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Linear Parameter Varying Control of Doubly Fed Induction Machines

H. Nguyen Tien, Carsten W. Scherer, *Fellow, IEEE*, Jacquélien M. A. Scherpen, *Senior Member, IEEE*, and Volkmar Müller

Abstract—This paper is concerned with the design of a self-scheduled current controller for doubly fed induction machines. The design is based on the framework of linear parameter-varying systems where the mechanical angular speed is considered to be a measurable time-varying parameter. The objective is to obtain robust dynamic performance for all variations of the mechanical rotor speed in a specified operating range. Employing the benefits of digital control, a continuous-to-discrete time transformation is utilized to replace the continuous-time current controller by an approximating discrete-time controller. Experimental results are provided to demonstrate the performance and robustness of the controlled system with respect to changes of the mechanical rotor speed and against stator voltage dips.

Index Terms—Doubly fed induction machine, linear parameter varying control, robustness.

NOMENCLATURE

v_{sd}, v_{sq}	d and q components of stator voltage.
v_{rd}, v_{rq}	d and q components of rotor voltage.
i_{sd}, i_{sq}	d and q components of stator current.
i_{rd}, i_{rq}	d and q components of rotor current.
Ψ_{sd}, Ψ_{sq}	Stator flux components.
L_m	Mutual inductance.
L_s, L_r	Stator and rotor inductances.
R_s, R_r	Stator and rotor resistances.
$\sigma = 1 - \frac{L_m^2}{L_s L_r}$	Total linkage coefficient.
$T_s = \frac{L_s}{R_s}$	Time constant of stator.
$T_r = \frac{L_r}{R_r}$	Time constant of rotor.

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Fig. 1. Wind energy conversion system.

ω_s Electrical angular speed of stator.
 ω_m Mechanical angular speed of rotor.

I. INTRODUCTION

DOUBLY FED induction machines (DFIMs) are widely used as generators for wind energy conversion systems in which the rotor of the DFIM is connected to the grid via the rotor side converter and the grid side converter coupled by a DC-voltage link as shown in Fig. 1. This configuration allows a reduction of the converter cost since they need to handle only about 30% of the generator power over a wide range of rotor speeds.

In the literature, conventional controller designs for DFIMs as described in [1]–[4] are based on field or vector control. These control strategies rely on the assumption that the machine parameters are all time-invariant. In some cases, feedforward terms are added to the output of the rotor current controllers for compensating cross coupling terms in the rotor current equations. Obviously, the performance for the designed current controllers could be very sensitive to any variations of parameter values, and a careful tuning is required to guarantee good performance over the whole operating range of the rotor speed.

Hence, various techniques have been proposed as alternatives to these vector control schemes, such as direct power control (DPC) [5] and direct torque control (DTC) [6], [7]. As a typical

feature, DPC does not require an inner current controller. This leads to a simplified control configuration and allows to obtain high dynamic responses. As a disadvantage, such control schemes produce large ripples in the active and reactive power at steady state. This issue has been addressed in [8]–[11]. In [8], all feasible and smooth connections between two adjacent vector sequences were taken into account in the framework of a novel three-vector-based strategy for a predictive DPC to achieve both active power and reactive power ripple reduction with simple calculations. In [9], a combination of DPC and vector control is considered to result in a high dynamic response and lower power ripples. However, there were no experimental results showing the robustness of the controlled system with respect to rotor speed variations in the operation range and low voltage ride through capabilities.

Additionally, the so-called finite response time (FRT or deadbeat-like) controllers in [3], [4] are designed under the assumption that the mechanical angular speed is constant during each sampling period. In every sampling cycle, their coefficients are determined under the assumption that the DFIM model is linear time-invariant, and this computation is repeated if a different value of the angular speed is measured. The main drawbacks of this approach are the high on-line computational load and the fact that the adjustment of the FRT controller parameters is rather ad-hoc. As a result, the performance of the system cannot be guaranteed over the entire operating range of the DFIM. The linear parameter-varying (LPV) control approach of [12], [13] provides a key solution to this problem, since the controller adaption is tailored toward obeying stability and performance for all admissible trajectories of the on-line measurable mechanical angular speed. However, it is emphasized that no digital implementation of the LPV design and no experimental results are reported in [12], [13].

