Analysis and Validation for an Inverter-side Current Controller in LCL Grid-connected Power Systems

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Abstract—Transformerless grid-connected inverters offer greater efficiencies when transferring power from renewable energy sources to the electrical grid. If the grid-inverter connection is done with an LCL filter, high attenuation of switching harmonics is achieved while preserving a small-size output filter. However, damping must be included in the controller to assure closed-loop stability. This paper proposes a reference computation methodology for the inverter-side current feedback in a photovoltaic (PV) generation system connected to the grid through an LCL filter. Theoretical analysis of the closed-loop system stability and of the steady-state performance are presented as well as experimental validation of the closed-loop performance. The feedback controller includes active damping and relies on a resonant control structure which improves the ability of dealing with grid harmonic distortion. The controller uses a reduced set of measurements, which requires the inverter-side current and grid voltage only, and assures a power factor close to unity.

Index Terms—LCL filter, transformerless inverter, grid connection, inverter-side current feedback.

I. INTRODUCTION

Transformerless grid-connected inverters offer a notorious usage among residential and industrial facilities. A noteworthy application that is gaining increasing interest is the connection to the grid of renewable sources such as photovoltaic (PV), wind turbine or hybrid generation systems [1]-[3]. In these applications, the electric energy is delivered by compelling a grid-side current and grid voltage only, and assures a power factor close to unity.

Several controllers providing active damping to the closed-loop system have been proposed so far. Most of them are based on the feedback of the grid-side or inverter-side currents and the feedback of the voltage or current of the LCL filter capacitor [7], [11]-[15]. Other approaches consider a reduced number of measured states by avoiding the measurement of the voltage or current of the LCL filter capacitor. In particular, two main control design approaches can be distinguished to assure active damping. The first approach relies on the feedback of the inverter-side current [7], [11], [14], while the second relies on the feedback of the grid-side current [12], [16], [17]. In the grid-side current feedback, the damping is added only under certain conditions on the LCL filter parameters and switching frequency. This approach may even present instability issues in the presence of weak grid conditions in [17]. In contrast, the inverter-side current
feedback adds sufficient damping to mitigate the resonant behavior of the LCL filter [7], [15] in a broader range of parameters, switching frequency and grid conditions. However, the computation of the inverter-side current reference following this approach is more elaborated. An interesting alternative has been presented in [14], which uses a reduced-order observer to realize a full state-feedback control, where the inverter-side current is the only available state.

This paper analyzes and experimentally validates an inverter-side current controller for power inverters connected to the grid through an LCL filter. The controller consists of a proportional term over the inverter-side current error, and a harmonic compensation mechanism, which is embodied in a bank of resonant filters tuned at the fundamental and harmonic frequencies and also operating on the inverter-side current error. The proportional term adds the required damping, while the resonant control structure compensates the effect of the grid voltage harmonic distortion to assure a high-quality grid current. This controller guarantees the tracking of the inverter-side current. Hence, it is important to provide an appropriate inverter-side current reference. This reference must be designed to assure that the grid-side current follows a sinusoidal waveform of the grid fundamental frequency component only, in phase with the grid voltage. The key to design such a reference is the use of a scheme that extracts the fundamental component of the grid voltage and its square-phase companion. The analysis shows that the inverter-side current controller guarantees asymptotic stability.

Experimental tests are performed in a three-level inverter prototype for simplicity. However, the controller can be extended to other inverter topologies. The experimental results show that the inverter, with the controller and the proposed current reference computation, supplies active power to the grid with a good performance and enough damped response despite the harmonic distortion in the grid voltage. Moreover, the solution uses a reduced number of sensors. The main contribution of the paper is the design of an inverter-side current controller to guarantee active damping for LCL grid-connected H-bridge (HB) based converters along with the corresponding stability analysis. The controller is able to achieve active power injection with guaranteed active damping by using a design method to obtain the references for the inverter-side current and the injected voltage. In addition, the corresponding modifications to these references are also included to deal with the more realistic scenario that considers harmonic distortion in the grid voltage.

II. SYSTEM MODELING

A general representation of a power inverter connected to the grid through an LCL filter to supply power to the grid is depicted in Fig. 1. The inductors $L_1$ and $L_2$ and the capacitor $C$ form the LCL filter between the power inverter and the grid. The grid is represented by a sinusoidal voltage source $V_g$ with fundamental angular frequency $\omega_g$ that may contain harmonic distortion.

