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Robust multisampled capacitor voltage active damping for grid-connected power converters

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ABSTRACT

The derivative feedback of the capacitor voltage is one of the most extended active damping strategies, used to eliminate stability problems in grid-connected power converters with an *LCL* filter. This strategy is equivalent to the implementation of a virtual impedance in parallel with the filter capacitor. This virtual impedance is strongly affected by the control loop delays and frequency, creating changes in the sign of the emulated virtual resistor, and raising instability regions where the active damping is ineffective. As a consequence, the *LCL* resonance frequency is restricted to vary, as the effective grid inductance changes, within the active damping stability region. This is an additional restriction imposed on the *LCL* filter design that can compromise the achievement of an optimised design. For this reason, in this work, a different strategy is presented; by adjusting the delay in the active damping feedback path, it becomes stable within the range where the *LCL* resonance frequency can be located for a given filter design, achieving a robust damping. Analytical expressions are provided to adjust this delay. To widen the stability region of the capacitor voltage derivative active damping, a multisampled derivative is implemented, overcoming its limitations close to the control Nyquist frequency. Experimental and simulation results validate the active damping strategy presented.

1. Introduction

Grid-connected voltage source power converters (VSC) are largely used as an interface with the grid for renewable power systems [1–4]. As the number of power converters increases, stringent grid codes regulating the power quality injected have been developed [5]. A common approach to meet these grid codes is the use of an *LCL* filter, reducing its overall size and cost if compared with the *L* filter [6]. In this filtering topology, the resonance between the filter and grid impedances must be damped, as stability issues arise otherwise due to the effects of the delays in the control loop [7]. To solve these stability problems, both passive and active damping methods are used in the existing literature [8,9].

There are several passive damping approaches, consisting in the addition of damping resistors that increase the converter power losses. Even though some damping topologies present reduced power losses [10], they introduce other passive components and the filter complexity grows.

Active damping (AD) approaches have been widely explored in previous papers as they can stabilise the system without increasing its power losses. The capacitor voltage feed-forward can effectively damp the filter resonant poles, but it becomes ineffective as the *LCL* resonance frequency approaches the converter control Nyquist frequency [11]. In high power converters, where the switching frequency, and accordingly the sampling frequency, is limited to reduce the power losses, and the resonance frequency is increased to lower the filter size, a different AD strategy is required. A notch filter, inserted in the current control loop and tuned at the resonance frequency, could damp the resonance [12], but it requires an estimation of the effective grid inductance, as it can change depending on the grid at which the VSC is connected and the power injected at the point of common coupling (PCC) [13,14]. Alternatively, a lead-lag controller can be tuned in the current control loop to avoid -180 degree crossings [12] that can lead to instability. However, when the resonance frequency approaches the converter control Nyquist frequency, it is unable to introduce enough phase lead. A suitable option to overcome this limitation is the introduction of additional delays and low-pass filters [15,16], which are able to stabilise the resonant poles. Nevertheless, they provide a poor damping at the resonant poles, compromising the grid current harmonic content and, therefore, the fulfillment of the grid codes [16]. The capacitor current proportional feedback is one of the preferred solutions, equivalent to the implementation of a virtual resistor in parallel with

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the filter capacitor [17–21]. The main drawback is that it requires additional sensors as this current is not normally measured a in grid-connected VSC. An alternative is using the capacitor voltage, measured for synchronisation purposes, using its derivative to estimate its current and performing an active damping strategy equivalent to the previous one [22–24].

Both, the capacitor current feedback and the capacitor voltage derivative AD strategies, are based on the emulation of a virtual damping resistor, which becomes a virtual impedance by the effects of the control delays [19,20]. The real part of the emulated virtual impedance varies with frequency and it can become negative, leading to instability if the resonance frequency is located in the negative region. This issue has been reported in the literature: in [17] the stability region where the capacitor current AD can effectively damp the resonance is limited to $\omega_s/6$, ω_s being the sampling frequency. The stability region calculated imposes constraints on the *LCL* filter design: the *LCL* resonance has to be lower than $\omega_s/6$. This restriction compromises the achievement of an optimised filter in order to meet the grid codes at the lowest price. In [19] they identified the same stability limits, suggesting that the resonance frequency should be limited to be lower than $\omega_s/6$, where the emulated virtual resistance is positive, so that the AD can stabilise the resonant poles. Alternatively, they proposed reducing the computation delay to widen the stability interval. Lastly [20], proposed an *RC* virtual damper that modifies the stability region, changing the feedback sign in order to operate at the region where the emulated virtual resistor is negative. In this case the resonance frequency is restricted to a wider interval limited by $\omega_s/5$ and $\omega_s/2$.

In commercial high power converters, as the one further described in this work, the converter side inductance is around 0.1 p.u., the filter capacitor is 0.03 p.u., while the grid side inductance is formed by the transformer leakage inductance, 0.05 p.u and the effective grid inductance, which is unknown and would modify the filter resonance frequency. The resonance frequency, for the values provided, is bounded within $0.15\omega_s$ and $0.27\omega_s$, as the grid inductance varies from a short circuit ratio (SCR) at the PCC of 1–300. This interval of possible resonance frequencies does not fall within the stability regions identified by the previous papers. To solve this issue, in this article, a different approach is presented. Instead of designing the filter to locate the resonance frequency where the AD is able to stabilise the resonant poles, the AD stability region is modified and adapted to be robust and stable within the range of frequencies where the *LCL* resonant poles can be located for a given power converter design. With this purpose, the delay in the AD path is modified after an analysis of the existing delays in the control loop, considering the filters applied to the measurements. These filters are neglected in the literature, even though they strongly affect the AD stability region and must be introduced to avoid noise amplification by the derivative. Analytical expressions are provided for the adjustment of the delay, which are valid for the capacitor current proportional feedback and the capacitor voltage derivative AD (CVDAD). However, this paper is focused on the latest strategy, as it avoids the use of additional sensors.

