mail: tomdr; nm@elektro.dtu.dk).

T. Geyer is with the ABB System Drives, 5300 Turgi, Switzerland (e-mail: t.geyer@ieec.org).

M. Abdelrahem is with the Institute for Electrical Drive Systems and Power Electronics, Technical University of Munich (TUM), 80333 Munich, Germany and also with the Electrical Engineering Department, Faculty of Engineering, Assiut University, 71516 Assiut, Egypt (e-mail: mohamed.abdelrahem@tum.de).

R. Kennel is with the Institute for Electrical Drive Systems and Power Electronics, Technical University of Munich (TUM), 80333 Munich, Germany (e-mail: ralph.kennel@tum.de).

Z. Zhang is with School of Electrical and Information Engineering, Tianjin University 92 Weijin Road, Nankai District, Tianjin 300072, Tianjin, China (e-mail: zhangz@tju.edu.cn).

R. P. Aguilera is with the School of Electrical and Data Engineering, at the University of Technology Sydney, Broadway, NSW 2007, Australia. (e-mail: raguilera@ieec.org).

**Abstract**—The application of Model Predictive Control (MPC) in electrical drives has been studied extensively in the last decade. This paper presents what the authors consider the most relevant contributions published in the last years, mainly focusing on three relevant issues: weighting factor calculation when multiple objectives are utilized in the cost function, current/torque harmonic distortion optimization when the power converter switching frequency is reduced, and robustness improvement under parameters uncertainties. Therefore, this paper aims to enable readers to have a more precise overview while facilitating their future research work in this exciting area.

**Index Terms**—Model predictive control, variable speed drives, weighting factors, electrical machine.

## I. INTRODUCTION

Electromobility is one of the most relevant actions to fight against global warming. Starting in 2030's, several countries will only approve new cars that do not use fossil fuels [1], [2]. Although the control of motor drive applications has been one of the most classic electrical engineering challenges, today, with the development of electric vehicles, it has become a very relevant and timely research subject [3]. The development of microprocessors has motivated the exploration of more sophisticated strategies to control electrical machines using power semiconductors. MPC is one of these control techniques [4]–[9]. The interest in using this technique in electrical drives is based on two essential facts: i) in electrical and mechanical engineering, some outstanding mathematical models allow the prediction of the future behavior of the system with high accuracy, and ii) modern microprocessors allow to do a very high number of real-time calculations at an affordable cost. Besides, MPC is easy to understand, easy to implement, can deal with nonlinearities, and can handle several variables simultaneously [10] [11]–[13].

The main challenges that should be solved when using MPC strategies in motor drive applications are:

- i. The quality improvement of the load current/torque when the power converter switching frequency is reduced.

- ii. The weighting factor calculation in the cost function to control more than one variable simultaneously.
- iii. The dependency of MPC strategy from the quality of the models and the parameter mismatches.

Consequently, this paper presents the most relevant contributions recently reported in the literature to overcome these challenges. In this regard, the machine currents' quality has been significantly improved using a Finite Control-Set Model Predictive Control (FCS-MPC) strategy with multiple prediction steps [14]–[20]. Furthermore, a very attractive solution has been developed for three-level neutral-point clamped converters in medium voltage [21]–[24]. Here, optimized pulse patterns are embedded with an MPC strategy to improve performance while reducing the converter losses notably. The weighting factor calculation in the cost function, running simulations, or using some iterative solutions is not particularly appreciated by the community because it does not guarantee the operation in an optimal point [25]. The use of Artificial Neural Networks (ANN) for the calculation of the weighting factors emerges as a very powerful solution that integrates power electronics with techniques of artificial intelligence [26]. Another interesting solution is to avoid the need for weighting factors, using what is known as Sequential MPC, obtaining very promising results [27]. The parameter dependency is addressed by a new strategy known as Model-Free MPC. Here, the prediction is performed using estimated variables, improving significantly the robustness of the MPC technique [28]. The next pages will explain the most relevant features of these strategies, including also additional references.

## II. CLASSICAL MPC APPROACHES IN DRIVES

As is well-known, there are two main control strategies for electrical drives: Field-Oriented Control (FOC) and Direct Torque Control (DTC) [29]–[32]. These approaches achieve closed-loop control of the stator currents with high dynamic performance. The first one relies on an inner PI-based current control loop and, the second one uses a predefined lookup table that relates the position of the stator flux and the sign of the torque and flux error. Starting from these two basic strategies, MPC has been applied in electrical drives to control the internal stator current (Model Predictive Current Control, MPCC) and also to directly control the machine's torque and flux (Model Predictive Torque Control, MPTC) [33]–[40].

The analysis here presented is based on the control of an induction machine (IM). Thus, the five classic electrical and mechanical equations in stationary reference frame are described as [10], [41]:

$$\mathbf{v}_s = \mathbf{i}_s R_s + \frac{d\boldsymbol{\psi}_s}{dt} \quad (1)$$

$$0 = \mathbf{i}_r R_r + \frac{d\boldsymbol{\psi}_r}{dt} - j\omega \boldsymbol{\psi}_r \quad (2)$$

$$\boldsymbol{\psi}_s = L_s \mathbf{i}_s + L_m \mathbf{i}_r \quad (3)$$

$$\boldsymbol{\psi}_r = L_r \mathbf{i}_r + L_m \mathbf{i}_s \quad (4)$$

$$T_e = \frac{3}{2} p \text{Im}\{\bar{\boldsymbol{\psi}}_s \mathbf{i}_s\}, \quad (5)$$

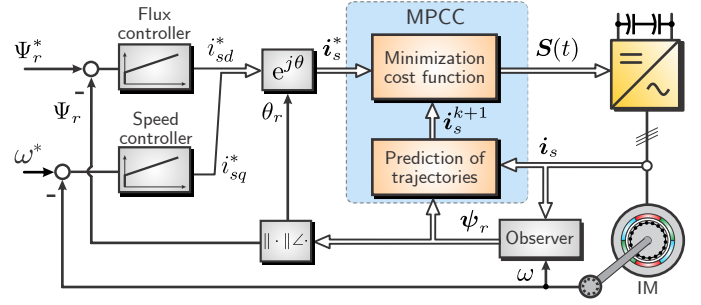


Fig. 1: A cascaded field-oriented control structure with speed and flux outer control loop and the MPCC inner control loop.

where  $\mathbf{v}_s$  is the stator voltage vector,  $\boldsymbol{\psi}_s$  and  $\boldsymbol{\psi}_r$  are the stator and rotor flux,  $\mathbf{i}_s$  and  $\mathbf{i}_r$  are the stator and rotor current,  $T_e$  is the electromagnetic torque, and  $\omega$  is the speed. The parameters  $R_s$  and  $R_r$  are the stator and rotor resistances,  $L_s$ ,  $L_r$  and  $L_m$  represent the stator, rotor and mutual inductances. In (5),  $\bar{\boldsymbol{\psi}}_s$  stands for the complex conjugate of  $\boldsymbol{\psi}_s$  and  $\text{Im}\{\cdot\}$  operator denotes the imaginary part.

### A. Model Predictive Current Control

Based on the IM model described above, the dynamic model for the stator current can be expressed as [10], [41], [42]:

$$\frac{d\mathbf{i}_s}{dt} = -\frac{R_\sigma}{\sigma L_s} \mathbf{i}_s + \frac{k_r}{\sigma L_s} \left( \frac{1}{\tau_r} - j\omega \right) \boldsymbol{\psi}_r + \frac{1}{\sigma L_s} \mathbf{v}_s, \quad (6)$$

where  $\sigma = 1 - k_r k_s$  is the total leakage coefficient,  $k_r = \frac{L_m}{L_r}$  and  $k_s = \frac{L_m}{L_s}$  are the magnetic coupling factors, and  $R_\sigma = R_s + k_r^2 R_r$ .

Several discretisation methods have been proposed to obtain the predictions of the stator currents from (6) [10]. The most utilised method is the Forward Euler discretisation, which leads to the following discrete-time prediction model for the stator currents:

$$\mathbf{i}_s^{k+1} = \left( 1 - \frac{T_s R_\sigma}{\sigma L_s} \right) \mathbf{i}_s^k + \frac{T_s k_r}{\sigma L_s} \left( \frac{1}{\tau_r} - j\omega^k \right) \boldsymbol{\psi}_r^k + \frac{T_s}{\sigma L_s} \mathbf{v}_s^k, \quad (7)$$

where  $T_s$  is the sampling period.

For the MPCC strategy, usually, the rotor flux reference is considered constant during the sampling period. Consequently, to control the motor, the expression (7) is continuously evaluated to predict the stator current selecting the proper voltage vector. The cost function is characterized by its high flexibility and the possibility to add systems constraints and, it is expressed as

$$g = (i_{s\alpha}^* - i_{s\alpha}^{k+1})^2 + (i_{s\beta}^* - i_{s\beta}^{k+1})^2. \quad (8)$$

Finally, the classical control scheme for MPCC is depicted in Fig. 1, which shows the inner control loop performing the predictive current control, and the outer loop handling speed and torque control [4], [10], [43].

