Active damping: an $\mathcal{H}_\infty$ model reference approach

Jorge Pérez, *Member, IEEE*, Santiago Cobreces, *Member, IEEE*, Robert Griñó, *Senior Member, IEEE*, Francisco Huerta, *Member, IEEE*, Francisco Javier Rodríguez, *Member, IEEE*

Abstract—This paper focuses on the design of an active damping controller that shapes the dynamics of an LCL-grid connected voltage source converter (VSC) to mimic an L-filter connected VSC. This makes possible to use simpler classical proportional plus resonant (PR) or proportional plus integral (PI) controllers design methods to regulate the grid current with minimum penalties in both performance and stability margins. The active damping design follows a model reference approach and is synthesized through a (sub-)optimal $\mathcal{H}_\infty$ algorithm. The controller uses the grid current and PCC voltage although it allows the selection of a different set of measurements. Effectiveness of the method is demonstrated with time and frequency domain experimental results.

I. INTRODUCTION

Distributed power generation systems generally rely on the use of controlled pulse-width modulation (PWM) converters as their grid-connection interfaces for energy conversion and conditioning. Multiple objectives can be achieved by implementing a suitable current controller, such as adjustable power factor, controllable DC-link voltage or reduced line current harmonic distortion [1]. To ensure low THD grid currents, a filter that reduces PWM commutation ripple is connected between the voltage-source converter (VSC) and the grid. One of the most common is the LCL filter, due to its good cost-filtering capabilities ratio [2].

In opposition to the simpler L filter, the use of an LCL filter increases control complexity as it introduces a pair of complex-conjugated poles close to the $j\omega$ axis that may cause current oscillations in response to system disturbances, which may lead to system instability in the presence of ill-conditioned closed-loop systems [3] or weak grids [2].

The progressive advance of LCL use has motivated two alternative control approaches. The first group is formed by standard single loop approaches, usually more sophisticated than those used for L design, that directly consider LCL dynamic complexity in the design process [4]–[7]. A second group proposes the use of several strategies to damp LCL

Fig. 1. Block-diagram description of the proposed active-damping approach.

resonances which, once applied, present to the reference-tracking controller, a simplified dynamic model keeping its design simple. This category is divided into passive -hardware- or active -software- approaches. The proposal on this article belongs to the latter category.

Passive damping [8]–[12] adds passive elements, commonly resistances, to the filter. This simple solution comes at the cost of extra power losses and, in some cases, reduction of the high-frequency attenuation capability [10].

Active damping techniques [1], [2], [13]–[19] try to achieve similar objectives dividing the design into two stages: a first stage that damps resonances, usually using an extra inner control loop, and a second stage consisting in the design of a reference-tracking controller that faces a simplified plant, ideally one similar to an L filter dynamic model. This lossless simplification of the outer controller makes active damping very popular on both industry and academia.

Most active damping methods are based on the emulation of a virtual resistance in some position of the grid filter. Approaches usually operate using the measurement of at least one extra LCL-filter state, commonly the capacitor voltage (or current), in addition to the controlled (grid-side or converter-side) current [15], [17], [19]. Other proposals include more filter measurements to achieve better results [20].

A second group of active damping methods try to achieve the same objective using the already available current measurement, without increasing the number of sensors. They are sometimes referred as filter-based active damping. Two main options exist in this category: the use of a lead-lag network (high-pass filter) [15], [21], to avoid negative gain margins placing extra gain on problematic frequencies, and the use of a notch filter, to attenuate the resonance by means of adjusting the anti-resonance peak of the notch filter at the resonance frequency of the LCL filter [22], [23].

This paper proposes an active damping approach that consists in the introduction of an inner control loop that makes the resulting plant emulate the dynamic behaviour of an L-filter connected converter. This inner controller is designed using

Jorge Pérez is with Gamesa Electric S.A.U. Madrid, Spain. Email: jorge.perez@depeca.uah.es

Santiago Cobreces and Francisco Javier Rodríguez Sánchez are with the Department of Electronics, Universidad de Alcalá, Spain. Email: cobreces@depeca.uah.es. Their work was supported by *Ministerio de Economía, Industria y Competitividad* and *Agencia Estatal de Investigación* with projects DPI2017-88505-C2-2-R, ENE2014-57760-C2-2-R and by *Universidad de Alcalá* with project CCG2015/EXP-064.

Robert Griñó is with the Institute of Industrial and Control Engineering (IOC), Universitat Politècnica de Catalunya (UPC), Spain. Email: roberto.grino@upc.edu. His work was supported in part by the Government of Spain through the *Ministerio de Economía, Industria y Competitividad* and *Agencia Estatal de Investigación* Projects DPI2013-41224-P and DPI2017-85404-P.

Francisco Huerta is with the Carlos III University of Madrid, 28911 Leganés, Spain. Email: fhuerta@ing.uc3m.es

a model-reference optimal $\mathcal{H}_\infty$ approach that translates this design complexity into a convex optimisation problem, with guaranteed performance. The high accuracy of the emulation allows directly using the controller that would be used in the L-filter case with almost identical results.

The proposal avoids the virtual-resistor paradigm, that was actually problematic in the most typical digital control scenario [21], [24], [25] and directly faces one of the active damping deeper objectives: to make the LCL filter as easy to control as the L filter, obtaining its improved filtering and cost advantages without worrying about its dynamic particularities-resonances-.

The casting of the problem inside the $\mathcal{H}_\infty$ problem converts the design process in an almost systematic procedure that formally guarantees an optimal controller calculated only from converter parameters, making it suitable for its implementation in a software tool that may be of interest for its industrial use. This easy design flow contrasts with that of traditional active damping approaches, which usually involve non trivial design iterations, with non-convex optimisation algorithms, parameter sweeping and outer controller re-tuning that may not success. $\mathcal{H}_\infty$ control is not new in grid converters control [26]–[32] but has not been explored for the active damping topic.

One inherent advantage of this proposal with respect to previous approaches is its flexibility regarding the measurements needed, offering very accurate results measuring only one or several of the filter variables: converter current, grid current, and/or capacitor voltage. When compared with previous approaches that only use one current measurement, perhaps the most interesting configuration, it offers better performance in the achieved damping and also in the resulting plant phase, that keeps positive on wide frequency regions. This fact is known to have important stability and robustness implications [33].

Next section develops open and closed-loop dynamics of the proposed application. Section III is centred on the proposed controller, its design, synthesis, implementation and limitations. Experimental results of the proposed method, both in time and frequency domains, are shown in section IV, as well as the analysis of the robustness of stability and damping and its performance comparison with related classical damping approaches. Finally, the paper conclusions can be found in section V.

II. PROBLEM DESCRIPTION AND MODELLING

Fig. 1 shows a simplified scheme of the considered application: a current-control loop of a VSC connected to the grid through an LCL filter.