The discrete implementation of the derivative close to the control Nyquist frequency is not possible without magnitude and phase distortion. For this reason, a multisampled derivative is used, as done in the passivity based analysis performed in [25]. With the multisampled approach the delay is reduced in the feedback path, achieving a wider stability interval in the CVDAD, as it will be shown in this work. Experimental results are provided to validate the proposed AD approach.

2. System modelling and stability analysis

2.1. System modelling and control

Fig. 1 shows the typical structure of a VSC connected to the grid with an *LCL* filter. This filter is formed by the converter inductance, L_c, the filter capacitor, C_f, the step-up transformer leakage inductance, L_{transf}, and the grid inductance, L_g at the PCC. The latter is usually unknown and variable, modifying the filter grid side effective inductance.

The converter side current is usually controlled by means of a proportional-integral (PI) controller in the synchronous reference frame (SRF). For this reason, the *LCL* filter model will be developed in the *dq* components. With this transformation, cross-coupling terms appear between the *d* and *q* phases in the reactive passive components, and thus the system has to be treated as a multiple-input multiple-output (MIMO) system, Fig. 2. In some cases, a simplified model is developed, neglecting some of the cross-couplings, but this model does not reproduce accurately the system dynamics and stability. In Fig. 2 it can be seen that the cross-coupling terms between both axis are the same but with an opposite sign, so the matrix representing the system dynamics will have the form and symmetry of Eq. (1). In Eq. (1) the converter current $I_{dq}(s)$ is correlated to the voltage imposed by the converter $V_{conv_{dq}}(s)$. This symmetry will be important for the stability analysis performed at the end of this section.

$$\begin{bmatrix} I_d(s) \\ I_q(s) \end{bmatrix} = \begin{bmatrix} G_1(s) & G_2(s) \\ -G_2(s) & G_1(s) \end{bmatrix} \begin{bmatrix} V_{conv_d}(s) \\ V_{conv_q}(s) \end{bmatrix} \tag{1}$$

The expression for $G_1(s)$ and $G_2(s)$ are given in Eqs. (2) and (3) respectively, reflecting that a sixth-order model is obtained. In these equations $\omega_r = 1/\sqrt{C_f L}$ is the filter resonance frequency, where $L = \sqrt{L_c L_{gt}/(L_c + L_{gt})}$, with L_{gt} equal to the sum of the transformer leakage inductance, L_{transf}, and the grid inductance, L_g. The parameters a and b are the ratios between ω_{par}/ω_r and ω_0/ω_r, respectively, where ω_{par} is the parallel resonance frequency between the grid side inductance and the filter capacitor and ω_0 is the frequency used for the SRF transformation. Lastly, Q represents the filter quality factor, defined as R_{da}/R_0, where R_{da} is a passive resistor connected in series with the filter capacitor to damp the resonance and R_0 is $\sqrt{L/C_f}$.

In Fig. 3 a schematic of the converter control loop is represented. The capacitor voltage and the converter current measurements are filtered by a first-order low-pass analog filter (LPAF) with a time constant τ_{lp}. These measurements are sampled by the DSP, where the control loops are executed twice per converter switching period. The PLL, receiving the capacitor voltage filtered measurement through a SOGI filter [26], provides the angle for the transformation to the SRF. Each current component is controlled by a PI controller. At the output of the controller, a feed-forward compensation of the capacitor voltage is added through a low-pass digital filter (LPDF). The sum of the controller and feed-forward voltages is saturated to avoid overmodulation, using an anti-windup (AW) for the PI.

The Laplace domain has been chosen to model the system because it is a convenient option to analyze it when the active damping with the multisampled approach is introduced. The Laplace equivalent transfer functions for the digital elements within the control loop, such as the feed-forward LPDF and the PI controller, are used. To properly study the system stability, a representation of all the elements in the same reference frame is required, in this case, the SRF. The LPAF is applied to the real magnitudes of the variables. This filter can be directly expressed in the stationary frame, $\alpha\beta$, without modifying the filter transfer function. In the control loop described in Fig. 3, elements defined in the stationary frame $\alpha\beta$, i.e. D_{conv} and the transformation of LPAF to $\alpha\beta$, coexist with elements in the SRF, i.e. the PI current controller and the LPDF. $D_{conv}(s)$ stands for the computation delay in the DSP, and the zero order hold (ZOH). These elements in $\alpha\beta$ are transformed to the SRF using the transformation proposed in [27]. The LPAF applied in the $\alpha\beta$ frame to the measured variables is shown in Eq. (4) for the converter current case, where $LPAF(s)$ is $1/(\tau_{lp}s + 1)$.

$$\begin{bmatrix} I_{\alpha,f}(s) \\ I_{\beta,f}(s) \end{bmatrix} = \begin{bmatrix} LPAF(s) & 0 \\ 0 & LPAF(s) \end{bmatrix} \begin{bmatrix} I_\alpha(s) \\ I_\beta(s) \end{bmatrix} \tag{4}$$

Eq. (4) is transformed to Eq. (5), its equivalent in the SRF.