### B. Model Predictive Torque Control

Similarly to MPCC, several reports demonstrate the advantages of model predictive torque control (MPTC), highlighting

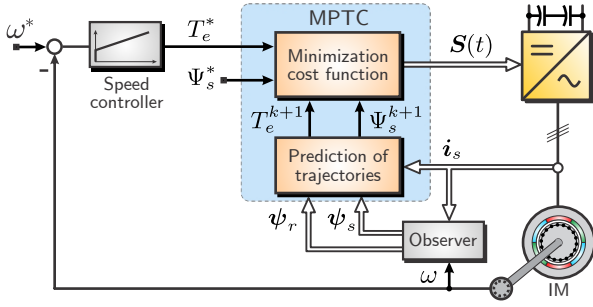


Fig. 2: Control diagram of the MPDTC strategy.

its very fast response under torque steps maintaining the stator flux amplitude constant. Besides, note that in comparison with MPCC no current controller is implemented in MPTC. To control the electromagnetic torque predictions of the stator current  $i_s^{k+1}$  and stator flux  $\psi_s^{k+1}$  must be evaluated for the seven possible voltage vectors of the 2L-VSI inverter. Expressions for future torque and flux are [10], [26], [38]:

$$\psi_s^{k+1} = \psi_s^k + T_s v_s^k - R_s T_s i_s^k, \quad (9)$$

$$T_e^{k+1} = \frac{3}{2} p \text{Im} \{ \bar{\psi}_s^{k+1} i_s^{k+1} \} \quad (10)$$

The cost function is defined as a linear combination of torque and flux errors, as described in the following equation

$$g = (T_e^* - T_e^{k+1})^2 + \lambda_\psi (\Psi_s^* - \Psi_s^{k+1})^2, \quad (11)$$

where  $\Psi_s = \|\psi_s\|$  is the amplitude of the stator flux vector and  $\lambda_\psi$  is the weighing factor. The classical control diagram of this method is illustrated in Fig. 2. In this figure, the torque reference is obtained from a speed loop controller and the stator flux magnitude is usually set as a constant. Finally, as is detailed in Fig. 2, the torque and flux variables are estimated by measuring the speed and the stator currents.

An important problem in (11) is the calculation of the weighting factor  $\lambda_\psi$ , which has a critical influence on the overall performance of the drive.

### C. Model Predictive Speed Control

In addition to torque and machine current control, MPC is also often considered to implement the speed control loop of a drive. In fact, MPC can provide benefits in this specific application, such as robustness to parameter variation and easy inclusion of flux weakening constraints. In [44] a linear implementation of MPC is proposed for the speed control loop of a servo PMSM drive, showing good results, superior to the one obtained with classical PI control. However, being the MPC solution obtained analytically considering a linear plant, external constraints cannot be added to the controller. In [45], [46], the case of MPC speed control for synchronous motor drive is studied. Also in this case, an unconstrained implementation of MPC, with the addition of integral terms in the cost function to compensate bias errors and external disturbances is considered. Results show improved performance with respect to classical PI speed control and excellent robustness to mechanical model mismatches and electrical

parameters uncertainties. Finally, in [47] a sensorless MPC speed loop for a PMSM drive is designed where the speed MPC is combined with the position estimation in the current loop. Since the estimated rotor position cannot be as reliable as the real position sensor information, the MPC reference speed is adjusted according to the position estimation reliability. This method prevents system instability and achieve achieving superior steady state and dynamic response. In [44]–[47] the MPC speed loop is always cascaded with an external torque control, which can be implemented using MPC or in a classical fashion. However, taking advantage of the MPC flexibility, it is possible to combine the speed and torque control in a single control loop. An effort in this direction has been taken in [48]–[50] where combined speed and torque MPC has been applied to Induction Machines [48], PMSM [49] and IPM synchronous motor drive [50], respectively. Results shows that, despite the more complex implementation and the additional calculation required by the control, the computational burden can be contained and performances, in terms of load variation recovery time and, in general, dynamic performances. Moreover, in this case a nonlinear constrained implementation of MPC is considered allowing a better response to the drive nonlinearities and improving robustness to system parameters variation.

### III. WEIGHTING FACTOR CALCULATION

Optimization of the weighting factors can be a very challenging task for cost functions with multiple objectives that are in conflict or can not be unified. Several solutions have been proposed to solve this optimization problem. Weightless MPC cost functions can be obtained by unifying the dimensions of the control objectives [51], or splitting the control problem into two cost functions [52]–[54]. However, the use of weighting factors can not always be avoided due to too many objectives that can not be unified. To calculate the weighting factors, methods that use artificial neural networks [26], [55], [56], fuzzy optimization [57] or multi-objective genetic algorithms [58] were proposed. It needs to be mentioned that the methods using online weighting factor calculation like in [55]–[57] can add additional computation burden to the control algorithm. On the other hand, the method introduced in [26] uses an offline calculation procedure of both the weighting factors and the flux reference providing a fast drive start and good performance during different loading conditions. The design process is straightforward and applicable to different cost functions [59]. Nevertheless, the price that has to be paid for the offline calculation is the additional memory to store the calculated weighting factors. In addition, the study cases are limited and interpolations techniques have to be implemented to cover all the online operation points.

#### A. Training the ANN and fitness function design

The function of the ANN in the weighting factor calculation is to become a fast and accurate surrogate of the system model with the ability to calculate the system performance metrics for different cost function parameters as shown in the Fig. 3 workflow. In order to create the surrogate, a system



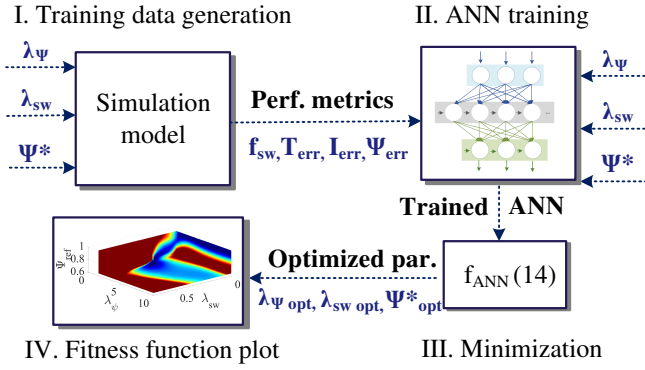


Fig. 3: Cost function parameter calculation workflow using the ANN.

simulation model can be used to collect the data for training the ANN. Alternatively, for more accuracy, experimental data can be used. However, it is more time-consuming and requires a human supervision. The training data of interest are the performance metrics like torque and stator current rms errors ( $T_{err}$ ,  $\Psi_{s\ err}$ ,  $I_{s\ err}$ ), average switching frequency ( $f_{sw\ avg}$ ) obtained for different cost function parameters. The MPTC cost function for controlling the torque, stator flux and the average switching frequency can be defined as:

$$g = (T_e^* - T_e^{k+1})^2 + \lambda_\psi (\Psi_s^* - \Psi_s^{k+1})^2 + \lambda_{sw} n_{sw} + h_{lim} \quad (12)$$

$$n_{sw} = \sum_{x=a,b,c} |S_x^k - S_x^{k-1}|, \quad (13)$$

where  $h_{lim}$  is limiting the stator current,  $\lambda_\psi$  and  $\lambda_{sw}$  are the weighting factors, and  $n_{sw}$  defines the number of switching transitions. Thus,  $S_x^{k-1}$  is the switching state applied in the previous sampling period and  $S_x^k$  is the current switching state.

When collecting the performance metrics for different cost function parameters  $\lambda_\psi$ ,  $\lambda_{sw}$  and  $\Psi_s^*$ , the range of the parameters should be defined to only sweep combinations that can lead to a successful drive start and steady state operation. A feed-forward ANN configuration was used with a back-propagating training algorithm to tune the parameters of the ANN. Once trained ANN can be used to calculate the performance metrics for all weighting factors within the defined range of the cost function parameters. Afterwards, a fitness function is defined using the performance metrics to find the cost function parameters that will provide the desired control performance. An example of a fitness function for different loading conditions is shown here:

$$f_{ANN} = \Psi_{s\ error}^2 + I_{s\ error}^2 + T_{s\ error}^2 + (f_1 - f_{sw\ avg})^2, \quad (14)$$

where  $f_1$  is the average switching frequency for which the optimum parameter values will be calculated and it is chosen arbitrary. By repeating the procedure in Fig. 3 for different operating conditions of the motor drive, the performance can be optimized for wide operating range of the motor drive. As it was shown in [26], the size of the optimum parameter region depends both on the load and the speed reference.

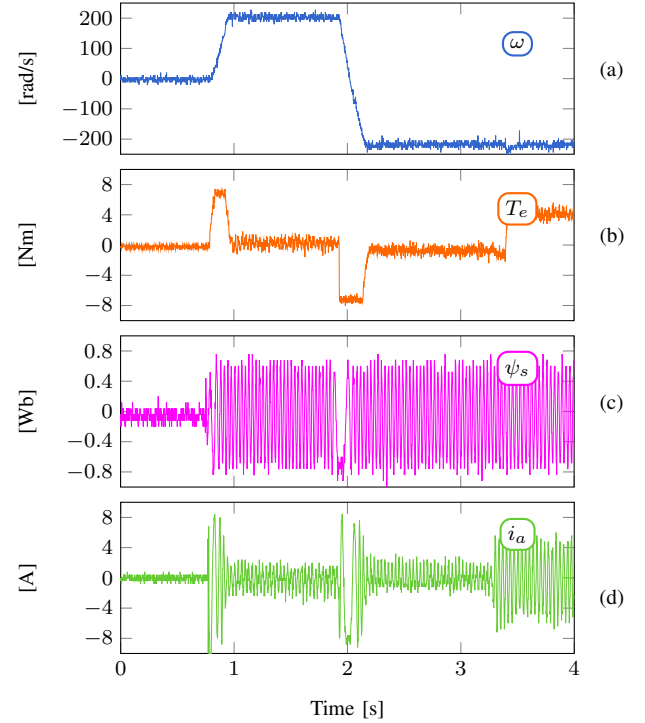


Fig. 4: Experimental waveforms of drive speed, torque, stator flux and current for  $f_{ANN}$  as per (14) and  $T_{load} = 5$  Nm,  $f_1 = 2.5$  kHz,  $\lambda_\psi = 9.64$ ,  $\lambda_{sw} = 0.13$  and  $\Psi_s^* = 0.65$  Wb.