Instead of facing this objective directly, in a single loop, this paper proposes to use nested control loops. In the inner loop of the control scheme shown in Fig. 1, the damping controller $\mathbf{K}_{AD}$ will reshape the LCL-filter open-loop dynamics so it resembles the one of a simpler L filter. Then, a simple outer grid-current controller (K_{cc} in Fig. 1), such as classic PR and PI regulators, can be designed for this emulated L-filter dynamic.

The design methodology is intended for three-phase system. The problem is approached in uncoupled $\alpha\beta$ axes [34],

reducing the original three-phase problem to two identical single-phase problems. For notation simplicity, only one of the controlled channels is considered in this paper for both plant modelling and controller design.

A. Open-loop modelling

System information can be obtained from three measurable variables (states): the grid and converter side currents i_2 and i_1 , respectively, and the capacitor voltage v_c . Although the presented methodology is able to operate only with i_2 , that is perhaps the most interesting case, the method is also able to work with a different set of measured variables.

Those variables are affected by two inputs: the VSC average output voltage over a PWM period, u , (the control variable) and the measured point of common coupling (PCC) voltage, v_s (the disturbance variable), as it is showed below:

$$\underbrace{\begin{bmatrix} I_2 \\ I_1 \\ V_c \end{bmatrix}}_{\mathbf{X}(s)} = \underbrace{\begin{bmatrix} F_{u \rightarrow i_2} \\ F_{u \rightarrow i_1} \\ F_{u \rightarrow v_c} \end{bmatrix}}_{\mathbf{G}(s)} U(s) + \underbrace{\begin{bmatrix} F_{v_s \rightarrow i_2} \\ F_{v_s \rightarrow i_1} \\ F_{v_s \rightarrow v_c} \end{bmatrix}}_{\mathbf{G}_d(s)} V_s(s), \quad (1)$$

where $\mathbf{G}(s)$ and $\mathbf{G}_d(s)$ are the MIMO dynamics from $U(s)$ and $V_s(s)$, respectively, to the measurements vector $\mathbf{X}(s)$, represented in the Laplace domain. These matrices are broken down below:

$$F_{u \rightarrow i_2}(s) = -\frac{Y_{L_2} \cdot Z_C \cdot Y_{L_1}}{1 + Z_C \cdot (Y_{L_1} + Y_{L_2})}, \quad (2)$$

$$F_{v_s \rightarrow i_2}(s) = \frac{Y_{L_2} \cdot (1 + Y_{L_1} \cdot Z_C)}{1 + Z_C \cdot (Y_{L_1} + Y_{L_2})}, \quad (3)$$

$$F_{u \rightarrow i_1}(s) = -\frac{Y_{L_1} \cdot (1 + Y_{L_2} \cdot Z_C)}{1 + Z_C \cdot (Y_{L_1} + Y_{L_2})}, \quad (4)$$

$$F_{v_s \rightarrow i_1}(s) = \frac{Y_{L_1} \cdot Z_C \cdot Y_{L_2}}{1 + Z_C \cdot (Y_{L_1} + Y_{L_2})}, \quad (5)$$

$$F_{u \rightarrow v_c}(s) = \frac{Y_{L_1} \cdot Z_C}{1 + Z_C \cdot (Y_{L_1} + Y_{L_2})}, \quad (6)$$

$$F_{v_s \rightarrow v_c}(s) = \frac{Y_{L_2} \cdot Z_C}{1 + Z_C \cdot (Y_{L_1} + Y_{L_2})}, \quad (7)$$

with

$$Y_{L_2}(s) = \frac{1}{L_2 s + R_2}, \quad Y_{L_1}(s) = \frac{1}{L_1 s + R_1}, \quad Z_C(s) = \frac{1}{C \cdot s},$$

where L_1, L_2 are the LCL converter and grid-side inductances, R_1, R_2 are their equivalent losses resistances and C is the LCL-filter capacitor.

B. Inner closed-loop damped grid-current dynamic

Active damping controller $\mathbf{K}_{AD}$ calculates the VSC average AC voltage, u , from the selected filter measurements, the PCC voltage v_s and the output of the outer grid-current controller u_{AD} :

$$U = K_{v_s} V_s + K_u U_{AD} + K_{i_2} I_2 + K_{v_c} V_c + K_{i_1} I_1. \quad (8)$$

The philosophy underlying this procedure is the creation of a control structure that involves the original plant, hiding its dynamics, and presents to the outer controller a simplified model. The new plant actuation variable is U_{AD} .

Substituting (8) in (1), grid-current (i_2) dynamic can be derived:

$$I_2(s) = G_{AD}(s) \cdot U_{AD}(s) + G_{dAD}(s) \cdot V_s(s), \quad (9)$$

where $G_{AD}(s)$ and $G_{dAD}(s)$ expressions are shown in (10).

From the outer current controller K_{cc} perspective, G_{AD} and G_{dAD} represent the dynamic model of a new grid-current plant with two inputs: u_{AD} , which is controlled by K_{cc} , and the PCC voltage v_s , which is an external uncontrolled disturbance.

III. ACTIVE DAMPING DESIGN

This section is centred in the design of the controller K_{AD} , its synthesis fundamentals, procedure and limitations. The philosophy lying under this procedure is the seek for a controller that minimizes the difference between the response of the controlled plant, an LCL-filter grid-connected VSC, and that of the reference model, an L-filter grid-connected VSC. This minimization is faced from a $\|\cdot\|_\infty$ optimization point of view.

A. Methodology

The controller K_{AD} is computed by means of an $\mathcal{H}_\infty$ synthesis. The input information that is needed to start the synthesis process is integrated in the generalised plant $\mathbf{P}$ [35]. $\mathbf{P}$ is a virtual plant that contains information of the open-loop system dynamics, the controller structure, and its performance objectives and constraints. Structurally, $\mathbf{P}$ is defined as:

$$\begin{bmatrix} z \\ v \end{bmatrix} = \mathbf{P} \begin{bmatrix} w \\ u \end{bmatrix}, \quad (11)$$

where w is the exogenous inputs vector to the system (usually formed by closed-loop disturbances and references), z is the vector to be minimised by the synthesised controller (or errors vector), u is the actuation vector (outputs of the obtained controller to be applied to the plant input) and v is the measurements plant output (input to the obtained controller).

With the information contained in $\mathbf{P}$, the synthesis algorithm will compute a (sub-)optimal controller K_{AD} which minimises the infinity norm¹ of the virtual closed-loop system $\mathbf{N}$ that results from the feedback interconnection of $\mathbf{P}$ and K_{AD} , and relates exogenous input vector w and error vector $z = \mathbf{N}w$:

$$\min_{K_d} \left(\|\mathbf{N}(\mathbf{P}, K)\|_\infty = \max_{w(\omega) \neq 0} \frac{\|z(\omega)\|_2}{\|w(\omega)\|_2} \right) \leq \gamma. \quad (12)$$

Thus, K_{AD} minimizes the energy of the error signals z for the *worst-case direction* and *worst-case frequency* of the

¹The infinity norm of a MIMO system $\mathbf{H}(s)$ in the frequency domain is defined as $\|\mathbf{H}(s)\|_\infty \triangleq \sup_\omega \bar{\sigma}(\mathbf{H}(j\omega))$, where $\bar{\sigma}(\mathbf{H}(j\omega))$ is the maximum singular value of $\mathbf{H}(j\omega)$. For SISO systems that is simplified to $\|\mathbf{H}(s)\|_\infty \triangleq \sup_\omega |H(j\omega)|$.

input vector w . The design keystone of this control paradigm is, then, to translate controller objectives to a proper definition of the generalized plant $\mathbf{P}$.