In this paper, the variation of the mechanical angular speed is also taken into account in the framework of an LPV controller synthesis to guarantee a bound on the worst-case energy gain for the specified operation range. Unlike [12], [13], we also treat the stator voltage as an input disturbance to achieve a reduction of the effect of grid voltage variations on the system, especially in the case of grid voltage dips. Furthermore, the implementation of an equivalent discrete-time version of the designed LPV controller is presented, along with experimental results that demonstrate the performance and robustness of the resulting digital control system. This is accompanied by an experimental comparison of the performance of the LPV control with a conventional FRT design.

A. Modeling of Doubly Fed Induction Machine

In a dq reference frame that has the d axis coinciding with the grid voltage vector, the DFIM model can be described by

$$\dot{x}_r = A_r(\omega_m)x_r + B_s v_s + B_r v_r \quad (1)$$

$$y_r = C_r x_r \quad (2)$$

where $x_r = (i_{rd} \ i_{rq} \ \Phi_{sd} \ \Phi_{sq})^T$, $v_s = (v_{sd} \ v_{sq})^T$, $v_r = (v_{rd} \ v_{rq})^T$, $y_r = i_r = (i_{rd} \ i_{rq})^T$ with $v_{sd}, v_{sq}, v_{rd}, v_{rq}, i_{sd}, i_{sq}, i_{rd}, i_{rq}$ denoting the voltage and current components of the

Fig. 2. Rotor side control block diagram.

stator and rotor, respectively, and Φ_{sd}, Φ_{sq} being the stator flux components; next to an obvious C_r we have

$$A_r(\omega_m) = \begin{pmatrix} a_{r11} & a_{r12} & a_{r13} & a_{r14} \\ -a_{r12} & a_{r11} & -a_{r14} & a_{r13} \\ a_{r31} & 0 & a_{r33} & a_{r34} \\ 0 & a_{r31} & -a_{r34} & a_{r33} \end{pmatrix}$$

$$B_s = \begin{pmatrix} -\frac{1-\sigma}{\sigma L_m} & 0 \\ 0 & -\frac{1-\sigma}{\sigma L_m} \\ 1 & 0 \\ 0 & 1 \end{pmatrix}, \quad B_r = \begin{pmatrix} \frac{1}{\sigma L_r} & 0 \\ 0 & \frac{1}{\sigma L_r} \\ 0 & 0 \\ 0 & 0 \end{pmatrix}$$

with $a_{r11} = -((a+1)/T_r) + (a/T_s)$, $a_{r12} = \omega_s - \omega_m$, $a_{r13} = a/L_m T_s$, $a_{r14} = -(a\omega_m/L_m)$, $a_{r31} = L_m/T_s$, $a_{r33} = -(1/T_s)$, $a_{r34} = \omega_s$; here L_s, L_r are the stator and rotor inductances, L_m is the mutual inductance, R_s, R_r are the stator and rotor resistances, $\sigma = 1 - (L_m^2/L_s L_r)$ is the total linkage coefficient and $a = (1 - \sigma)/\sigma$; moreover $T_s = L_s/R_s$ and $T_r = L_r/R_r$ denote the time constants of the stator and rotor, ω_m is the mechanical angular velocity of the rotor, and ω_s is the electrical angular velocity of the stator (or grid).

B. Rotor Side Control Structure

The control structure on the rotor side of the DFIM is shown in Fig. 2. The control objectives of the LPV current controller in the inner loop are to obtain high dynamic performance and robust tracking of the rotor currents. The proportional-integral (PI) controllers in the outer loops are used for tracking optimal values of the electrical torque T_e^{ref} and the active power Q_g^{ref} . Note that, in this configuration, the DC-link voltage is maintained at a constant value by the grid side controller.

The electrical torque of the DFIM can be calculated by

$$T_e = -\frac{3}{2} p \frac{L_m}{L_s} \Phi_{sq} i_{rd}. \quad (3)$$

When the stator voltage v_s is measured, the reactive power can be obtained as

$$Q_g = -3 v_{sd} \frac{\Phi_{sq} - i_{rq} L_m}{L_s}. \quad (4)$$

Equations (3) and (4) show that the electrical torque and the reactive power of the DFIM can be regulated via the components of the rotor currents i_{rd} and i_{rq} .