## B. Design validation

Fig. 4 shows the experimental waveforms obtained using  $\lambda_\psi$ ,  $\lambda_{sw}$  and  $\Psi_s^*$  that were calculated with (14) for the frequency  $f_1 = 2.5$  kHz and the load torque of 5 Nm. It can be observed that the design allows a fast start, successful speed reversing and a good steady-state performance of the drive.

## C. Analytical Tuning for MPTC

As discussed above, the weighting factors' choice for MPTC strategy—and thus the tuning procedure—may be difficult because one has to decide on the relative importance between the torque error and the flux magnitude error. This trade-off also affects the stator current distortions, which are typically higher than the ones obtained with MPCC. Recently, in [39], [60], the value of the weighting factors that minimize the current distortions are analytically derived. However, MPCC still achieves lower current distortions, particularly at non-zero torque reference and low switching frequencies. To overcome this issue, a slight modification to the conventional MPTC is proposed in [40]. Here, the MPTC is made equivalent to MPCC by tracking the rotor flux magnitude  $\Psi_r^*$  instead of the stator flux magnitude  $\Psi_s^*$ . As a result, analytical expressions for the weighting factors can be derived, ensuring the same closed-loop performance as MPCC. Consequently, the proposed model predictive torque and flux controller achieves the same low current and torque distortions as MPCC, without requiring an outer field-oriented controller that sets the current references (see Fig. 1).

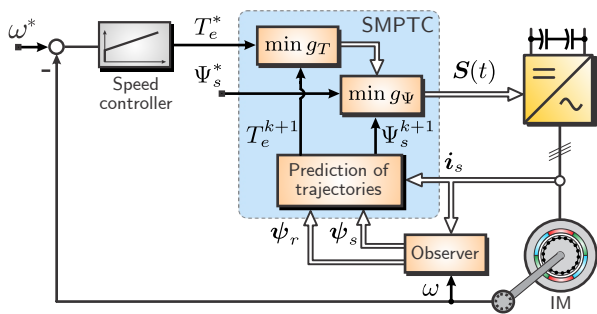


Fig. 5: Sequential model predictive flux and torque control.

#### IV. MPC WITHOUT WEIGHTING FACTORS

As discussed in Section III, the selection of weighting factors for a multi-objective cost function does not have an analytical solution. Its selection has been through a heuristic search and currently with advanced techniques such as ANN, which goes in the opposite direction of the conceptual simplicity of predictive control. It is for this reason that many authors have proposed techniques to avoid the use of weighting factors. In [61], a Model Predictive Torque Control (MPTC) with multi-objective optimization based on a ranking table is proposed, achieving an independence of the weighting factor, however a relative magnitude errors are missed and changed for discrete table. In [51] the electrical torque and stator flux magnitude references are converted into an equivalent reference vector of stator flux. There are also works that propose to control the torque and stator flux in an indirect way controlling the direct and quadrature stator current [62], in this way the cost function has objectives with the same magnitude. A modified MPTC is proposed in [63], where the electrical torque and stator flux are controlled through the control of torque and reactive torque, both have the same order magnitude so a weighting factor is not required. A deadbeat predictive direct torque control is proposed in [64], achieving control of torque and flux through the tracking of voltage vectors. One of the latest multi-objective control strategies for motor drives is the sequential predictive control [54]. This method achieves torque and flux control by evaluating two cost functions sequentially. Because it is intuitive and straightforward, this strategy has been implemented in other applications [65], improved with more sophisticated techniques [66], and analyzed in depth [67], [68].

The Sequential Model Predictive Torque Control (SMPTC) proposed in [54] is illustrated in Fig. 5. As shown in this figure, the SMPTC strategy has two control objectives: the electromagnetic torque and amplitude of the stator flux, controlled using two cost functions sequentially,

$$g_T = (\Delta T_e)^2 = (T_e^* - T_e^{k+1})^2, \quad (15)$$

$$g_\Psi = (\Delta \Psi_s)^2 = (\Psi_s^* - \Psi_s^{k+1})^2. \quad (16)$$

SMPTC as well as any finite-set model predictive torque control needs a discrete model to predict the behavior of variables. The prediction of electric torque and stator flux of the machine is based on (9) and (10), considering the forward Euler discretization as presented in [69], [70] and [71].

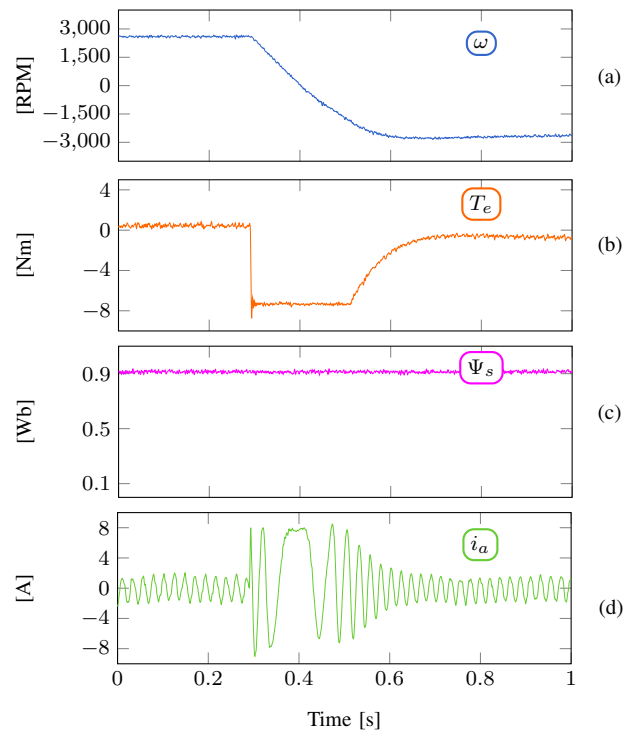


Fig. 6: SMPTC during speed reversal maneuver: (a) Motor speed  $\omega$ ; (b) Electric torque  $T_e$ ; (c) Stator flux magnitude  $|\psi_s|$ ; (d) Stator current  $i_\alpha$ .

In summary, the SMPTC can be described in the following three simple steps:

- i. The electric torque is predicted for the  $(k+1)$ -th instant considering all the possible vectors of the 2L-VSI using the discrete-time model of the stator flux and torque.
- ii. The cost function  $g_T$  in (15) is evaluated. The two vector that minimizes the cost function will be considered for the next optimization stage.
- iii. The two optimal vectors found in the previous step are evaluated in the cost function  $g_\Psi$ , (16). The vector that minimizes  $g_\Psi$  will be the optimal vector and will be applied in the 2L-VSI in the next sampling time.

Thus, using the SMPTC strategy, weighting factors are avoided while keeping good performance in the motor drive.

An inversion speed maneuver from nominal speed to negative nominal speed in an induction machine controlled with SMPTC is depicted in Fig. 6. This figure shows the good behavior of this control technique, where the electric torque and the stator flux have a good dynamic and stationary behavior. The electric torque allows the speed inversion to be carried out with good dynamic performance. Simultaneously, the magnitude of the stator flux remains controlled and constant at all times, showing that there is no coupling between both control objectives.

Nevertheless, as analyzed in [72], the simplicity of SMPTC removes degrees of freedom that can be exploited to improve the overall system efficiency. This feature limits the controllability of the torque and machine magnetization, what can lead to suboptimal performance. As shown in [72], when

the torque and flux terms are minimized sequentially higher current distortions result over the whole range of switching frequencies.

## V. MULTI-STEP FCS-MPC

Multi-step FCS-MPC has recently gained attention due to its role as a key enabler for efficiency improvement, specially in high power electrical drives [42], [72], since it has the potential to reduce the stator current THD while retaining the power converter switching frequency. This leads to a power loss reduction in the electric machine, improving thus the overall system efficiency [42]. To achieve this, the control input in the problem formulation is either the switch positions [42] or the distinctive output voltage levels [16]. For the multi-step case, the optimal control input vector,  $\mathbf{u}_{\text{opt}}(k)$ , is the first element of the optimal control input sequence:

$$\mathbf{U}_{\text{opt}} = [(\mathbf{u}_{\text{opt}}(k))^T \cdots (\mathbf{u}_{\text{opt}}(k+N-1))^T]^T. \quad (17)$$

The vector  $\mathbf{U}_{\text{opt}}$  is found formulating a standard multi-step MPC problem that minimizes the stator current tracking error versus switching effort,  $\Delta \mathbf{u}(\ell) = \mathbf{u}(\ell) - \mathbf{u}(\ell-1)$ , over an extended prediction horizon  $N$ , i.e.:

$$\begin{aligned} \min_{\mathbf{U}} \quad & \sum_{\ell=k}^{k+N-1} \|\mathbf{i}_s(\ell+1) - \mathbf{i}_s^*(\ell+1)\|_2^2 + \lambda_u \|\Delta \mathbf{u}(\ell)\|_2^2 \\ \text{s.t.} \quad & \mathbf{U} \in \mathbb{U}^{3N} \text{ and } \|\Delta \mathbf{u}(\ell)\|_\infty \leq 1, \end{aligned} \quad (18)$$

where  $\mathbf{U}$  is the multi-step decision variable and  $\lambda_u > 0$ .