Fig. 2 shows the proposed structure for $\mathbf{P}$ (displayed in red color). The original LCL plant, $G(s)$ and $G_d(s)$ –described in (1)– is displayed in orange. The extra elements added to incorporate design objectives into the problem are drawn in purple. Those elements generate an error signal, e_d , that compares the output current, i_d , of a reference model, G_{ref} , with that of the actual plant, i_2 .

A controller yielding a small e_d will make the closed-loop system mimic the dynamic behaviour of the reference model, so e_d is included in the vector to be minimized, z . Control signal, u , needs to be also included in z to avoid falling in a solution (i.e. controller) that needs an unrealistic control level and, also, to limit the effective control bandwidth. Transfer functions W_d and W_u are weighting functions that determine the range of frequencies where the reference model will be imitated (i.e. e_d is minimized) and the actuation signal u will be limited, respectively.

The exogenous inputs vector w is formed by the two disturbances to the damped plant, u_{AD} and v_s (see (9)). In addition to feedforward these signals, K_{AD} will feedback some filter measurements, forming the vector v . In Fig. 2, as well as in (8), (9) and (10), a complete states vector x feedback was considered to not lose generality. The method is, in any case, quite flexible and allows using one or several measurements of the grid filter (e.g. the grid current i_2) being only necessary in that case to modify the definition of the vector v .

Plant $\mathbf{P}$ inputs and outputs are summarized below:

$$z = \begin{bmatrix} W_d \cdot e_d \\ W_u \cdot u \end{bmatrix}, \quad v = \begin{bmatrix} v_s \\ u_{AD} \\ x \end{bmatrix}, \quad w = \begin{bmatrix} v_s \\ u_{AD} \end{bmatrix}, \quad u = u. \quad (13)$$

B. Reference model and weighting function selection

The design of the generalised plant $\mathbf{P}$ involves the selection of three transfer functions: the model reference transfer function G_{ref} and the two involved weighting functions W_d and W_u .

G_{ref} is the way the designer specifies the damping strategy of the plant. This article proposes to mimic an L-connection scenario, where the inductance and equivalent resistance match the addition of the actual LCL-filter inductances and resistances:

$$G_{ref}(s) = \frac{1}{R_{ref} + sL_{ref}} = \frac{1}{(R_1 + R_2) + s(L_1 + L_2)}. \quad (14)$$

This transfer function closely resembles that of a damped version of the LCL one. Given the dipole nature of the L filter, a single reference model is enough to shape both G_{AD} and G_{dAD} (defined in (10)), subtracting to that end u_{AD} from v_s , as it can be observed in Fig. 2 (purple colour).

Besides, the dynamic particularities derived from the sampled-data nature of the controller are also the cause of one fundamental limitation of the closed-loop system. The sampling frequency of the controller, together with the delay at the plant input, induces a limitation on the maximum achievable controller bandwidth, f_c , of

$$\omega_{c_{MAX}} = 2\pi f_{c_{MAX}} \approx \frac{1}{T_s} \text{ rad/s}, \quad (17)$$

being T_s the sampling period of the digital system [35]. It is important to remark that this bandwidth limitation affects to both feedback and feedforward paths. Good damping results can not be assured, then, if the LCL resonance frequency f_{res} is above $f_{c_{MAX}}$. Consequently, the LCL resonance frequency imposes a minimum sampling and control frequency to achieve active damping. This is a classical theoretical limitation and affects to any sampled-data linear strategy to damp LCL filter resonance.

2) *Weights design limitations*: Elements in $\mathbf{P}$ must be strictly stable and proper. From a design point of view, it means that the defined weights can not have pure integral/derivative terms or more zeros than poles. This limitations have, in any case, a very small influence in the controller performance, as it is possible to emulate pure integral/derivative terms by placing poles or zeros arbitrarily close to the $j\omega$ axis, and to make $\mathbf{P}$ proper by adding as many high frequency poles as necessary.

In addition, it must be taken into account that the synthesised active damping controller $\mathbf{K}_{AD}$ will have the same order of the generalised plant $\mathbf{P}$. This means that, even though higher-order frequency weights in $\mathbf{P}$ can be translated in a better damping performance, their definition will result in higher controller complexity and, then, higher computational times.

A more detailed explanation of a controller $\mathcal{H}_\infty$ synthesis and implementation, including the delays modelling and the final discrete controller obtaining, can be found in [36].

IV. RESULTS

This section analyses experimentally and numerically the validity of the proposed control procedure. Both cases assume a converter with the following parameters.

A. Validation conditions

The main converter parameters used in numerical and experimental analysis are shown on Table I.

TABLE I
SETUP PARAMETERS

S_n	9 kVA	L_1	3.4 mH
V_g	120 V	R_1	28.8 mΩ
ω_1	$2\pi 60 \text{ rad}\cdot\text{s}^{-1}$	L_2	1.7 mH
V_{DC}^*	700 V	R_2	18.6 mΩ
T_s	200 100 μs	C	18 μF
T_{sw}	$2T_s$	C_{DC}	4.7 mF

The study considers two different control sampling periods to better evaluate the influence of this parameter on controller design and performance.

1) *Controller location on complete control loop*: The controllers obtained using the proposed design procedure are not able to fully control the converter on their own, and need additional external control loops. These controllers have been chosen so they are as standard as possible.

The reference tracking is accomplished by a classical outer PR grid current controller, designed for the L equivalent filter reference in (14) [17]:

$$K_{cc}(z) = K_p \left(1 + \frac{\sin(\omega_1 T_s)(z^2 - 1)}{T_r 2\omega_1 (z^2 - 2z \cos(\omega_1 T_s) + 1)} \right), \quad (18)$$

where the resonant time constant and the proportional gain are fixed for both sampling periods to $T_r = 0.004$ and $K_p = 12.648$, respectively. The controller bandwidth is the same, then, for both considered sampling periods. The resonant time constant T_r is chosen for a fast reference tracking. The proportional gain K_p is tuned so applying K_{cc} to the LCL filter (without damping) for a sampling period $T_s = 200 \mu\text{s}$ results in a unstable system, but with stability margins close to zero. This is done to test the active damping efficiency as a system stabilizer (see the test in Fig. 14).