II. LPV CONTROLLER SYNTHESIS

Let us briefly recall the LPV control design paradigm from [14]. For this purpose consider a generic LPV-system

$$\begin{pmatrix} \dot{x}(t) \\ z_p(t) \\ y(t) \end{pmatrix} = \begin{pmatrix} A(\delta(t)) & B_p(\delta(t)) & B(\delta(t)) \\ C_p(\delta(t)) & D_p(\delta(t)) & E(\delta(t)) \\ C(\delta(t)) & F(\delta(t)) & 0 \end{pmatrix} \begin{pmatrix} x(t) \\ w_p(t) \\ u(t) \end{pmatrix} \quad (5)$$

where $A(\cdot)$, $B_p(\cdot)$, $B(\cdot)$, $C_p(\cdot)$, $D_p(\cdot)$, $E(\cdot)$, $C(\cdot)$, $F(\cdot)$ are affine functions of the parameter vector $\delta(t)$ that varies in the polytope δ_c with vertices $\delta^1, \dots, \delta^k$. Here, u is the control input and y is the measured output, while w_p is a (generalized) disturbance input and z_p denotes the controlled variable to be kept small. An LPV controller is described similarly as

$$\begin{pmatrix} \dot{x}_c(t) \\ u(t) \end{pmatrix} = \begin{pmatrix} A_c(\delta(t)) & B_c(\delta(t)) \\ C_c(\delta(t)) & D_c(\delta(t)) \end{pmatrix} \begin{pmatrix} x_c(t) \\ y(t) \end{pmatrix}. \quad (6)$$

The closed-loop system of (5) and (6) admits the description

$$\begin{pmatrix} \dot{\xi}(t) \\ z_p(t) \end{pmatrix} = \begin{pmatrix} \mathcal{A}(\delta(t)) & \mathcal{B}(\delta(t)) \\ \mathcal{C}(\delta(t)) & \mathcal{D}(\delta(t)) \end{pmatrix} \begin{pmatrix} \xi(t) \\ w_p(t) \end{pmatrix} \quad (7)$$

where the matrices $\mathcal{A}(\cdot)$, $\mathcal{B}(\cdot)$, $\mathcal{C}(\cdot)$, and $\mathcal{D}(\cdot)$ are given in (8), shown at the bottom of the page. The goal in LPV synthesis is to find a controller (6) such that the closed-loop interconnection (7) is exponentially stable and the L_2 -gain of $w_p \rightarrow z_p$ is bounded by a given number $\gamma > 0$ for all time-varying parameters with $\delta(t) \in \delta_c$. Guarantees for these properties of the closed-loop system (7) can be formulated as follows.

Theorem 1: If there exists some $\mathcal{X} \succ 0$ such that

$$\begin{pmatrix} * \\ * \\ * \end{pmatrix}^T \begin{pmatrix} 0 & \mathcal{X} & 0 & 0 \\ \mathcal{X} & 0 & 0 & 0 \\ 0 & 0 & -\gamma I & 0 \\ 0 & 0 & 0 & \frac{1}{\gamma} I \end{pmatrix} \begin{pmatrix} I & 0 \\ \mathcal{A}(\delta) & \mathcal{B}(\delta) \\ 0 & I \\ \mathcal{C}(\delta) & \mathcal{D}(\delta) \end{pmatrix} \prec 0 \quad (9)$$

holds for all $\delta \in \delta_c$, then (7) is uniformly exponentially stable and the L_2 -gain from w_p to z_p is bounded by $\gamma > 0$ for all parameter trajectories $\delta : [0, \infty) \rightarrow \delta_c$. ■

Since the system (5) is affinely dependent on the parameter $\delta \in \delta_c$, the state-space matrices of (5) range in the following polytope of generating system matrices [14]

$$\text{Co} \left\{ \begin{pmatrix} A(\delta^j) & B_p(\delta^j) & B(\delta^j) \\ C_p(\delta^j) & D_p(\delta^j) & E(\delta^j) \\ C(\delta^j) & F(\delta^j) & 0 \end{pmatrix} : j = 1, \dots, k \right\}. \quad (10)$$

If B , E and C , F are parameter independent, the closed-loop matrices in (7) are also affine in the parameter δ . Therefore, guaranteeing (9) for all $\delta \in \delta_c$ is equivalent to guaranteeing (9) for $\delta \in \{\delta^1, \dots, \delta^k\}$. Consequently, to ensure stability and performance for the controlled system, we need to find some $\mathcal{X} \succ 0$ with (9) for $\delta = \delta^1, \dots, \delta^k$; this boils down to solving a standard linear matrix inequality (LMI) problem.