The first constraint in (18) can consider either the switching states or voltage levels, while the second constraint limits the output voltage level transitions to avoid internal damage in the converter and excessive  $dV/dt$  in the load.

Normally, to find  $\mathbf{U}_{\text{opt}}$ , it is required to evaluate all the possible combinations. However, for the multi-step case, the number of combinations exponentially increases with  $N$ , leading to a high computational burden. Recently [14], [15],

have shown that the control problem (18) can be reformulated as an integer least squares optimization problem. Then, a sphere decoding algorithm (SDA) enables microprocessors to efficiently solve (18) in real-time to obtain the multi-step optimal control vector  $\mathbf{U}_{\text{opt}}$  [16]–[19].

To do this, the SDA forms an initial sphere centered in  $\mathbf{U}_{\text{unc}}$ , which is the well known explicit unconstrained optimal solution. The initial sphere radius,  $\rho_{\text{ini}}$ , is determined as the Euclidean distance between  $\mathbf{U}_{\text{unc}}$  and an initial suitable candidate,  $\mathbf{U}_{\text{ini}}$ , chosen from the pool of available input combinations. Then, using a branch-and-bound technique, the SDA quickly discards all input combinations that lie outside of the initial sphere. As soon as the SDA finds a candidate inside the initial sphere, a new smaller sphere is formed and the process is repeated by evaluating the remaining candidates. Finally, the optimal solution  $\mathbf{U}_{\text{opt}}$  is found when the sphere cannot be further shrink, so its first element,  $\mathbf{u}_{\text{opt}}(k)$ , is taken to synthesize the switching state  $\mathcal{S}$  applied to the converter.

Similar to the system shown in Fig. 1, but considering a multi-step MPCC, the best trade-off between performance, weighting factors, sampling interval, optimality and length of

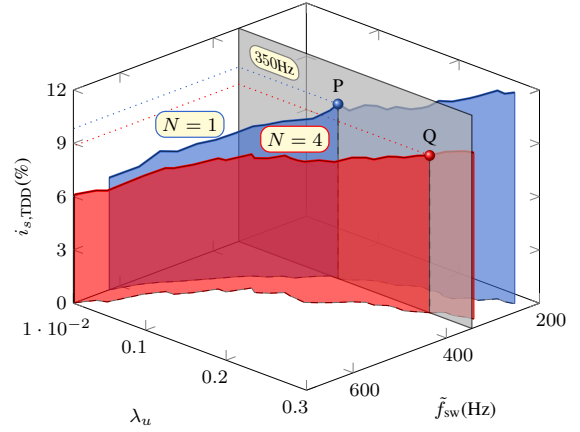


Fig. 7: Experimental sweep of the weighting factor  $\lambda_u$  for the prediction horizon  $N = 1$  and 4, and sampling interval  $T_s = 125 \mu\text{s}$  [17].

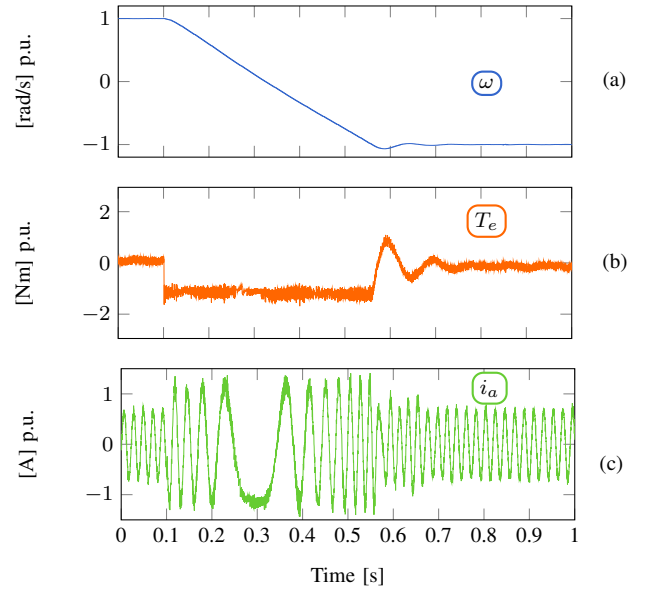


Fig. 8: A typical speed reversal maneuver using multi-step FCS-MPC [17].

the prediction horizon was obtained considering that transients are one of the main causes of impracticability due to the SDA initialization [73]–[75]. Thus, a selective initialization approach choose the box-constrained initialization during transients while preserving the well-known dynamic performance of FCS-MPC at minimum sub-optimality cost, without affecting the steady-state performance. Fig. 7 shows results of multi-step FCS-MPC reducing the stator current total demand distortion (TDD) ( $i_{s\text{TDD}}$ ), defined by:

$$i_{s\text{TDD}} = \frac{1}{\sqrt{2}I_{\text{nom}}} \sqrt{\sum_{h \neq 1} (\hat{i}_{s,h})^2}, \quad (19)$$

where  $I_{\text{nom}}$  refers to the nominal current of the IM and  $\hat{i}_{s,h}$  is the amplitude of the  $h$ -th stator current harmonic. The points P and Q in Fig. 7 are selected to demonstrate that for the same average switching frequency of  $f_{\text{sw}} = 350 \text{ Hz}$ ,

$i_{sTDD}$  is reduced by 10.6% for a prediction horizon  $N = 4$ . Finally, Fig. 8 shows a typical speed reversal maneuver, which demonstrates that the dynamic performance of the algorithm remains unaltered; see [17] for further details.

An FPGA implementation of the SDA is detailed in [19] achieving prediction horizons of up to five steps. Multi-step FCS-MPC is particularly beneficial for higher-order systems, such as converters with LC filters. As shown in [76], the current distortions can be reduced by an order of magnitude when using long prediction horizons of 15 to 20 steps.

## VI. MODEL PREDICTIVE PULSE PATTERN CONTROL

Optimal Pulse Patterns (OPP) allow the current distortion in the load to be minimized for a specific switching frequency [77]. Consequently, OPPs are a very attractive choice for medium-voltage drives because they achieve a substantial reduction of the switching frequency at comparable harmonic distortion [78]. The OPP's switching angles  $\alpha_j$  with  $j \in \{1, \dots, d\}$  are calculated offline and stored in the memory of the real-time control platform as functions of the modulation index  $m$  and the pulse number  $d$ , [79]. Fig. 9 shows the three-level OPP for  $m = 0.7$  and  $d = 5$ ; the corresponding stator flux reference trajectory is illustrated in Fig. 9(b).

The use of OPPs provides optimal performance under steady-state conditions. However, undesired transients are observed whenever the operating point is changed, or transitions between different pulse patterns occur. To overcome this problem, closed-loop control strategies have been proposed in the literature [21]–[23], [78]–[81]. Initial concepts use deadbeat controllers to track a current reference trajectory [80], [81] or stator flux reference trajectory to improve the robustness under parameter variations [78], [79]. However, these methods require a dedicated observer to identify the instantaneous fundamental components of the stator currents and flux linkage vectors. More recently, a model predictive controller, referred as Model Predictive Pulse Pattern Control (MP<sup>3</sup>C), has been introduced in [21]–[23] to control the instantaneous stator flux instead of its fundamental component. As a result, a simple and standard flux observer can be used, making the implementation much simpler.

### A. MP<sup>3</sup>C Control Strategy

Starting with the initial version in [21], the MP<sup>3</sup>C strategy has been extended and revised during the past decade [21]–[23], [82]–[89]. To explain the MP<sup>3</sup>C fundamentals, the three-level NPC converter will be used in this review. The block diagram of the overall MP<sup>3</sup>C scheme is shown in Fig. 10.

1) *Outer Control Loops*: As depicted in Fig. 10, the speed  $\omega$  is regulated by manipulating the reference torque  $T_e^*$ , which in turn is translated into the reference angle between the stator and rotor flux vectors  $\gamma^*$ , as illustrated in Fig. 9(c). Then, the angular position of the stator flux reference is determined as:

$$\theta^* = \angle \psi_r + \sin^{-1} \left( \frac{T_e^*}{k_T \Psi_s^* \Psi_r} \right), \quad (20)$$

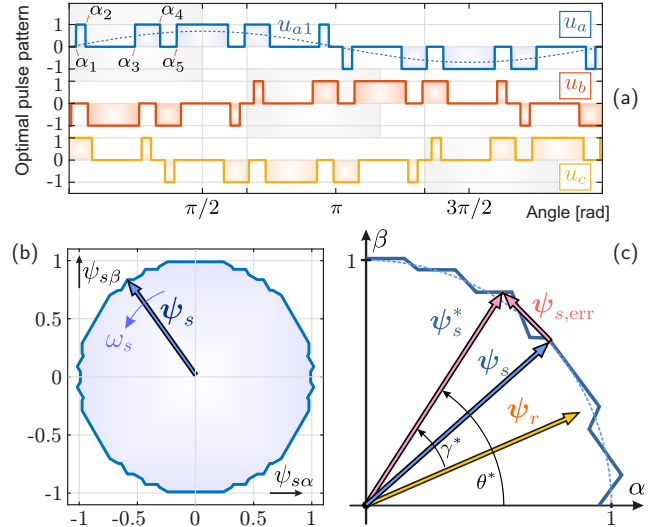


Fig. 9: (a) OPP for  $d = 5$  switching angles and  $m = 0.7$ ; (b) reference trajectory for the OPP shown in (a); (c) control principle of MP<sup>3</sup>C.

where  $\Psi_s^*$  is the desired stator flux magnitude,  $\Psi_r = \|\psi_r\|$  is the estimated rotor flux vector magnitude provided by the flux observer and the parameter  $k_T = \frac{3}{2} p \frac{k_r}{\sigma L_s}$ .