Considering the damped LCL plant, the closed-loop dynamic of the controlled current i_2 follows the expression:

$$I_2 = \underbrace{\frac{G_{AD} \cdot K_{cc}}{1 + G_{AD} \cdot K_{cc}}}_{T(s)} I^* + \underbrace{\frac{G_{dAD}}{1 + G_{AD} \cdot K_{cc}}}_{Y(s)} V_s, \quad (19)$$

where $T(s)$ and $Y(s)$ are the closed-loop tracking and admittance transfer functions. The stability of the damped system can be obtained by means of linear time invariant (LTI) analysis of the loop function $L_{AD} = K_{cc} \cdot G_{AD}$. A perfect damping and improved stability margins would be obtained if L_{AD} is equal to its reference value $L_{ref} = -K_{cc} \cdot G_{ref}$. If the inner active damping loop is not considered, L_{AD} would be equal to $L = K_{cc} \cdot F_{u \rightarrow i_2}$.

The considered application used for controller verification is an active PWM rectifier that needs a DC-voltage control loop. This outer loop consists of a standard PI-based control loop ($K_{DC}(z) = K_p + \frac{K_I T_s}{z-1}$, $K_p = 0.0992$, $K_I = 2.0742$), whose input is the difference between the DC-bus voltage reference and the measured value. Grid current references are computed from the K_{DC} controller output, together with the reactive power reference and the information extracted from grid PCC voltages.

2) *Numerical analysis and simulation platform*: All numerical results shown on this section have been obtained using Mathworks Matlab software tool. Simulation experiments have been performed using Matlab Simulink tool, with the aid of SimPowerSystems (sub)library.

3) *Experimental platform*: All experiments have been conducted on GEISER group facilities in Universidad de Alcala.

Fig. 4 shows a picture of the experimental set-up. It consist of an AC programmable power supply Pacific SmartSource 345-AMX, emulating the grid, and a two-level VSC of nominal power S_n connected to a LCL filter and a DC-bus capacitance C_{DC} . Passive loads R_L are connected to the DC-bus to test the platform under different operating points.

Fig. 4. Picture of the implemented experimental platform.

Fig. 5. LCL shaping results for the two different considered sampling periods, feeding-back only i_2 . Dynamic references (i.e. the one of an L filter) are represented in dark blue, the open loop LCL dynamic is represented in red and the obtained modified LCL dynamic is shown in light blue. Damping results improves for a smaller sampling period, thanks to the higher K_{AD} maximum bandwidth f_{cMAX} .

Control algorithm is implemented on a Texas Instruments DSP TMS320DSK6713.

B. Damping effectiveness analysis

Fig. 5 shows the frequency-domain shaping (numerical) results of the synthesized controller for the two considered sampling periods, feeding-back only the grid current i_2 (i.e. $\mathbf{x} = i_2$ in Fig. 2). As it can be seen, both modified LCL-filter

Fig. 6. Modified (i.e. damped) LCL dynamic considering different feedback loops: (green) only the grid current i_2 (dark blue) only the capacitor voltage v_c (red) only the converter current i_1 (light blue) the complete LCL filter states vector $\mathbf{x}$.

dynamics, G_{AD} and G_{dAD} , follow the L-filter reference, G_{ref} , damping the resonance of the unmodified LCL-filter dynamics, $F_{u \rightarrow i_2}$ and $F_{v_s \rightarrow i_2}$. Note that the proposed controller loses effectiveness around the maximum bandwidth, f_{cMAX} , (i.e. all K_{AD} terms magnitude in (10) tends to zero, and so does its output u), so the modified LCL plant dynamics G_{AD} and G_{dAD} tend to zero and its open loop value $F_{v_s \rightarrow i_2}$, respectively. The results are better, then, for the smaller sampling period case $T_s = 100 \mu s$, as the resonance frequency f_{res} is inside the achievable controller bandwidth. Damping results can be further improved for LCL filters with lower resonance frequencies and/or for lower implementation sampling periods.

Although the result shown in Fig. 5 corresponds to a design where only the grid-current (i_2) is measured for the synthesized active damper, the design is flexible enough to achieve similar results feeding back any subset of the filter variables, as shown in Fig. 6. In it, the frequency-domain (numerical) results of the damped dynamics (G_{AD} and G_{dAD}) are shown for a sampling period $T_s = 200 \mu s$ and different measured variables (identified in parentheses). Slightly worse results are obtained when only the converter current (i_1) is used. In any case, the results are very similar inside the active control band, before the band constraints described in § III-C1.

For that reason, all the experiments and conclusions reported on the rest of this section have been obtained using a damping controller that only uses the grid current i_2 and voltage v_s as measured variables, because it is considered the most interesting case in most applications.

Fig. 7 shows different numerical open-loop functions, and their deduced stability margins, when the designed grid current controller K_{cc} is: applied to the damped LCL plant G_{AD}

Fig. 7. Open-loop function frequency response numerical representation for the unmodified LCL filter (red), the reference L filter (dark blue) and the modified LCL filter (light blue). Their corresponding stability margins are detailed in TABLE II.

(see L_{AD}); applied to the L filter reference G_{ref} (see L_{ref}); and applied to the unmodified LCL plant $F_{u \rightarrow i_2}$ (see L). As stated before, K_{cc} is designed to be unstable if applied to the unmodified LCL plant with a sampling period $T_s = 200 \mu s$, with gain GM and phase PM margins close to zero (see GM_1 and PM_1 in table II). K_{cc} is again designed for a sampling period $T_s = 100 \mu s$, following (18), with the same design parameters K_p and T_r . This will result in a clearly unstable system if it is directly applied to the undamped plant (see GM_2 and PM_2 in table II).

TABLE II
STABILITY MARGINS FOR $T_s = 200 \mu s$ (TOP) AND $T_s = 100 \mu s$ (BOTTOM).

	No AD	AD	ref	$\frac{AD}{ref}$
PM_1	-0.31°	36.8°	41.3°	0.891
GM_1	0.05 dB	4.07 dB	5.86 dB	0.694
PM_2	44.3°	58.4°	63°	0.927
GM_2	-17.2 dB	8.9 dB	12 dB	0.741

Designed damping controller stabilizes both cases. Table II summarizes the achieved stability margins seen in Fig. 7. In it, sub-index AD and ref stands for the damped LCL filter case (loop L_{AD}) and the L filter reference case (loop L_{ref}), respectively, meanwhile the sampling periods $T_s = 200 \mu s$ and $T_s = 100 \mu s$ are identified by sub-index 1 and 2, respectively. Both designed controllers increase stability margins almost to their respective reference values, with better results for the smaller sampling period design, as can be seen in the rightmost

Fig. 8. Comparison between the sensitivity transfer functions of the reference L-filter system (S_{ref}) and the damped LCL-filter system (S_{AD}) for the two sampling periods considered (Upper: $T_s = 200 \mu s$, lower: $T_s = 100 \mu s$).

Fig. 9. Closed-loop numerical and experimental admittance frequency response for $T_s = 200 \mu s$ (a) without the proposed active damping (b) with the proposed active damping.

column that shows a performance index that compares the achieved stability margins to its reference values.