By a well-known controller parameter elimination procedure, we arrive at the following LMI conditions that guarantee the existence of a polytopic LPV controller [14], [15]

$$\begin{pmatrix} Y & I \\ I & X \end{pmatrix} \succ 0 \quad (11)$$

$$\Phi_x^T \begin{pmatrix} XA(\delta^j) + A(\delta^j)^T X & XB_p(\delta^j) & C_p(\delta^j)^T \\ B_p(\delta^j)^T X & -\gamma I & D_p(\delta^j)^T \\ C_p(\delta^j) & D_p(\delta^j) & -\gamma I \end{pmatrix} \Phi_x \prec 0 \quad (12)$$

$$\Phi_y^T \begin{pmatrix} A(\delta^j)Y + A(\delta^j)^T Y & B_p(\delta^j) & YC_p(\delta^j)^T \\ B_p(\delta^j)^T & -\gamma I & D_p(\delta^j)^T \\ C_p(\delta^j)Y & D_p(\delta^j) & -\gamma I \end{pmatrix} \Phi_y \prec 0 \quad (13)$$

for $j = 1, \dots, k$; here Φ_x and Φ_y form bases for the null-spaces $\ker(B^T \ 0 \ E^T)$ and $\ker(C \ F \ 0)$, respectively.

After obtaining X and Y satisfying the LMIs (11)–(13), one can reconstruct $\mathcal{X}$ and the vertex controller matrices K_j at each vertex δ^j with the help of the projection lemma [16].

III. GAIN-SCHEDULING DESIGN FOR THE ROTOR SIDE CURRENT CONTROLLER

In this section we show how to apply the LPV control procedure in the framework of mixed sensitivity loop shaping H_∞ -design for the rotor current control loop.

A. Setting Up the LPV Model

The mechanical angular speed ω_m in (1) is considered as a time-varying parameter that can be measured on-line and varies by $\pm 30\%$ around the synchronous speed ω_s . Hence, we express the mechanical angular speed as $\omega_m(t) = \omega_s(1 + p_\omega \delta_\omega(t))$ with $-1 \leq \delta_\omega(t) \leq 1$ and $p_\omega = 0.3$. Then (1) can be rewritten in an affinely parameter dependent form as

$$\dot{x}_r = (A_{rs} + \delta_\omega(t)A_{r\omega})x_r + B_sv_s + B_rv_r \quad (14)$$

where $A_{rs} \triangleq A_r(\omega_s)$ and $A_{r\omega} \triangleq A_r(\omega_s(1 + p_\omega)) - A_{rs}$ are time-invariant matrices. This leads to the LPV system (14), (2) with $\delta_\omega(t) \in \text{Co}\{-1, 1\}$ that is denoted as G_{rn} .

$$\begin{pmatrix} \mathcal{A}(\delta) & \mathcal{B}(\delta) \\ \mathcal{C}(\delta) & \mathcal{D}(\delta) \end{pmatrix} = \begin{pmatrix} A(\delta) + B(\delta)D_c(\delta)C(\delta) & B(\delta)C_c(\delta) & B_p(\delta) + B(\delta)D_c(\delta)F(\delta) \\ B_c(\delta)C(\delta) & A_c(\delta) & B_c(\delta)F(\delta) \\ C_p(\delta) + E(\delta)D_c(\delta)C(\delta) & E(\delta)C_c(\delta) & D_p(\delta) + E(\delta)D_c(\delta)F(\delta) \end{pmatrix} \quad (8)$$

Fig. 3. Interconnection of the system.