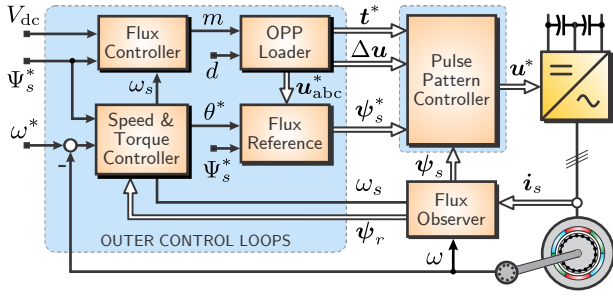
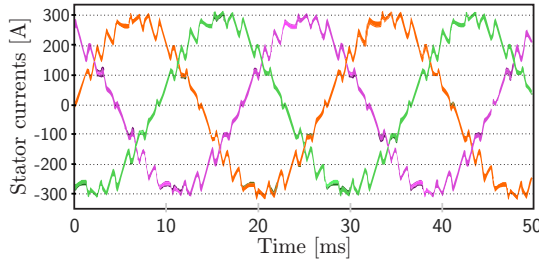
In its simplest form, the modulation index is determined by the flux controller from the angular stator frequency  $\omega_s = 2\pi f_s$  and the dc-link voltage  $V_{dc}$  as  $m = \omega_s \Psi_s^* (2/V_{dc})$ . The pulse number  $d$  is computed from the maximal allowed switching frequency of the semiconductor devices  $f_{sw}$  according to  $d = \text{floor}(f_{sw}/f_s)$ , where  $f_s$  is the fundamental stator frequency. Thus, by using the inputs  $d$  and  $m$ , the switching sequence of the OPP in the  $\alpha\beta$  frame can be stated as  $\mathbf{u}_s^* = \mathbf{T}_{\alpha\beta} \mathbf{u}_{abc}^*$ , where  $\mathbf{T}_{\alpha\beta}$  is the Clarke transformation. Then, if the induction machine is fed by the OPP, the stator flux space vector should track the following trajectory:

$$\psi_s^*(\theta) = \psi_s(0) + \frac{\Psi_s^*}{m} \int_0^\theta \mathbf{u}_s^*(\vartheta) d\vartheta. \quad (21)$$

Therefore, under ideal conditions, the stator voltage matches precisely the OPP voltage waveform  $\mathbf{v}_s = \mathbf{u}_s V_{dc}/2$ , and consequently, the stator current distortion is minimized. Nevertheless, the dc-link voltage ripple, voltage drop over the stator resistance, and the nonlinear effects of the inverter produce deviations of the stator flux space vector from its reference trajectory. Trajectory deviations also occur during transients or transitions between different pulse patterns, among other disturbances. Therefore, a closed-loop controller must be implemented to compensate these errors quickly and to ensure the proper operation of the drive.

To this end, an MPC strategy is designed, which operates at regularly-spaced sampling instants  $kT_s$ , where  $k \in \mathbb{N}$  and  $T_s$  is the sampling interval. A variable-length prediction horizon  $N$  is utilized such that this time interval includes at least one switching transition per phase. The switching instants of the pulse pattern are shifted forward or backwards such that the stator flux error is corrected within the horizon  $N$  to achieve fast closed-loop control of the stator flux space vector.




 Fig. 10: MP<sup>3</sup>C scheme for induction machines.

 Fig. 11: Medium-voltage experimental results: steady-state stator currents when operating with MP<sup>3</sup>C at 30 Hz with  $d=5$ .

2) *Pulse Pattern Controller*: Modifying the switching instants of the OPP, the following stator flux correction is obtained:

$$\Delta\psi_s(\Delta t) = -\frac{V_{dc}}{2} \mathbf{T}_{\alpha\beta} \mathbf{N} \Delta t, \quad (22)$$

with  $\mathbf{N} = \text{diag}\{\Delta\mathbf{u}^T\}$  and  $\Delta\mathbf{u} = [\Delta\mathbf{u}_a^T \ \Delta\mathbf{u}_b^T \ \Delta\mathbf{u}_c^T]^T$ , where  $\Delta\mathbf{u}_x \in \{-1, 1\}^{n_x}$  comprises the  $n_x$  switching transitions for the phase  $x \in \{a, b, c\}$  within the prediction horizon. The correction of the switching instants  $\Delta\mathbf{t} = \mathbf{t} - \mathbf{t}^*$  contains all modifications from the OPP's nominal switching times in the vector  $\mathbf{t}^* = [t_a^{*T} \ t_b^{*T} \ t_c^{*T}]^T$ . Thus, by using  $\Delta\mathbf{t}$  as the decision variable, the MP<sup>3</sup>C problem is formulated as:

$$\begin{aligned} \min_{\Delta\mathbf{t}} \quad & \|\psi_{s,\text{err}} - \Delta\psi_s(\Delta\mathbf{t})\|_2^2 + q\|\Delta\mathbf{t}\|_2^2 \\ \text{s.t.} \quad & \mathbf{t} = \Delta\mathbf{t} + \mathbf{t}^* \in \mathbb{T}, \end{aligned} \quad (23)$$

where  $\psi_{s,\text{err}} = \psi_s^* - \psi_s$  is the instantaneous stator flux error. The flux reference  $\psi_s^*$  is obtained by substituting the reference stator flux angle (20) in (21). Notice that vectors  $\mathbf{t}^*$  and  $\Delta\mathbf{u}$  are determined by the OPP, and consequently, they are provided by the OPP Loader block in Fig. 10, and then used as parameters in (23). To ensure that the MP<sup>3</sup>C-based switching instants  $\mathbf{t}^* = \Delta\mathbf{t} + \mathbf{t}^*$  occur within the prediction horizon  $N$  in ascending order but before the first switching transition beyond  $N$  (transition with the nominal time  $t_{x[n_x+1]}^*$ ), the set constraint  $\mathbb{T}$  in (23) is defined as  $\mathbb{T} = \mathbb{T}_a \times \mathbb{T}_b \times \mathbb{T}_c$ , where for each phase

$$\mathbb{T}_x = \left\{ \mathbf{t}_x \in \mathbb{R}^{n_x} \mid 0 \leq t_{x1} \leq \dots \leq t_{x[n_x]} \leq t_{x[n_x+1]}^* \right\}. \quad (24)$$

Solving (23) provides a trade-off between the flux error correction and modification to the nominal OPP. The scalar  $q > 0$  is used to adjust this compromise.

In summary, the MP<sup>3</sup>C problem in (23) is a small-scale

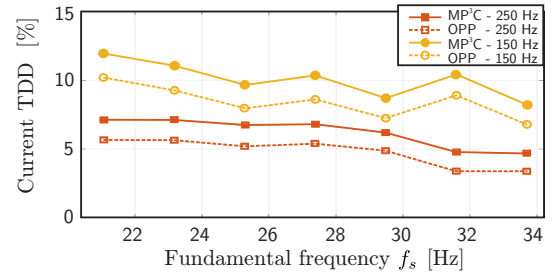


Fig. 12: Medium-voltage experimental results: measured vs theoretical OPP current TDD at rated torque when varying the fundamental frequency, [85].

quadratic program that can be solved in real-time by adopting the active set method or a fast gradient method [83], [90]. Its solution provides a sequence of optimal control actions within the horizon and, following the receding horizon policy, only the first control action of this sequence (the pulse pattern over  $T_s$ ) is applied to the drive system. Noteworthy extensions of the trajectory control principle have been developed, such as the optional insertion of additional switching transitions to shorten torque transients [23]. The pulse insertion provides the controller, when required, with an additional degree of freedom to remove the flux error as quickly as possible. On the other hand, to address the balancing of the neutral point (NP) potential of an NPC converter in the MP<sup>3</sup>C strategy, the cost function (23) is extended to also control the predicted error of the NP voltage at the end of the prediction horizon to zero [88]. As a result, the stator flux trajectory control and NP balancing are treated in a single control loop by modifying the switching instant of the OPP.

## B. Experimental Results

Experimental results for a medium-voltage NPC inverter driving a 3.3 kV induction machine rated at 1140 kVA are summarized in this review (further details can be found in [85]). To evaluate the performance of the drive system, the current TDD is utilized.

The stator current waveforms are depicted in Fig. 11, showing operation with MP<sup>3</sup>C at a fundamental frequency of 30 Hz and 60% of the rated machine torque. For a pulse number equal to 5, the resulting switching frequency is 150 Hz and the current TDD is 8.7%.

The current TDD is depicted in Fig. 12 for two upper bounds on the switching frequency (150 and 250 Hz) when the drive operates at various fundamental frequencies and rated torque. The current TDD that results from the MP<sup>3</sup>C strategy (solid lines) is compared with the theoretical one from the OPP at open-loop (dash-dotted line). As shown in Fig. 12, the difference between the measurement and the theoretical TDD is less than 1.85%. This discrepancy tends to increase with the torque set-point. Thus, the results depicted in Fig. 12 correspond to the worst-case at which the IM operates, i.e., at rated torque.