The improvement of the system stability robustness when the sampling period (T_s) is reduced is further demonstrated in the sensitivity functions comparison between the reference L-filter dynamic ($S_{ref} = (1 + L_{ref})^{-1}$) and the damped LCL-filter dynamic ($S_{AD} = (1 + L_{AD})^{-1}$) shown in Fig. 8. As it can be seen, the maximum sensitivity gain of the actively damped design ($\|S_{AD2}\|_\infty$) decreases when the sampling period is reduced to $T_s = 100 \mu s$, being its value much closer to its reference at that sampling period ($\|S_{ref2}\|_\infty$).

Fig. 9 shows the closed-loop input admittance frequency response results of the considered system (a) without the

Fig. 10. Closed-loop numerical and experimental admittance frequency response for $T_s = 100 \mu s$ and K_{AD} connected.

proposed active damping and (b) with the proposed active damping for a sampling period $T_s = 200 \mu s$. The purple line represents the numerical value of the obtained admittance (Y), whose dynamic was defined in (19), meanwhile red circles represents experimental identified values of it (Y_{exp}) taken at different frequencies². The resonance in the LCL open loop admittance ($F_{v_s \rightarrow i_2}$), in black, is highly damped with K_{AD} connected, but it is not with it disconnected³, as can be deduced from the experimental results.

Fig. 10 repeats the results shown in Fig. 9, this time for a reduced sampling period $T_s = 100 \mu s$. In this case, the undamped system is unstable in the experimental platform (as predicted theoretically, see GM_2 and PM_2 in Table II), not allowing to measure experimentally the obtained closed-loop admittance. Fig. 10 shows, then, only the results when the proposed active damping (K_{AD}) is connected. As it can be seen, the LCL filter resonance (see $F_{v_s \rightarrow i_2}$) is slightly better damped experimentally for this reduced sampling period.

Closed-loop input admittance is an important indicator of active damping performance as it specifies how the converter is going to react in the presence of grid disturbances, predicting possible oscillations or instability situations.

C. Robustness analysis

The grid impedance (Z_g in Fig. 1) seen by the converter from the PCC is usually unknown to a great extent. Particularly, the uncertainty associated with its inductance, L_g , is perhaps the main practical concern regarding converter closed-loop robustness and a potential source of cost increase due to maintenance and converter unavailability. Fig. 11 summarizes the robustness analysis, showing (a) the (numerical) closed-loop admittance frequency response magnitude, (b) the current time response and (c) the trajectories of the system closed-loop

²To do so, the system is excited with PCC disturbances at different frequencies, thanks to the grid simulator. The grid current response at each frequency is then measured. The relations between excitations and responses, in both phase and amplitude, give the experimental admittance values Y_{exp} .

³The undamped system (i.e. with K_{AD} disconnected) considering a $T_s = 200 \mu s$ is stable in the tested experimental platform, even though it is not theoretically (see GM_1 and PM_1 in table II), allowing to measure the experimental admittance (Y_{exp}) shown in Fig. 9(a). That is probably due to the more damped experimental behaviour of the system (compare Y_{exp} and Y phase around the resonant frequency).

Fig. 11. Analysis of the effect of L_g perturbation. (a) Expected admittance frequency response magnitude, (b) current time-response and (c) closed-loop poles positions for changes of an inductive grid $Z_g = L_g s$.

poles for variations of an inductive grid impedance $Z_g = L_g s$. As it can be seen, the system is well damped and stable for all possible variations of the grid inductance between $L_g = [0, 0.14]$ H.

Fig. 12 shows the damping robustness of the method towards parametric variations of the LCL filter. It represents, in the top, the closed-loop frequency response of the expected (numerical) admittance magnitude ($|Y|$) for changes of the LCL parameters L_1 , L_2 and C . High (resonant) values of $|Y|$ in Fig. 12 would indicate ineffective action of the designed active damping. A sampling period of $T_s = 100 \mu s$ was considered for the grid current controller (K_{cc}) and the damping controller (K_{AD}) designs. Then, the resonance frequency

Fig. 12. (Top) Expected admittance magnitude modification for changes of the LCL resonance frequency. High (resonant) values would indicate ineffective action of the designed active damping. Red boxes represent unstable cases. (Bottom) Grid current response to a reference change for different values of the filter parameters.

(ω_{res}) is changed from 0.85 to 1.25 pu of its nominal value by modifying each of the LCL elements. $|Y|$ in Fig. 12 for $\omega_{res} = 1$ pu is, then, equivalent to $|Y|$ in Fig. 10. The red boxes represent resonance frequency limits from which the system becomes unstable. In addition, the bottom part of Fig. 12 shows the simulated time domain response of the controlled current for the nominal filter and for stable variations of each filter parameter. The system is stable for the next variations of the filter parameters: $L_1 = [8.23, 0.56]$ pu⁴, $L_2 = [2.03, 0.62]$ pu and $C = [1.32, 0.66]$ pu; well above the expected tolerance on commercial filter inductances and capacitors, which is usually within $\pm 10\%$. To further demonstrate the robustness of the system, the top part of Fig. 13 shows the obtained closed-loop admittance for the nominal filter (see Y), for the worst damping case (i.e. larger $\|Y\|_\infty$) considering filter parameters within a tolerance of $\pm 10\%$ and the considered L_g uncertainty range (see Y_{wc}), and for this worst case scenario but with a non-ideal grid impedance (see Y_{wcL_g}). In addition, the bottom part of the figure shows the (simulated) current response to a reference change for the aforementioned cases. As it can be seen, the damping results for the worst filter parameters⁵ are still enough to result in small oscillations in the controlled current. It is also important to remark that a non zero grid inductance modifies the filter resonance frequency but does not worsen the damping performance.

⁴The nominal value of the filter resonance is $\omega_{resN} = \sqrt{(L_{1N} + L_{2N}) / (L_{1N} L_{2N} C_N)} \approx 7000$ rad/s, where L_{1N} , L_{2N} and C_N are the filter nominal parameters shown in table I

⁵The worst damping case for variations of the filter parameters within the considered tolerance are $C = 1.1$ pu, $L_1 = 1.1$ pu and $L_2 = 1.1$ pu.

Fig. 13. Closed-loop admittance and current response to a reference change; (red) for the nominal filter conditions and an ideal grid impedance; (dashed green) for the worst filter case, in terms of resonance damping, with a parameters tolerance of $\pm 10\%$ and an ideal grid impedance; (dashed blue) for the worst filter case and a non-ideal grid (dashed blue).

D. Time domain results

Fig. 14 shows simulated time domain results of the performance of the designed active-damping (K_{AD}) for a $T_s = 200$ μ s. Two applications were simulated simultaneously to obtain these results: one VSC connected to the grid through the equivalent L filter, controlling its grid current (i_L) with only the proposed external current controller (K_{cc}), and another

Fig. 14. Simulated time domain active damping results. i_{abcL} and i_{abcLCL} represent the grid current for the emulated L-filter and the considered LCL-filter, respectively. Both are equivalent if K_{AD} is connected. An unstable oscillation grows if K_{AD} is disconnected. This oscillation is quickly dissipated if K_{AD} is again connected.