B. H_∞ Control for the LPV System

In a first step and for frozen parameter values $\delta_\omega \in [-1, 1]$, we use the mixed sensitivity interconnection for controller synthesis in an H_∞ -framework as shown in Fig. 3. The generalized disturbance input w_{rc} consists of the stator voltages and reference rotor currents $w_{rc} = (v_{sd} \ v_{sq} \ i_{rd}^{ref} \ i_{rq}^{ref})^T$, while the controlled output is chosen as $i_r = (i_{rd} \ i_{rq})^T$. The tracking error $e_r = (e_{rcd} \ e_{rcq})^T = (i_{rd}^{ref} - i_{rd} \ i_{rq}^{ref} - i_{rq})^T$ is the input to the controller and its output is $v_r = (v_{rd} \ v_{rq})^T$. Note that the components of v_{sd} and v_{sq} of w_{rc} are viewed as external disturbances and their influence on the controlled outputs must be reduced as much as possible.

The weighting function $W_{rs} = \begin{pmatrix} W_{rsd} & 0 \\ 0 & W_{rsq} \end{pmatrix}$ is used to shape the sensitivity S_{rc} which corresponds to the transfer matrix from the reference input i_r^{ref} to the tracking error e_r . W_{rs} is kept large over the low frequency range for good tracking. On the other hand, the weighting function $W_{rt} = \begin{pmatrix} W_{rtd} & 0 \\ 0 & W_{rtq} \end{pmatrix}$ is used to shape the transfer matrix T_{rc} from the reference input i_r^{ref} to the controlled output i_r . The weight W_{rt} is not only chosen to keep the closed loop bandwidth at a desired value, but also to reject the effect of the components v_{sd} and v_{sq} onto the controlled output i_r .

According to the standard H_∞ control paradigm, the weights are tuned so as to achieve

$$\left\| \begin{bmatrix} W_{rs} S_{rc} \\ W_{rt} T_{rc} \end{bmatrix} \right\|_\infty \leq \gamma_{H_\infty}$$

for some stabilizing controller and with a minimal value of γ_{H_∞} . This specification should be achieved with a stabilizing LTI controller $K_{rc}(\delta_\omega)$ for all frozen values of $\delta_\omega \in [-1, 1]$.

C. Synthesis of Gain-Scheduled Current Controller

After having tuned the weights, the parameter $\delta_\omega(t) \in \delta_c$ is again considered to be varying with time. This permits to derive a description of the interconnection in Fig. 3 as in (5). It is easy to verify that all required properties are satisfied to apply the design procedure in Section II. The LMI Control Toolbox [17] is employed to minimize γ over the LMI constraints (11)–(13) with an optimal value γ_{LPV} (which is bounded from

below by γ_{H_∞}). This results in an exponentially stabilizing controller $K_{rc}(\delta_\omega(t))$ such that the $\mathcal{L}_2$ -gain of the channel $w_{rc} \rightarrow z_{rc}$ is bounded by γ_{LPV} for all time-varying parameter trajectories satisfying $\delta_\omega(t) \in \delta_c$. In terms of the two vertex controllers $(A_{Krcj}, B_{Krcj}, C_{Krcj}, D_{Krcj})$, $j = 1, 2$, the state-space matrices of the LPV controller at the measured speed $\omega_m(t)$ then equal

$$\begin{pmatrix} A(\delta_\omega(t)) & B(\delta_\omega(t)) \\ C(\delta_\omega(t)) & D(\delta_\omega(t)) \end{pmatrix} = \frac{1}{2} (1 - \delta_\omega(t)) \begin{pmatrix} A_{Krc1} & B_{Krc1} \\ C_{Krc1} & D_{Krc1} \end{pmatrix} + \frac{1}{2} (\delta_\omega(t) - 1) \begin{pmatrix} A_{Krc2} & B_{Krc2} \\ C_{Krc2} & D_{Krc2} \end{pmatrix}$$

with $\delta_\omega(t) = (\omega_m(t) - \omega_s)/\omega_s p_\omega$. This completes the design of the continuous-time LPV controller.

A similar design of a continuous-time LPV current controller for DFIMs has been also presented in [18].