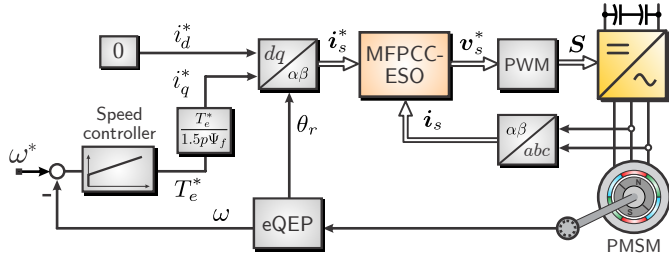


Fig. 13: Control diagram of the PMSM drive system with an ESO based on an ultra-local model.

## VII. MODEL-FREE MPC

The performance of MPC will be degraded if the system parameters used in the controller are different from the actual system parameters [91]. For example, the stator resistance and  $dq$ -axis inductance of a Permanent Magnet Synchronous Machine (PMSM) may vary with the operating point and environment. If the machine parameter variations are not taken into account, the current will show a steady-state error and the noise will increase during the operation of the machine [92].

Various methods have been proposed in the literature to tackle the problems of machine parameter variations, including online parameter identification [93], disturbance observer-based methods [94], and model free control [28], [95]–[98]. These methods differ in operating principle, complexity and computational burden. However, one of the main differences is that the online identification and disturbance observer-based methods are still using the mathematical model of machines, while model free control does not use the actual machine models. This feature distinguishes model free control from other kinds of methods due to its universality and flexibility.

Recently, model free control has been introduced in MPC and is gaining increasing attention. The early work on model free MPC in [95] simply uses the current difference in the past to predict the future current. Although the principle is simple, it requires high sampling frequency and the performance is affected by the updating rate of current difference table. Furthermore, the use of only one voltage vector during one control period limits its steady state performance.

The other work on model free MPC uses the so-called ultra-local model developed by Michel Fliess [99]. As pointed in [99], in most cases the physical system can be described using the first/second order ultra-local model. For example, the PMSM can be described by a first-order ultra-local model using complex vector in stationary frame as [97]:

$$\frac{d\mathbf{i}_s}{dt} = \alpha\mathbf{v}_s + \mathbf{F}, \quad (25)$$

where  $\mathbf{F}$  represents the known and unknown part of the system, which can be summarized as a total disturbance;  $\alpha$  represents a nonphysical scaling factor selected by the designer;  $\mathbf{v}_s$  and  $\mathbf{i}_s$  are the stator voltage vector (input) and stator current vector (output), respectively. According to [99], the total disturbance  $\mathbf{F}$  can be estimated from the input and

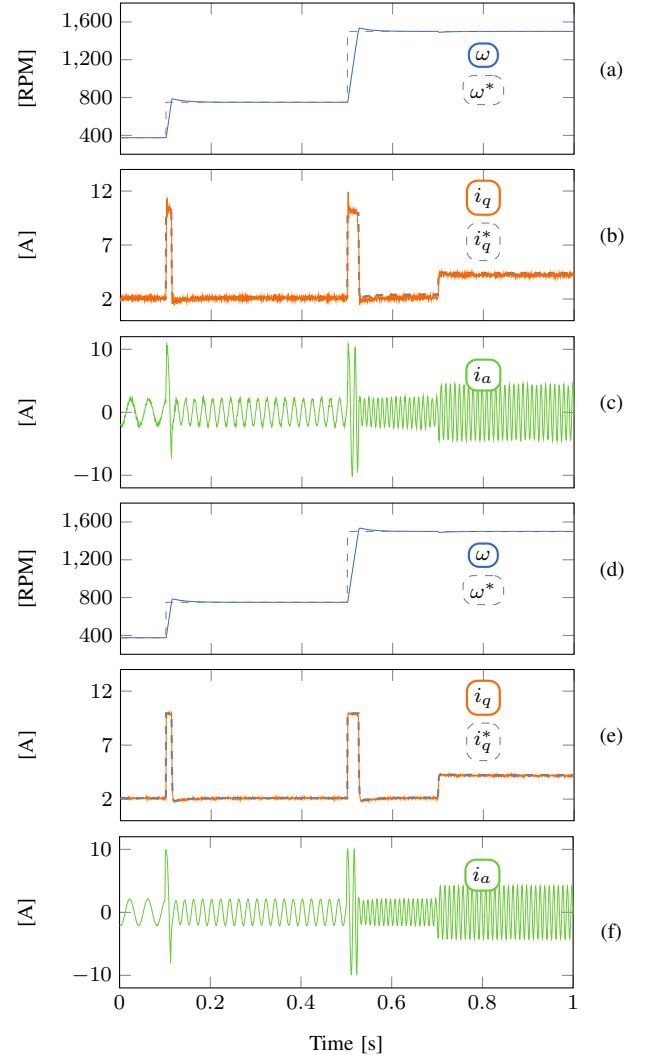


Fig. 14: Experimental results of model-free MPC during a dynamic process with mismatched parameters at  $1R$ ,  $2\psi$ ,  $2L$  under the condition of 10 kHz sampling frequency, (a)-(b)-(c)  $\mathbf{F}$  estimated using differential algebra method, (d)-(e)-(f)  $\mathbf{F}$  estimated using ESO [97].

output using a differential algebra method as

$$\hat{\mathbf{F}} = -\frac{6}{L} \int_{t-L}^t (L-2\sigma)\mathbf{i}_s(\sigma) + \alpha\sigma(L-\sigma)\mathbf{v}_s(\sigma)d\sigma \quad (26)$$

where  $L = n_F T_{sc}$  and  $n_F$  is the number of control periods in the integral step  $L$ . Owing to the existence of  $n_F$ , a lot of past information is used in the differential algebra method. This will affect the dynamic response of the system. In [97], an Extended State Observer (ESO) is employed to observe the total disturbance. However, it should be noted that the ESO is still based on the ultra-local model rather than real machine model. After estimating  $\mathbf{F}$ , one can track the stator current reference using deadbeat control based on SVM or finite-control-set MPC. The control diagram of the PMSM drive system with an ESO based on ultra-local model is shown in Fig. 13.

The dynamic responses of a PMSM drive based on model

TABLE I: Main characteristics of MPC strategies.

	MPCC	MPTC	SMPC	Multi-Step	MP <sup>3</sup> C	Model-Free
Need of weighting factors	✓✓	✗	✓✓	✗	✓✓	✗
Harmonic distortion	✓	✓	✗	✓	✓✓	✓
Low switching frequency	✗	✗	✗✗	✓✓	✓✓	—
Calculation and complexity	✓	✓	✓✓	✗	✓	✗✗
Parameter robustness	✓	✓	✓	✗	✓	✓✓

Scaled from best (✓✓) to worst (✗✗), (—) not studied.

free MPC is illustrated in Fig. 14, where  $\mathbf{F}$  is estimated using differential algebra method in (a)-(c) and using ESO in (d)-(f). From top to bottom, the curves shown in Fig. 14 are the rotor speed, the reference and the actual values of the  $q$ -axis current, and the  $a$ -phase stator current. The permanent magnet flux and inductance in the controller are twice of the actual value. It is seen that model-free MPC can achieve accurate tracking of  $q$ -axis current with machine parameter mismatches. It is worth to highlight that, by estimating  $\mathbf{F}$  using ESO instead of differential algebra method, much lower dynamic overshoot is achieved in the current.

### VIII. CHALLENGES AND FUTURE WORK

MPC has shown a high degree of flexibility, allowing the control of torque, flux, and currents, generating different control schemes that are easy to implement. Table I summarizes the main characteristics of the different MPC strategies described in the previous chapters. According to the experience of these researchers, MPC still has to overcome some challenges in terms of behavior. One of the challenges is the operation at very low speed, including developing full torque at zero speed. In the case of sequential MPC, it is important to gain more practical experience concerning the selection of the sequence: first selecting the torque and then the flux or vice versa is an aspect that needs more maturity. Also, the number of voltage vectors that are chosen at each stage of the sequence is a matter that needs further investigation.

To reduce the intrinsic dependence that MPC has on the motor parameters, Model-Free predictive control appears as an incipient and very attractive alternative, which needs further investigation regarding the implementation, observation techniques, and experimental results. In particular, the performance to parameter mismatch of Model-Free predictive control should be compared to the one obtained using field-oriented control (with linear control).

Another future research is the study of MP<sup>3</sup>C for higher order systems, such as drives with LC output filter. However, the optimal dynamic response using multi-step prediction should be subject of additional study.

Finally, Artificial Intelligence (AI) techniques are being considered to be applied in power converters [100]. In principle, MPC is appropriate for AI techniques because it works with basic models and instantaneous variables. However, this is a new line of research that must be explored in the future.

### IX. CONCLUSIONS

This paper showed that Model Predictive Control is being used successfully in high-performance motor drive applica-

tions. With little effort, MPC has been applied in field-oriented control and direct control of flux and torque.

Two different and interesting strategies have solved the classical problem of weighting factor calculation in the cost function when using predictive control. The first method uses sequential predictive control to avoid the use of weighting factors. This solution is straightforward to understand and implement without sacrificing the drive's high-quality transient behavior. The second strategy uses Artificial Neural Networks to obtain the optimal value for the weighting factor, introducing artificial intelligence techniques in the core of the control method, which opens a very attractive area for future research.