![Fig. 1. Power inverter connected to grid through LCL filter.](image)

Let $i_s$ and $i_c$ denote the currents through $L_1$ and $L_2$, respectively, namely the inverter-side and the grid-side currents, and let $v_c$ denote the voltage across $C$. The model of the system in Fig. 1, is described by the following equations:

$$L_1 \frac{di_s}{dt} = -R_1 i_s - v_c + e$$  \hspace{1cm} (1)

$$L_2 \frac{di_c}{dt} = -R_2 i_s + v_c - v_g$$  \hspace{1cm} (2)

$$Cd\frac{dv_c}{dt} = i_i - i_s$$  \hspace{1cm} (3)

where each inductor is represented by an inductance $L$, followed by a parasitic resistor $R_i$, $i=1,2$. The main objective is to supply active power to the grid by forcing the grid-side current $i_c$ to have a pure sinusoidal waveform in phase with the fundamental component of the grid voltage $v_g$. In order to comply with this purpose, the voltage source inverter (VSI) must reconstruct an appropriate voltage signal $e$, referred to as the inverter voltage and the control objective all along the paper.

In the present work, it is assumed that the switching frequency is high enough, that is, the cutoff frequency of the LCL filter is selected well below the switching frequency $f_{sw}$. In this way, the ripple in all signals can be neglected, and $e$ can be regarded as a continuous signal. In fact, it can be calculated as $e = u V_{dc}$ as shown in Fig. 1, where $u \in [-1,1]$ is considered to be continuous and represents the duty ratio of a pulse width modulation (PWM) signal.

The inverter-side current controller, to be described in the next section, can be used without major modifications in several renewable based systems that supply electric energy to the grid. Furthermore, the VSI in Fig. 1 could be any single-phase suitable topology, which is the case of HB-based topologies such as the H5 [18], the HERIC [19], the H6 [20] or the ZVFBR [21] topology. A PV system is taken as an example, where the direct-current (DC) power supply is replaced by an array of PV panels. In this case, the controller must be enhanced with a regulation loop to maintain the DC-link voltage in an average reference value $V_{dc}$ dictated by an additional maximum power point tracking (MPPT) algorithm.

In general, the inductor series parasitic resistors $R_1$ and $R_2$ shown in Fig. 1 add damping to the system, which is in benefit of the stability. The control design shown in the next section neglects such parasitic resistors, making the problem
more challenging as the damping must be provided only by the controller. Notice that the third-order state-space representation of the system model (1) - (3) is a linear system. Therefore, it admits a transfer function representation, which governs the open-loop performance, facilitates the design of the LCL filter, and allows the computation of its resonant frequency. This information is then used to suggest a switching frequency where the inverter can operate normally. If $R_i$ and $R_2$ are neglected, then the transfer functions $G_i(s) = I_i(s)/E(s)$, $G_v(s) = I_S(s)/E(s)$ and $G_{RC}(s) = V_c(s)/E(s)$ are given by:

$$G_i(s) = \frac{s^2 + 1/(L_2C)}{(L_1A(s))}$$ (4)

$$G_v(s) = \frac{1}{(L_1L_2C)(A(s))}$$ (5)

$$G_{RC}(s) = \frac{s}{(L_1CA(s))}$$ (6)

where the characteristic polynomial $A(s)$ and resonance frequency $\omega_{res}$ are given by:

$$A(s) = s^3 + \left(\frac{L_1 + L_2}{L_1L_2C}\right)s$$

$$\omega_{res} = \sqrt{\frac{L_1 + L_2}{L_1L_2C}}$$ (7)

A common design of the LCL filter, intended to mitigate switching frequency components and to reduce the filter size [2], [22], [23], consists in computing the minimum value of $L_1$ to guarantee the required inverter-side current ripple. Afterwards, the LCL filter capacitor $C$ is selected based on the maximum reactive power handled by the capacitor. It is common in practice to fix the reactive power handled by $C$ between 5% and 15% of the rated active power supplied to the grid. However, a compromise arises between the amount of the capacitor reactive power and the filter ability to attenuate the switching ripple. On one hand, if the capacitance is large, the handled reactive power is large as well. On the other hand, if the capacitance is small, the attenuation ability of the LCL filter diminishes. Finally, the grid-side inductor value is selected to meet the required ripple attenuation at the grid-side current, and also to locate the LCL resonance frequency $\omega_{res}$ in (7) between $10\omega_C$ and $\pi f_{sw}$.