Fig. 15. Experimental time domain results of the controlled current with and without the proposed active damper for a $T_s = 100 \mu s$.

VSC connected to the same grid through the considered LCL filter, controlling its current (i_{LCL}) with the proposed active damper (K_{AD}) and the aforementioned outer current controller (K_{cc}). The response of the LCL-filter grid current (i_{LCL}) is very similar to the one of the emulated L-filter (i_L), demonstrating the good dynamics shaping achieved with K_{AD} . It can be seen how the grid current of the LCL-filter (i_{LCL}) starts to oscillate at the resonant frequency if K_{AD} is disconnected. If K_{AD} is connected again, this oscillation is quickly dissipated.

Fig. 15 shows the experimental evolution of a controlled system under K_{AD} different connection status. The top of the figure shows how the current becomes unstable if K_{AD} is disconnected during a long period. The bottom of the figure shows a short-period K_{AD} disconnection. As it can be seen, connecting again K_{AD} will stabilize the system.

Fig. 16 to Fig. 18 show experimental results of the implemented damped LCL grid current control at sampling period $T_s = 100 \mu s$ for different tests: Fig. 16 shows the response to a change of the passive DC-load power demand from $P = 0$ to $P = 4.2$ kW, Fig. 17 shows the results of a sudden change of the reactive power reference from $Q^* = 0$ to $Q^* = 4$ kVar, and Fig. 18 shows the response of the system to a balanced grid voltage dip of 70% of its nominal value (with a load

Fig. 16. Connection of a 4.2 kW DC-load with a null reactive reference. The grid current increases to supply the DC-load, maintaining the DC-bus voltage v_{DC} constant.

Fig. 17. Sudden change to a reactive power reference of $q^* = 4$ kVar with a null active power reference. The grid current quickly track the new reactive power demand.

of $P = 4.2$ kW). Damped system is proven to be quick and stable for all the aforementioned tests.

E. Performance comparison with related classic approaches

The performance of the proposed method is compared with the main two classical approaches to active damping that only measure one current to reduce the number of sensors needed: the notch filter approach [15], [22], [23] and the high-pass filter-lead-lag network- [21]. Both proposals have been long studied in literature. Authors have selected the most recent relevant articles on both approaches.

The dynamic of these two elements in the Laplace domain are as follows:

$$Notch(s) = \left(\frac{s^2 + 2D_z\omega_{nf}s + \omega_{nf}^2}{s^2 + 2D_p\omega_{nf}s + \omega_{nf}^2} \right)^n, \quad (20)$$

$$HPF(s) = \frac{k_{HPF}s}{s + \omega_{HPF}}. \quad (21)$$

Fig. 18. Initial response and recovery to a balanced grid voltage dip of 70%. The system is supplying a load of $P = 4.2$ kW. The grid current increased to compensate the power interchange decrease during the voltage dip.

The notch filter, *Notch*, is placed in series to the outer current controller (equivalent to: $K_u = \text{Notch}(s)$, $K_{v_s} = 0$, $K_{i_2} = 0$, $K_{v_c} = 0$, $K_{i_1} = 0$ in (8)), selecting $\omega_{nf} = \omega_{res}$ to avoid feeding the filter resonance with the VSC output voltage, but still actuating at the rest of frequencies. D_z is usually selected equal to zero, resulting in $|\text{Notch}(j\omega)| = 0$ at $\omega = \omega_{res}$. D_p will control the notch bandwidth: the wider is this bandwidth, the more robust the notch will be to variations of the filter resonance, affecting, in turn, other low control frequencies. In that regard, increasing the notch order n makes possible to increase the notch bandwidth without affecting so much other frequencies, increasing in turn the computational complexity of the damper. As it shown in [23], a good compromise is to select $n = 2$.

The high-pass filter $\text{HPF}(s)$ has two parameters: ω_{HPF} and k_{HPF} , that represent its bandwidth and maximum gain, respectively. This filter will feedback the grid current i_2 to modify the outer current controller output (i.e. similar to set $K_{i_2} = -\text{HPF}$ in the proposed damper $\mathbf{K}_{AD}$, with all its other terms set to zero in (8)).

The optimal design of these two proposals is usually based on the analysis of the resonant closed-loop poles trajectories for variations of their design parameters. Considering the notch filter, this process is relatively simple, as in practice it only depends on the optimal selection of D_p . On the other hand, the selection of optimal values for the high-pass filter is more complicated, not only because in this case the two design parameters affect the obtained performance, but also because their variation not always follow a *predictable* effect on the poles trajectories, depending on other factors as the sampling and resonant frequencies [21]. Moreover, the performance of the notch filter and the high-pass filter strongly depends on the design of the outer current controller, mainly on its bandwidth [21], [23].

Fig. 19 and Fig. 20 show, on the top, the frequency domain response of the admittance (Y) and tracking (T) closed-loop transfer functions, respectively, when the proposed PR outer current controller (K_{cc}) is implemented with a sample period

Fig. 19. Comparison of the closed-loop admittance (Y) and the current (i) response to a PCC voltage (v_s) dip for: the proposed active damper (red), the notch filter (green) and the high-pass filter (blue).

Fig. 20. Comparison of the closed-loop admittance (Y) and the current (i) response to a current reference (i^*) change for: the proposed active damper (red), the notch filter (green) and the high-pass filter (blue).

$T_s = 100 \mu s$, damping the filter resonance with the proposed damper ($\mathbf{K}_{AD}(s)$), the notch filter ($\text{Notch}(s)$) and the high-pass filter ($\text{HPF}(s)$). In addition, the bottom part of Fig. 19 and Fig. 20 show the time domain response of the grid current to a PCC voltage (v_s) dip and a current reference (i^*) change, respectively, for the three aforementioned methods. The design criterion for the two considered damping filters is similar and based on the analysis of the root locus of the system loop for variations of a proportional outer current controller (i.e. setting $K_{cc} = K_p$). The design parameters are then selected so the system is as robust as possible to variations of K_p , what usually results in a better damping of the filter resonance.

As it can be seen in Fig. 19 and Fig. 20 results, the proposed method shows a better performance than the previous approaches for the pre-designed PR controller (K_{cc}). Fig. 19 needs particular attention due to the small damping that is

achieved with notch-based active damping in the closed-loop admittance transfer function. This translates, as shown in Fig. 19, in a long oscillating transient in the voltage dip response. Nevertheless these oscillations eventually vanish and current reference tracking is never compromised. Additionally, Y_{HPF} suffers from steep phase shifts that may compromise the system passivity and, then, its stability with weak grids [33]. On the other hand, Fig. 20 shows that both damping filters have a larger overshoot than the proposed method for variations of the current reference (i^*) due to their smaller damping.

It can be concluded that, even though the proposed active damper may have a bigger computational design complexity than previous active damping proposals, it is compensated with the improvement in the performance and clearness for outer controller design.