IV. DISCRETE-TIME LPV CONTROLLER

A. Discretization Problem of Predesigned LPV Controllers

Unlike a direct digital design method, such as in [19], the designed continuous-time controller has to be replaced with an equivalent discrete-time controller to be implemented in an industrial digital control system. The transformation from continuous to discrete time requires that the sampling frequency is high so as to obtain a good digital approximation and to retain system stability and performance [20]. On the other hand, the maximal possible sampling frequency depends on the particular digital control system used for implementation and cannot be arbitrarily large. This motivates to consider the closed-loop performance for choosing a suitable sampling time. In [21], the input and the feedback gains are modified to push the response of the sampled-data model toward that of the original system. In [22], the stability of the resulting control system is guaranteed by minimizing the H_∞ -norm of the error transfer function. The so-called digital redesign technique based on closed-loop performance [23] or on minimizing a local quadratic performance index [20] is applied to achieve the same goal. Recently, new discretization methods of continuous-time controllers taking account of the inter-sample behavior of the resulting sampled-data system have been presented in [24]. Unfortunately, these techniques are not directly applicable to LPV controllers because of their limitation to time-varying systems [25]. This motivates to discuss a dedicate approach for discretizing LPV controllers.

Let us consider the LPV controller designed in the previous part with the state-space description

$$K_{rc} : \begin{cases} \dot{x}_c(t) &= A_c(\delta_\omega(t)) x_c(t) + B_c(\delta_\omega(t)) y_c(t) \\ u_c(t) &= C_c(\delta_\omega(t)) x_c + D_c(\delta_\omega(t)) y_c(t) \end{cases} \quad (15)$$

where $\delta_\omega(t)$ stands for time-varying parameter vector in δ_c .

The objective is to design a discrete-time LPV controller K_{rc}^d such that the performance of the closed-loop system is as close as possible to that of the closed-loop system with the

continuous-time LPV controller. In this paper, the trapezoidal technique is used for the real-time implementation of the LPV controller since it guarantees the necessary approximation error with a reasonable sampling rate. The following discussion about this method is mainly based on [25].

For a sample time $T > 0$ let us introduce $x_{c,k} \triangleq x(kT)$, $y_{c,k} \triangleq y(kT)$ and $\delta_k \triangleq \delta_\omega(kT)$. If the scheduling parameter that is measured in the time interval $[kT, (k+1)T]$ is considered to be equal to its value at time kT , we infer

$$x_{c,k+1} = x_{c,k} + \int_{kT}^{(k+1)T} A_c(\delta_k)x_c(\tau) + B_c(\delta_k)y_c(\tau)d\tau. \quad (16)$$

Approximating the integral (16) using the trapezoidal rule [25] results in $x_{c,k+1} \approx x_{c,k} + (T/2)(A_c(\delta_k)x_{c,k} + B_c(\delta_k)y_{c,k} + A_c(\delta_{k+1})x_{c,k+1} + B_c(\delta_{k+1})y_{c,k+1})$. If assuming that

$$H(\delta) := \left(I - \frac{T}{2}A_c(\delta)\right) \text{ is invertible for all } \delta \in \delta_c \quad (17)$$

we get $x_{c,k+1} = H(\delta_{k+1})^{-1}(I + (T/2)A_c(\delta_k)x_{c,k} + (T/2)H(\delta_{k+1})^{-1}(B_c(\delta_k)y_{c,k} + B_c(\delta_{k+1})y_{c,k+1}))$. Then

$$z_{c,k+1} \triangleq \frac{1}{\sqrt{T}}H(\delta_{k+1})x_{c,k+1} - \frac{\sqrt{T}}{2}B_c(\delta_{k+1})y_{c,k+1} \quad (18)$$

leads, after a short computation, to

$$z_{c,k+1} = \left(I + \frac{T}{2}A_c(\delta_k)\right)H(\delta_k)^{-1}z_{c,k} + \sqrt{T}H(\delta_k)^{-1}B_c(\delta_k)y_{c,k}. \quad (19)$$

Similarly, the output equation can be expressed as $u_{c,k} = C(\delta_k)x_{c,k} + D(\delta_k)y_{c,k}$ which results in a similar fashion in

$$u_{c,k} = \sqrt{T}C_c(\delta_k)H(\delta_k)^{-1}z_{c,k} + \left[\frac{T}{2}C_cH(\delta_k)^{-1}B_c(\delta_k) + D_c(\delta_k)\right]y_{c,k}. \quad (20)$$

Equations (19) and (20) are used to implement the discrete-time LPV controller as described in Section II. If we denote the spectral radius of a matrix by $\rho(\cdot)$, we note that the invertibility condition (17) is guaranteed if the sampling time T satisfies [25] $T < 2\rho(A_c(\delta))^{-1}$ for all $\delta \in \delta_c$.