### III. ANALYSIS AND CONTROL DESIGN

This section presents the design of the inverter-side current controller to achieve active damping for the LCL grid connected inverter. The controller is based on the inverter-side current feedback to damp the system resonance. It includes a proportional term over the inverter-side current error. Therefore, an appropriate design of the inverter-side current reference is essential to accomplish the main control objective, i.e., to compel a proper grid-side current. A resonant control structure [24], [25] is also included in the controller, which operates on the inverter-side current error as well. The aim of the resonant structure is to compensate harmonic distortion that may be present in the grid, which is well supported by the internal model principle [26]. It is worth mentioning that the controller addressed herein requires a reduced number of sensors. In fact, only grid voltage and inverter-side current are measured. The design of the controller is presented in the following subsections. First, the current error feedback is addressed, and the arguments on stability and dynamics are presented. Second, the computation of the inverter-side current reference is presented, which relies on the estimation of the fundamental frequency component of the grid voltage and its square-phase companion. An analysis of the steady state of the current reference computation scheme is performed to bound the error that arises when the grid voltage has harmonic distortion. Finally, a resonant control structure is introduced to compensate grid frequency harmonics.

#### A. Current Error Feedback

In this section, the error system is obtained by considering an output feedback, where only the inverter-side current is measured. The controller consists of a proportional gain for which an stability analysis is performed. A resonant control structure is added to the main controller to deal with possibly harmonic distortion in the grid voltage.

The controller operates on the inverter-side current error, which is the difference between the measured and an appropriately designed reference of the inverter-side current. The following assumptions are considered to simplify the error dynamics in the controller derivation.

1) The grid fundamental frequency has a known constant value of $f_C = \omega_C/(2\pi)$.

2) Parasitic resistors $R_1$ and $R_2$ have an arbitrarily small value and can be neglected in the system dynamics (1)-(3).

If $\omega_C$ is unknown, an estimator such as the one presented in [27] can be used. Notice that the second assumption above poses a more challenging control problem, as the missing damping must be introduced only by the controller.

To supply only active power to the grid, $i_S$ must asymptotically track a reference $i_S^{\text{ref}}$, which is composed of fundamental components only and is in phase with the grid voltage. The amplitude of $i_S^{\text{ref}}$ is proportional to the amount of power to be supplied to the grid, namely, the power reference $P_{ref}$. This control objective is denoted by $i_S \to i_S^{\text{ref}}$ as $t \to \infty$, and must be fulfilled with uniform stability.

Consider $i_S^{\text{ref}}$ to be an admissible trajectory for $i_s$, then $i_s$, $v_C$, and $e$ track references $i_s^{\text{ref}}$, $v_C^{\text{ref}}$, and $e^{\text{ref}}$, respectively, which are governed by system dynamics (1)-(3). Define the error state variables as $\tilde{i}_1 = i_1 - i_s^{\text{ref}}$, $\tilde{i}_2 = i_2 - i_s^{\text{ref}}$, $\tilde{v}_C = v_C - v_C^{\text{ref}}$ and $u = e - e^{\text{ref}}$. Based on these error definitions, the system (1)-(3) is transformed into the following system, which is referred to as the error system:

$$L_1 \frac{d\tilde{i}}{dt} = -\tilde{v}_C + u$$ (8)
\[ L_c \frac{d\tilde{v}_c}{dt} = \tilde{v}_c \]
\[ C \frac{d\tilde{v}_c}{dt} = \tilde{i}_1 - \tilde{i}_s \]

The system described in (8)-(10) is linear and time-invariant, and has a pole at the origin and two complex conjugate poles. The magnitude of the latter equals the resonant frequency \( \omega_{res} \) in (7). In order to fulfill the control objective, the above error system must have the origin as an asymptotically stable equilibrium point. Then, the following controller is proposed, which relies on the feedback of \( \tilde{i}_1 \), which is expressed in (11).