The use of a multiple-step prediction has reduced the distortion of the load current generated by the inverter. Besides, for higher power motor drives, optimized pulse patterns can be integrated with predictive control to further reduce the distortion in the load current while operating with a low switching frequency.

Finally, the model-free strategy has demonstrated that it is possible to control the machine with high quality, without the need for a precise model, which offers a significant opportunity to improve the drive's robustness, introducing modern estimation techniques into the control algorithm.

These selected advanced topics showed that the application of Model Predictive Control opens the possibility to continue improving the behavior of high-performance motor drives, including more intelligent techniques and advanced optimization algorithms.

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**Jose Rodriguez** (M'81-SM'94-F'10-LF'20) received the Engineer degree in electrical engineering from the Universidad Tecnica Federico Santa Maria, in Valparaíso, Chile, in 1977 and the Dr.-Ing. degree in electrical engineering from the University of Erlangen, Erlangen, Germany, in 1985. He has been with the Department of Electronics Engineering, Universidad Tecnica Federico Santa Maria, since 1977, where he was full Professor and President. Since 2015 he was the President and since 2019 he is full professor at Universidad Andres Bello in

Santiago, Chile. He has coauthored two books, several book chapters and more than 400 journal and conference papers. His main research interests include multilevel inverters, new converter topologies, control of power converters, and adjustable-speed drives. He has received a number of best paper awards from journals of the IEEE. Dr. Rodriguez is member of the Chilean Academy of Engineering. In 2014 he received the National Award of Applied Sciences and Technology from the government of Chile. In 2015 he received the Eugene Mittelmann Award from the Industrial Electronics Society of the IEEE. In years 2014 to 2020 he has been included in the list of Highly Cited Researchers published by Web of Science.



**Cristian Garcia** (M'15) received the M.Sc. and Ph.D. degrees in electronics engineering from the Universidad Tecnica Federico Santa Maria, Valparaíso, Chile, in 2013 and 2017, respectively. From 2017 to 2019, he was with the Engineering Faculty of the Universidad Andres Bello, Santiago, Chile, as an Assistant Professor. Since 2019, he has been with the Department of Electrical Engineering of the University of Talca, Curico, Chile, where he is currently an Assistant Professor. During 2016 he was a visiting Ph.D. student in the Power Electronics

Machines and Control (PEMC) Group at the University of Nottingham, U.K. His research interests include electric transportation applications, variable-speed drives and model predictive control of power converters and drives.



**Andrés Mora** (S'14–M'17) received the B.Sc. and M.Sc degrees in electrical engineering from the Universidad Técnica Federico Santa María (UTFSM), Valparaíso, Chile, in 2010, and the Ph.D. degree in electrical engineering from the Universidad de Chile, Santiago, Chile, in 2019.

Since 2011, he has been an Assistant Professor with the Department of Electrical Engineering at the UTFSM. His research interests include multi-level power converters, variable speed drives, model predictive control, and renewable energy conversion systems. In 2019, Dr. Mora received the IEEE Transactions on Industrial Electronics Best Paper Awards.



**Freddy Flores** (Member, IEEE) received the M.Sc. and Ph.D. degrees in electronics engineering from Universitat Rovira I Virgili (URV), Tarragona, Spain, in 2009 and 2013, respectively.

In 2015 he joined to the Advanced center for electrical and electronic engineering (AC3E), from the Universidad Técnica Federico Santa María (UTFSM), Valparaíso, Chile, as a postdoctoral fellow. Consequently, in 2017 he was in charge of the energy area in the technology transfer unit of the AC3E developing and managing industrial projects related to the energy and electric power systems. Dr. Flores-Bahamonde is currently Assistant Professor in the Engineering Sciences Department at the Universidad Andrés Bello, where he is also researcher in the Center for Energy Transformation.

His main interest includes the design and control of power converters for renewable energies, automotive power systems, and DC microgrids.



**Pablo Acuna** (M'12) received the B.Sc. degree in electronics engineering, the B.Eng. degree in electronics engineering, and the Ph.D. degree in electrical engineering from the University of Concepción, Concepción, Chile, in 2004, 2007, and 2013, respectively. From 2014 to 2017, he was a Research Associate in the University of New South Wales, Australia.

From 2018 to 2019, he was Lecturer in the School of Engineering, RMIT University, Melbourne, Australia. He is currently a Assistant Professor in the Department of Electrical Engineering, Universidad de Talca, Chile.

His research interests include electrical power conversion systems and their applications in industry, transportation, and utility.





**Mateja Novak** (S'17-M'21) received the M.Sc. degree in Electrical Engineering and Information Technology from Zagreb University, Croatia, in 2014 and the Ph.D. degree in Electrical Engineering from Aalborg University, Denmark, in 2020. She is currently working as a postdoctoral researcher at AAU Energy, Aalborg University, Denmark. Her research interests include model predictive control, multilevel converters, statistical model checking, reliability of power electronic systems and renewable energy systems.



**Yongchang Zhang** (M'10-SM'18) received the B.S. degree from Chongqing University, China, in 2004 and the Ph.D. degree from Tsinghua University, China, in 2009, both in electrical engineering. From August 2009 to August 2011, he was a Post-doctoral Fellow at the University of Technology Sydney, Australia. He joined North China University of Technology in August 2011 as an associate professor, and was promoted to a full professor in January 2015. Since August 2021, he is a full professor at North China Electric Power University.

He has published more than 100 technical papers in the area of motor drives, pulsewidth modulation and AC/DC converters. His current research interest is model predictive control for power converters and motor drives.

Dr. Zhang is a fellow of the Institute of Engineering and Technology. He serves as the associate/guest editor of several international journals, such as IEEE Journal of Emerging and Selected Topics in Power Electronics. He is the Technical Program Co-Chair of 5th/6th IEEE International Conference on Predictive Control of Electrical Drives and Power Electronics



**Luca Tarisciotti** (S'12-M'15-SM'21) received the master's degree in Electronic Engineering from The University of Rome "Tor Vergata" in 2009 and his Ph.D. degree in Electrical and Electronic Engineering from the PEMC group, University of Nottingham in 2015. In the same year he became Research Fellow at the University of Nottingham, UK, until 2018.

He is currently working as Assistant Professor at the University Andres Bello, Santiago, Chile. His research interests include Matrix converters, DC/DC

converters, Multilevel converters Advanced power electronics and Smart Grids control.



**S. Alireza Davari** (SM'17) was born in Tehran, Iran, in 1981. He received the M.Sc. and Ph.D. degrees both from the Iran University of Science and Technology (IUST), Tehran, Iran, in 2006 and 2012, respectively.

Between 2010 and 2011, he left for a sabbatical visit at Technische Universitaet Muenchen, Germany. Between 2013 and 2020 he was at Shahid Rajaei Teacher Training University as an Assistant Professor. Since 2020 he has continued at the same university as an Associate Professor.

His research interests include encoder-less drives, predictive control, power electronics, and renewable energy.



**Zhenbin Zhang** (Senior Member, IEEE) was born in Shandong, China, in 1984. He received his PhD in electrical engineering in 2016 (summa cum laude) from the Institute for Electrical Drive Systems and Power Electronics (EAL), Technical University of Munich (TUM), Munich, Germany.

From 2016 to 2017, he was a Research Fellow and the Group-Leader for "Modern Control Strategies for Electrical Drives" Group in EAL of TUM. Since 2017, he has been a Full Professor and Laboratory Director with Shandong University, Jinan, China. Since 2020, he has been the Director of the International Center for Intelligent Energy and Power Conversion Systems, Shandong University. His research interests include power electronics and electrical drives, sustainable energy systems, and smart- and microgrids.

Dr. Zhang received the VDE-Award, Suedbayern, Germany, in 2017. He is an Associate Editor for the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS.



**Fengxiang Wang** (S'13-M'14-SM'18) was born in Jiujiang, China, in 1982. In 2014 he received Ph.D. degree at the Institute for Electrical Drive Systems and Power Electronics, Technische Universitaet Muenchen, Munich, Germany.

Currently he is working as full professor and deputy director of Quanzhou Institute of Equipment Manufacturing, Haixi Institutes, Chinese Academy of Sciences, China.

He serves as Associate Editor for IEEE Transactions on Industrial Electronics. His research interests include predictive control and sensorless control for electrical drives and power electronics.



**Margarita Norambuena** (S'12-M'14-SM'21) received the B.S. and M. Sc. degrees in electric engineering from the Universidad Tecnica Federico Santa Maria (UTFSM), Valparaiso, Chile, in 2013. She received the Ph.D. degree (summa cum laude) in electronics engineering from the UTFSM in 2017 and received the Doktoringenieur (Dr-Ing.) degree (summa cum laude) from the Technische Universitaet Berlin (TUB) in 2018.

She received the IEEE IES Student Best Paper Award in 2019 for her doctoral work. Currently she

is an Assistant Professor at Universidad Tecnica Federico Santa Maria. Her research interest include multilevel converters, model predictive control of power converters and drives, energy storage systems, renewable energy and electromobility. Dr. Norambuena serves as an Associate Editor for IEEE JESTPE.



**Tomislav Dragicevic** (S'09-M'13-SM'17) received the M.Sc. and the industrial Ph.D. degrees in Electrical Engineering from the Faculty of Electrical Engineering, University of Zagreb, Croatia, in 2009 and 2013, respectively. From 2013 until 2016 he has been a Postdoctoral researcher at Aalborg University, Denmark. From 2016 until 2020 he was an Associate Professor at Aalborg University, Denmark. Currently, he is a Professor at the Technical University of Denmark.