A part of the problem concerning the discretization of the continuous-time LPV controllers and the implementation in a real setup has been also presented in [26].

V. EXPERIMENTAL RESULTS

The subsequently described experiments were performed on the test platform at the Institute of Electrical Power Engineering, Dresden University of Technology, Germany. This platform is based on the multi processors ALPHA-COMBO system produced by dSPACE. A rapid control prototyping structure used for supervision and control has been developed by the author of [4]. The laboratory setup involves an 11 kW induction

Fig. 4. Doubly fed induction machine and induction motor.

motor as the prime mover and a 4 kW DFIM was used for all experiments. The moment of inertia of the DFIM is 0.032 kgm^2 , the stator resistance is 1.070Ω , the rotor resistance is 1.32Ω , the mutual inductance is 0.1601 H , and the stator and rotor leakage inductances are 0.0066 H and 0.0098 H , respectively. The experimental DFIM is coupled to the induction motor as shows in Fig. 4. The experimental results have been performed for the rotor speed range from 1050 rpm to 1950 rpm. Experiments involving stator voltage dips were implemented by grid symmetrical faults created by short circuits of resistors.

The following weighting functions have been chosen for designing the continuous-time LPV current controller:

$$W_{rs}(s) = \begin{pmatrix} \frac{650}{s+0.05} & 0 \\ 0 & \frac{950}{s+1.05} \end{pmatrix}, \quad W_{rt}(s) = \begin{pmatrix} \frac{0.01}{s+0.95} & 0 \\ 0 & \frac{0.01s+0.03}{s+0.95} \end{pmatrix}.$$

This results in an optimal $\mathcal{L}_2$ -gain of $\gamma_{LPV} = 0.4483$. The invertibility condition (17) and a simulation-based comparison of the continuous-time with the discretized LPV controller lead to the choice $T = 0.25 \text{ ms}$ for the sampling time.

A. Step Responses of Electrical Torque and Reactive Power

The results with the rotor speed fixed at 1350 rpm are presented in Figs. 5 and 6. In Fig. 5, the test is carried out with step changes of the reference values of the reactive power Q_g from 0 VAR to 700 VAR. The performance of the controlled system with step changes of the electrical torque from -3 Nm to -10 Nm is shown in Fig. 6. The experimental results show that the tracking and decoupling of the electrical torque and reactive power are very good in the presence of noisy measurements. The ripples in the measured rotor current, the electrical torque, and the reactive power are also small because of the good performance achieved for the inner current-control loop. Note that the rotor speed oscillations are due to changes of the electrical torque developed by the machine since the rotor speed depends mainly on the difference between the electrical and mechanical torque.

Fig. 5. Step response of the reactive power near synchronous speed (1350 rpm).

Fig. 6. Step response of the electrical torque near synchronous speed.

B. LPV Controller Performance Under Rotor Speed Variations

Since the above tests have been performed with fixed values of the rotor speed they do not provide an indication about the performance of the system during rotor speed changes. The following experiments are aimed at an evaluation of this issue.

Figs. 7–9 show how the reactive power Q_g , the electrical torque T_e , and the dq components of the rotor currents change during variations of the rotor speeds. These results reveal the important point that the performance of the controlled system is indeed maintained during rotor speed changes in the whole operating region from 1050 rpm to 1950 rpm as expected from the LPV design.

C. Grid Voltage Dip Tests With the LPV Controller

Since the total power of installed wind turbine systems worldwide has incessantly increased, the separation of wind turbines from the grid is only allowed under some specific

Fig. 7. Tracking behavior of the controlled system with LPV controller for variations of the rotor speed from 1050 rpm to 1350 rpm.

Fig. 8. Tracking behavior of the controlled system with LPV controller for variations of the rotor speed from 1350 rpm to 1800 rpm.