Besides, the design of this proposal is mainly based on the selection of two simple frequency domain weights that, besides the control bandwidth (dependent of controller sampling frequency) does not depend on the LCL filter parameters (e.g. its resonance frequency) or the outer current controller to achieve the optimal performance. This could make the proposal valid for its implementation on a systematic tool with possible industry uses.

V. CONCLUSION

This paper has presented an active damping controller design for LCL-filter grid-connected VSC. It modifies grid current dynamic in order to mimic the one of an equivalent L filter, making possible the use of simple PI or PR controllers to track the desired current with minimum stability and performance penalties.

Proposed controller is obtained by (sub-)optimal $\mathcal{H}_\infty$ synthesis following a model reference design. The designer specifies the desired LCL grid current dynamic, in this case one equivalent to an L filter. The synthesis algorithm obtains, then, an optimal controller that complies with the specifications, resulting in a damped system. Design methodology is flexible enough to specify the number of inputs to the controller, considering different LCL or disturbances measurements, increasing its suitability for industry configurations. Limitations of the method are presented, being the most important a bandwidth limitation coming from the discrete-time implementation of the controller.

Obtained frequency domain results show the damping capabilities of the presented method. Stability margins are, then, minimally decreased if an L filter pre-designed grid current controller is applied to the modified LCL filter. In addition, damping and stability robustness of the method is proved for variations of the LCL filter and the grid impedance parameters. Finally, good time domain results are demonstrated by experimental tests. Damping performance of the method has been compared, with positive results, with related existing methods.

REFERENCES

- [1] J. Dannehl, F. W. Fuchs, S. Hansen, and P. B. Thøgersen, "Investigation of active damping approaches for PI-based current control of grid-connected pulse width modulation converters with LCL filters," *IEEE Transactions on Industry Applications*, vol. 46, no. 4, pp. 1509–1517, 2010.
- [2] M. Liserre, R. Teodorescu, and F. Blaabjerg, "Stability of photovoltaic and wind turbine grid-connected inverters for a large set of grid impedance values," *IEEE Transactions on Power Electronics*, vol. 21, no. 1, pp. 263–272, 2006.
- [3] L. Harnefors, L. Zhang, and M. Bongiorno, "Frequency-domain passivity-based current controller design," *IET Power Electronics*, vol. 1, no. 4, pp. 455–465, 2008.
- [4] B. Bahrani, M. Vasiladiotis, and A. Rufer, "High-order vector control of grid-connected voltage-source converters with LCL-filters," *IEEE Transactions on Industrial Electronics*, vol. 61, no. 6, pp. 2767–2775, June 2014.
- [5] F. Huerta, D. Pizarro, S. Cobrecas, F. J. Rodriguez, C. Giron, and A. Rodriguez, "LQG servo controller for the current control of lcl grid-connected voltage-source converters," *IEEE Transactions on Industrial Electronics*, vol. 59, no. 11, pp. 4272–4284, Nov 2012.
- [6] J. Kukkola, M. Hinkkanen, and K. Zenger, "Observer-based state-space current controller for a grid converter equipped with an LCL filter: Analytical method for direct discrete-time design," *IEEE Transactions on Industry Applications*, vol. 51, no. 5, pp. 4079–4090, Sept 2015.
- [7] R. Guzman, L. G. de Vicuña, J. Morales, M. Castilla, and J. Miret, "Model-based active damping control for three-phase voltage source inverters with LCL filter," *IEEE Transactions on Power Electronics*, vol. 32, no. 7, pp. 5637–5650, July 2017.
- [8] M. Liserre, F. Blaabjerg, and S. Hansen, "Design and control of an LCL-filter-based three-phase active rectifier," *IEEE Transactions on Industry Applications*, vol. 41, no. 5, pp. 1281–1291, 2005.
- [9] R. D. Middlebrook, "Input filter considerations in design and application of switching regulators," *Proc. of IAS Record*, 1976.
- [10] R. Pena-Alzola, M. Liserre, F. Blaabjerg, R. Sebastián, J. Dannehl, and F. W. Fuchs, "Analysis of the passive damping losses in LCL-filter-based grid converters," *IEEE Transactions on Power Electronics*, vol. 28, no. 6, pp. 2642–2646, 2013.
- [11] R. W. Erickson, "Optimal single resistor damping of input filters," in *Proc. of Applied Power Electronics Conference and Exposition*, vol. 2. Citeseer, 1999, pp. 1073–1079.
- [12] M. Liserre, A. Dell'Aquila, and F. Blaabjerg, "Stability improvements of an LCL-filter based three-phase active rectifier," in *Proc. of 33th Annual IEEE Power Electronics Specialists Conference, PESC*, vol. 3. IEEE, 2002, pp. 1195–1201.
- [13] V. Blasko and V. Kaura, "A novel control to actively damp resonance in input LC filter of a three-phase voltage source converter," *IEEE Transactions on Industry Applications*, vol. 33, no. 2, pp. 542–550, 1997.
- [14] E. Wu and P. W. Lehn, "Digital current control of a voltage source converter with active damping of LCL resonance," in *Proc. of Twentieth Annual IEEE Applied Power Electronics Conference and Exposition APEC 2005*, vol. 3. IEEE, 2005, pp. 1642–1649.
- [15] J. Dannehl, M. Liserre, and F. W. Fuchs, "Filter-based active damping of voltage source converters with filter," *IEEE Transactions on Industrial Electronics*, vol. 58, no. 8, pp. 3623–3633, 2011.
- [16] J. He and Y. W. Li, "Generalized closed-loop control schemes with embedded virtual impedances for voltage source converters with LC or LCL filters," *IEEE Transactions on Power Electronics*, vol. 27, no. 4, pp. 1850–1861, 2012.
- [17] S. G. Parker, B. P. McGrath, and D. G. Holmes, "Regions of active damping control for LCL filters," *IEEE Transactions on Industry Applications*, vol. 50, no. 1, pp. 424–432, 2014.
- [18] Y. Tang, P. C. Loh, P. Wang, F. H. Choo, and F. Gao, "Exploring inherent damping characteristic of LCL-filters for three-phase grid-connected voltage source inverters," *IEEE Transactions on Power Electronics*, vol. 27, no. 3, pp. 1433–1443, 2012.
- [19] J. L. Agorreta, M. Borrega, J. López, and L. Marroyo, "Modeling and control of paralleled grid-connected inverters with LCL filter coupled due to grid impedance in pv plants," *IEEE Transactions on Power Electronics*, vol. 26, no. 3, pp. 770–785, 2011.
- [20] C. A. Busada, S. G. Jorge, and J. A. Solsona, "Full-state feedback equivalent controller for active damping in lcl-filtered grid-connected inverters using a reduced number of sensors," *IEEE Transactions on Industrial Electronics*, vol. 62, no. 10, pp. 5993–6002, Oct 2015.
- [21] X. Wang, F. Blaabjerg, and P. C. Loh, "Grid-current-feedback active damping for LCL resonance in grid-connected voltage-source converters," *IEEE Transactions on Power Electronics*, vol. 31, no. 1, pp. 213–223, Jan 2016.
- [22] M. Liserre, A. D. Aquila, and F. Blaabjerg, "Genetic algorithm-based design of the active damping for an LCL-filter three-phase active

rectifier," *IEEE Transactions on Power Electronics*, vol. 19, no. 1, pp. 76–86, 2004.