He made a guest professor stay at Nottingham University, UK during spring/summer of 2018. His research interest is application of advanced control, optimization and artificial intelligence inspired techniques to provide innovative and effective solutions to emerging challenges in design, control and diagnostics of power electronics intensive electrical distributions systems and microgrids. He has authored and co-authored more than 330 technical publications (more than 150 of them are published in international journals, mostly in IEEE), 10 book chapters and a book in this field, as well as filed for several patents.

He serves as an Associate Editor in the IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS, in IEEE TRANSACTIONS ON POWER ELECTRONICS, in IEEE Emerging and Selected Topics in Power Electronics and in IEEE Industrial Electronics Magazine. Prof. Dragičević is a recipient of the Končar prize for the best industrial PhD thesis in Croatia, a Robert Mayer Energy Conservation award, and he is a winner of an Alexander von Humboldt fellowship for experienced researchers.



**Tobias Geyer** recently joined ABB Medium-Voltage Drives in Switzerland as the R&D platform manager of the ACS6080. He is also an extraordinary Professor at Stellenbosch University, South Africa. Dr. Geyer received the Dipl.-Ing. degree in electrical engineering, the Ph.D. in control engineering and the Habilitation degree in power electronics from ETH Zurich in the years 2000, 2005 and 2017, respectively.

After his Ph.D., he spent three years at GE Global Research, Munich, Germany, three years at the University of Auckland, Auckland, New Zealand, and eight years at ABB's Corporate Research Centre, Baden-Dättwil, Switzerland, where his last position was that of a Senior Principal Scientist for power conversion control.

He is the author of 35 patent families and the book "Model predictive control of high power converters and industrial drives" (Wiley, 2016). He teaches a regular course on model predictive control at ETH Zurich. His research interests include medium-voltage and low-voltage drives, utility-scale power converters, optimized pulse patterns and model predictive control.

Dr. Geyer received the Semikron Innovation Award and the Nagamori Award, both in 2021. He is also the recipient of the 2017 First Place Prize Paper Award in the Transactions on Power Electronics, the 2014 Third Place Prize Paper Award in the Transactions on Industry Applications, and of two Prize Paper Awards at conferences. He is a former Associate Editor for the Transactions on Industry Applications (from 2011 until 2014) and the Transactions on Power Electronics (from 2013 until 2019). He was an international program committee vice chair of the IFAC conference on Nonlinear Model Predictive Control in Madison, WI, USA, in 2018. Dr. Geyer is a Distinguished Lecturer of the Power Electronics Society in the years 2020 and 2021.



**Frede Blaabjerg** (S'86-M'88-SM'97-F'03) was with ABB-Scandia, Randers, Denmark, from 1987 to 1988. From 1988 to 1992, he got the PhD degree in Electrical Engineering at Aalborg University in 1995. He became an Assistant Professor in 1992, an Associate Professor in 1996, and a Full Professor of power electronics and drives in 1998. From 2017 he became a Villum Investigator. He is honoris causa at University Politehnica Timisoara (UPT), Romania and Tallinn Technical University (TTU) in Estonia.

His current research interests include power electronics and its applications such as in wind turbines, PV systems, reliability, harmonics and adjustable speed drives. He has published more than 600 journal papers in the fields of power electronics and its applications. He is the co-author of four monographs and editor of ten books in power electronics and its applications.

He has received 33 IEEE Prize Paper Awards, the IEEE PELS Distinguished Service Award in 2009, the EPE-PEMC Council Award in 2010, the IEEE William E. Newell Power Electronics Award 2014, the Villum Kann Rasmussen Research Award 2014, the Global Energy Prize in 2019 and the 2020 IEEE Edison Medal. He was the Editor-in-Chief of the IEEE TRANSACTIONS ON POWER ELECTRONICS from 2006 to 2012. He has been Distinguished Lecturer for the IEEE Power Electronics Society from 2005 to 2007 and for the IEEE Industry Applications Society from 2010 to 2011 as well as 2017 to 2018. In 2019-2020 he served as a President of IEEE Power Electronics Society. He has been Vice-President of the Danish Academy of Technical Sciences.

He is nominated in 2014-2020 by Thomson Reuters to be between the most 250 cited researchers in Engineering in the world.



**Ralph 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 Universität 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, the EPE Association Distinguished Service Award in 2015 as well as the 2019 EPE Outstanding Achievement Award. In 2018 Dr. Kennel received the Doctoral degree honoris causa from Universitatea Stefan cel Mare in Suceava (Romania). 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.



**Davood Arab Khaburi** was born in 1965. He has received B.Sc. in Electronic Engineering, in 1990 from Sharif University of Technology, Tehran, Iran, and M.Sc. and Ph.D. in Electrical Engineering, from ENSEM, INPEL, Nancy, France in 1994 and 1998, respectively. He has joined to UTC in Compiègne, France for one year (1998-1999). Since January of 2000 he has been as a faculty member in Electrical Engineering Department of Iran University of Science Technology (IUST), where he is currently as an Associate Professor.

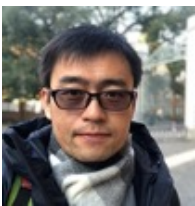
He is one of the founders of Iranian Association of Power Electronics and currently he is a board member of this association. He is also a member of Center Of Excellence for Power Systems Automation and Operation. Currently, he is the head of power group at IUST. His research interests are Power Electronics, Motor Drives and Digital Control.



**Mohamed Abdelrahem** (S'14-M'18-SM'21) was born in Assiut, Egypt, in 1985. He received the B.Sc. (Hons.) and M.Sc. degrees in electrical engineering from Assiut University, Assiut in 2007 and 2011, respectively, and the Ph.D. degree (Hons.) in electrical engineering from the Technical University of Munich (TUM), Munich, Germany, in 2020.

Since 2019, he is the head of the research group "Renewable Energy Systems" at the Institute for Electrical Drive Systems and Power Electronics (EAL), TUM. Since 2020, he is an assistant professor at

the Electrical Engineering Department, Assiut University, Egypt. In 2020, Dr. Abdelrahem received Walter Gademann prize from faculty of Electrical and Computer Engineering, TUM, in recognition of his excellent PhD dissertation entitled "Predictive Control and Finite-Set Observers for Variable-Speed Wind Generators". Furthermore, he has received a number of best paper awards from high prestigious international conferences of the IEEE. Dr. Abdelrahem is recorded in the world's top 2% scientist's list by Stanford University. His research interests include power electronics, predictive and encoderless control of variable-speed wind generators, photovoltaic energy systems, and energy storage systems.



**Zhen Zhang** (S'11-M'13-SM'15) received the B.Eng. and M.Eng. degrees in automation from Tianjin University, Tianjin, China, in 2004 and 2007, respectively, and the Ph.D. degree in electrical engineering from The University of Hong Kong, Hong Kong, in 2014.

He is currently a Professor of Electrical Engineering with the School of Electrical and Information Engineering, Tianjin University. His research interests focus on the advanced control for power conversion with emphasis on wireless power transfer and motor

drives.

Dr. Zhang is currently the Chair of IEEE Beijing Section IES Chapter and an Associate Editor for the IEEE INDUSTRIAL ELECTRONICS MAGAZINE, IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS, and IEEE TRANSACTIONS ON INDUSTRIAL INFORMATICS. He is the recipient of the IEEE J. David Irwin Early Career Award, Humboldt Research Fellowship, and 2020 Best Paper Award for IEEE TRANSACTIONS ON INDUSTRIAL ELECTRONICS.



**Nenad Mijatovic** after obtaining his Dipl.Ing. education in Electrical Power Engineering at University of Belgrade, Serbia in 2007, was enrolled as a doctoral candidate at Technical University of Denmark. He received his Ph.D. degree from Technical University of Denmark for his work on technical feasibility of novel machines and drives for wind industry. Upon completion of his PhD, he continued work within the field of wind turbine direct-drive concepts as an Industrial PostDoc.

Dr. N. Mijatovic currently holds position of Associate Professor at Technical University of Denmark where he is in charge of managing research projects and education related to the field of electrical machines and drives, power electronic converters, motion control, application of energy storage and general applications of low frequency electromagnetism with main focus on emerging eMobility and renewable energy generation. He is a member of IEEE since 2008 and senior member of IEEE since 2018 and his field of interest and research includes novel electrical machine drives/actuator designs, operation, control and diagnostic of electromagnetic assemblies, advance control of drives and grid connected power electronics, energy storage and eMobility.



**Ricardo Aguilera** (S'01-M'12) received the B.Sc. degree in electrical engineering from the Universidad de Antofagasta, Antofagasta, Chile; the M.Sc. degree in electronics engineering from the Universidad Tecnica Federico Santa Maria, Valparaíso, Chile; and the Ph.D. degree in electrical engineering from The University of Newcastle (UoN), Newcastle, NSW, Australia, in 2003, 2007, and 2012, respectively.

From 2012 to 2013, he was a Research Academic at UoN, where he was part of the Centre for Complex Dynamic Systems and Control. From 2014 to 2016, he was a Senior Research Associate at The University of New South Wales, Australia, where he was part of the Australian Energy Research Institute. Since September 2016, he has been with the School of Electrical and Data Engineering, at the University of Technology Sydney, Australia, where he currently holds a Senior Lecturer position. His main research interests include theoretical and practical aspects on model predictive control with application to power electronics, renewable energy integration, and microgrids.