Fig. 9. Tracking behavior of the controlled system with LPV controller for variations of the rotor speed from 1800 rpm to 1950 rpm.

situations [27], [28]. In fact, the unsupervised disconnection of wind turbines from the grid might lead to instabilities in the electric power distribution. Therefore, the robustness of the

Fig. 10. Grid voltage dips test at near-synchronous speeds of the rotor.

controlled system under the effects of grid voltage sags is an important aspect for controller design [29].

Fig. 10 shows the system performance when the grid undergoes faults at a rotor speed of 1350 rpm. The grid voltage drops to 50% of its rated value for a period of 200 ms before it recovers to its normal value. The overshoots of the controlled variables get larger along with the rotor speed. The oscillations of the rotor currents are well damped after only about 40 ms corresponding to about two sampling periods after the fault is cleared. This indicates that the controlled system with the LPV controller is robust against stator voltage dips and is capable of riding through grid faults rapidly.

D. Comparison With FRT Controllers

In this last part we compare the experimentally achieved performance with that for a conventional FRT design, again concerning electrical torque and reactive power tracking in view of rotor speed changes, and the robustness of the controlled system in the presence of stator voltage dips.

Fig. 11. Performance of the controlled system with FRT controller for variations of the rotor speed from 1050 rpm to 1350 rpm.

Fig. 12. Performance of the controlled system with FRT controller for variations of the rotor speed from 1350 rpm to 1500 rpm.

Figs. 11 and 12 show the reactive power Q_g , electrical torque T_e , and the dq components of the rotor currents with respect to the variation of the rotor speed from 1050 rpm to 1350 rpm and from 1350 rpm to 1500 rpm, respectively. As can be seen, the performance of the closed-loop system with the FRT controller degrades when the rotor speed increases, in contrast to the results for the LPV controller in Figs. 7–9 and for the whole operating region from 1050 rpm to 1950 rpm. The poor performance for the FRT controller demonstrates the benefits of LPV synthesis over conventional FRT designs in a real setup.

E. Grid Voltage Dip Tests With the FRT Controller

As mentioned above, a high rotor speed induces more oscillations in the controlled variables for the FRT controller.

Fig. 13. Grid voltage dips test at near-synchronous speeds of the rotor with the FRT current controller.

Especially, a higher over current will occur during the fault if the rotor speed is at a high value before a stator voltage fault. The following grid voltage dip tests have only been carried out at near-synchronous rotor speed (1350 rpm) to avoid exceeding the current limit of the converter in transient periods.

Fig. 13 shows the system performance when the grid undergoes faults as described above. The experimental results show that the oscillations of the rotor currents are not well damped after the fault clearance. In contrast, for the LPV controller, the rotor currents quickly recover to their normal values after only about two sampling periods (see Fig. 10) when the fault is cleared. The slow settling to normal values for the FRT controller leads to significant fluctuations of the reactive power and the electrical torque after the fault. This phenomenon becomes more serious if the rotor speed is above the synchronous speed since the performance degrades as discussed above. In brief, the controlled system with the FRT controller is not robust against stator voltage dips, especially at high rotor speeds.

VI. CONCLUSION

A self-scheduled LPV current controller for a DFIM applied to a wind energy conversion system has been synthesized to maintain the performance requirements for all trajectories of the rotor speed over its whole range of operation. A suitable discrete-time controller has been designed to ensure the implementability of the resulting digital control system for a chosen sampling time. The laboratory-scale test results show that, with the chosen sampling frequency, the designed controller obtains the required performance for all trajectories of the rotor speed over its operating range and guarantees robustness of the controlled system under grid voltage faults.

The experimental results also confirm the superior performance of LPV controllers if compared to a conventional FRT design. The LPV controller leads to faster tracking of the reference values with less overshoot. It also induces less oscillations in the controlled variables in comparison with that of the FRT controller. Moreover, as the most important feature, the LPV controller guarantees stability and performance of the closed-loop system for time-varying rotor speeds over the whole operating range. In case of grid voltage dips and in contrast to an FRT controller, the experimental results for the LPV controller show that the oscillations of the rotor currents are well damped after the fault is cleared.

A thorough experimental robustness analysis of the designed LPV controller against variations in all machine parameters is left for future research.

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