- [23] R. Peña-Alzola, M. Liserre, F. Blaabjerg, M. Ordóñez, and T. Kerekes, "A self-commissioning notch filter for active damping in a three-phase LCL-filter-based grid-tie converter," *IEEE Transactions on Power Electronics*, vol. 29, no. 12, pp. 6754–6761, Dec 2014.
- [24] M. Orellana and R. Griño, "On the stability of discrete-time active damping methods for vsc converters with a LCL input filter," in *Proc. of 38th Annual Conference on IEEE Industrial Electronics Society IECON 2012*. IEEE, 2012, pp. 2378–2383.
- [25] M. Cespedes, L. Xing, and J. Sun, "Constant-power load system stabilization by passive damping," *IEEE Transactions on Power Electronics*, vol. 26, no. 7, pp. 1832–1836, 2011.
- [26] H. Mosskull, "Optimal stabilization of constant power loads with input LC-filters," *Control Engineering Practice*, vol. 27, pp. 61–73, 2014.
- [27] Q.-C. Zhong and T. Hornik, *Control of power inverters in renewable energy and smart grid integration*. John Wiley & Sons, 2012, vol. 97.
- [28] T. Hornik and Q.-C. Zhong, "Parallel PI voltage H-inf current controller for the neutral point of a three-phase inverter," *IEEE Transactions on Industrial Electronics*, vol. 60, no. 4, pp. 1335–1343, 2013.
- [29] G. Weiss, Q.-C. Zhong, T. C. Green, and J. Liang, "H-inf repetitive control of DC-AC converters in microgrids," *IEEE Transactions on Power Electronics*, vol. 19, no. 1, pp. 219–230, 2004.
- [30] S. Cobreces, E. J. Bueno, F. J. Rodríguez, D. Pizarro, and F. Huerta, "Robust loop-shaping h-inf control of lcl-connected grid converters," in *Proc. of IEEE International Symposium on Industrial Electronics (ISIE)*. IEEE, 2010, pp. 3011–3017.
- [31] R. Naim, G. Weiss, and S. Ben-Yaakov, "H-inf control applied to boost power converters," *IEEE Transactions on Power Electronics*, vol. 12, no. 4, pp. 677–683, 1997.
- [32] T.-S. Lee, S. Chiang, and J.-M. Chang, "H-inf loop-shaping controller designs for the single-phase ups inverters," *IEEE Transactions on Power Electronics*, vol. 16, no. 4, pp. 473–481, 2001.
- [33] L. Harnefors, X. Wang, A. Yepes, and F. Blaabjerg, "Passivity-based stability assessment of grid-connected VSCs-an overview," *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 4, no. 1, pp. 116 – 125, 2015.
- [34] P. C. Krause, O. Wasynczuk, and S. D. Sudhoff, *Analysis of Electric Machinery and Drive Systems. 2nd Edition*. IEEE Press. Wiley-Interscience, 2002.
- [35] S. Skogestad and I. Postlethwaite, *Multivariable feedback control: analysis and design*. Wiley New York, 2007, vol. 2.
- [36] J. Pérez, S. Cobreces, R. Griño, and F. Rodríguez, " $\mathcal{H}_\infty$ current controller for input admittance shaping of VSC-based grid applications," *IEEE Transactions on Power Electronics*, vol. 32, no. 4, pp. 3180–3191, 2016.

Jorge Pérez (S'16-M'17) Jorge Pérez was born in Madrid, Spain, in 1988. He received B.Sc, M.Sc and PhD degree in industrial electronics from University of Alcalá (UAH), Madrid, Spain, in 2011, 2013 and 2017 respectively. Until 2017, he was a researcher for GEISER Group of the UAH. He is currently with Gamesa Electric S.A.U, Madrid, Spain. His research interests include automatic control applied to power electronic systems, power quality, and distributed power generation systems.

automatic control and system identification applied to power electronic systems power quality, and distributed power generation systems.

nonlinear control and control of power electronics converters. Dr. Griño is an affiliate member of International Federation of Automatic Control (IFAC) and a member of the Spanish Society on Automation and Control-IFAC.

power electronics converters, power quality and distributed power generation systems.

a Professor. He is the author of more than 142 refereed publications in international journals, book chapters, and conference proceedings. Also, he has directed more than 45 investigation projects funded by public institutions and private industry. His research interests include the areas of control electronics, real-time processing, and embedded systems applied to power electronic systems.

Santiago Cobreces (S'03-M'09) Santiago Cobreces was born in Alcalá de Henares, Spain, in 1980. He received the B.Sc. and He received the M.Sc. degree in telecom engineering and the Ph.D. degree in electronics engineering from the University of Alcalá, Alcalá de Henares, Spain, in 2003 and 2009, respectively. Since 2012, he is an Associate Professor with the Department of Electronics, University of Alcalá, where he is a member of the Research Group "Electronics Engineering Applied to the Renewable Energies." His current research interests include automatic control and system identification applied to power electronic systems power quality, and distributed power generation systems.

Robert Griño (M'99-SM'12) Robert Griño received the M.Sc. degree in electrical engineering and the Ph.D. degree in automatic control from the Universitat Politècnica de Catalunya (UPC), Barcelona, Spain, in 1989 and 1997, respectively. From 1990 to 1991, he was a Research Assistant with the Instituto de Cibernética, UPC. From 1992 to 1998, he was an Assistant Professor with the Automatic Control Department, Universitat Politècnica de Catalunya, where he has been an Associate Professor since 1998. His research interests include digital control,

Francisco Huerta (S'08-M'11) Francisco Huerta received the M.Sc. degree and the Ph.D. degree in electronics engineering from the University of Alcalá, Alcalá de Henares, Spain, in 2006 and 2011, respectively. He worked from 2007 and 2012 with the University of Alcalá. From 2013 until 2016, he was a Postdoctoral Researcher with IMDEA Energy. From 2016 to 2017, he worked as a Researcher with the University of Alcalá. Currently he is an Assistant Professor with the Carlos III University of Madrid. His research interests include control of

Francisco J. Rodríguez (S'99-M'00) Francisco J. Rodríguez received the B.Sc. degree in technical telecommunication engineering from the University of Alcalá, Alcalá de Henares, Spain, in 1985, the M.Sc. degree in telecommunication from the Technical University of Madrid, Madrid, Spain, in 1990, and the Ph.D. degree in electronics engineering from the University of Alcalá in 1997. He worked in the private electronic industry for two years. Since 1986, he has been a Lecturer with the Department of Electronics, University of Alcalá, where now he is