\[ u = -k\tilde{i}_1 \]

where \( k \) is a strictly positive constant gain. The closed-loop error system (8)-(11) can be rewritten in the form of an autonomous system as follows:

\[ Q \frac{d\tilde{x}}{dt} = \hat{A}\tilde{x} \]

\[ Q \frac{d\tilde{x}}{dt} = \begin{bmatrix} -k & 0 & -1 \\ 0 & 0 & 1 \\ 1 & -1 & 0 \end{bmatrix} \tilde{x} \]

where \( Q = \text{diag}(L_1, L_2, C) \) is the diagonal matrix with filter parameters in its diagonal, and the error state vector is given by \( \tilde{x} = \begin{bmatrix} \tilde{i}_1, \tilde{i}_s, \tilde{v}_c \end{bmatrix}^T \). The closed-loop state matrix is given by (14).

\[ Q^{-1} \hat{A} = Q^{-1} \begin{bmatrix} -k & 0 & -1 \\ 0 & 0 & 1 \\ 1 & -1 & 0 \end{bmatrix} \]

The model in (13) represents the closed-loop error average model in time domain with inverter-side current error feedback. The roots of its closed-loop characteristic polynomial in the frequency domain establishes stability. Closed-loop polynomial together with the related transfer function is used to prove closed-loop stability with the inverter-side current error feedback by appealing to the root locus method. The filter inductor ratio is defined by \( \alpha = L_2 / L_1 \). Then the closed-loop characteristic polynomial is given by (15) and its roots are the closed-loop poles, or equivalently, the eigenvalues of \( Q^{-1} \hat{A} \).

\[ A(s) = s^3 + \frac{1+\alpha}{aL_1C} s + k \left( \frac{1}{L_1} s^2 + \frac{1}{aL_1^2C} \right) \]

Then, the closed-loop poles coincide with the solutions of \( 1 + kG(s)=0 \), where \( G(s) \) is expressed in (16).

\[ G(s) = \frac{1}{L_1} s^2 + \frac{1+\alpha}{aL_1^2C} \]

The open-loop poles of \( G(s) \) are located at zero and at \( \pm j\sqrt{(1+\alpha)/(aL_1C)} \), and the zeros at \( \pm j\sqrt{1/(aL_1C)} \). The root locus of the closed-loop poles as a function of gain \( k \), which varies from zero to infinity, may adopt either of the sketches shown in Fig. 2. For \( k = 0 \), the closed-loop poles coincide with the open-loop poles. As \( k \) increases, the closed-loop poles follow the root loci of Fig. 2. The value of the inductor ratio \( \alpha = L_2 / L_1 \) relative to 8 determines the behavior of the root loci as \( k \) varies. Figure 2 shows that if the gain \( k \) is positive, then the closed-loop poles have negative real parts, which guarantees closed-loop stability. Additionally, the value \( \alpha = L_2 / L_1 \) differentiates the performance of several stable closed-loop systems. This has an effect on the performance of the transient response rather than on the stability. In practice, the case \( \alpha = 8 \) or even \( \alpha > 8 \) rarely appears. In fact, \( L_1 \) is usually designed to be larger than \( L_2 \) as the inverter-side current gets larger switching ripple. If \( \alpha \geq 8 \), then there exist break-in or breakaway points of the closed-loop poles to the real line as observed in Fig. 2(b) and Fig. 2(c). This is not the case for \( \alpha < 8 \) as shown in Fig. 2(a). That is, for \( \alpha \geq 8 \), there are discrete values of \( k \) for which multiple closed-loop real poles occur. Such values of \( k \) can be obtained by solving the critical points of function \( f(s) = 1 + kG(s) \) [28].

![Fig. 2](image-url)
If $a<8$, $L_2<8L_1$, and the root locus is depicted in Fig. 2 (a) where one closed-loop pole has negative real part for any $k>0$ and tends to a zero at $-\infty$ as $k \to \infty$. The two remaining closed-loop poles are complex conjugates with negative real part for any $k>0$, and tend to the zeros located in the imaginary axis at $\pm j\sqrt{1/(L_1C)}$ as $k \to \infty$. This is a typical root locus that occurs when the value of the inverter-side inductor is larger than that of the grid-side inductor. This selection is done by considering that the switching harmonic content is larger in the inverter-side current than in the grid-side current.

If $a=8$, $L_2=8L_1$, and the root locus is shown in Fig. 2(b). In this case, there is one negative real closed-loop pole that tends to the zero located at $-\infty$, and there are two closed-loop poles that have negative real part and tend to the zeros located in the imaginary axis as $k \to \infty$. There is a point in the real line $s_j=-(\sqrt{6}/4)\sqrt{1/(L_1C)}$ as shown in Fig. 2(b), where the three closed-loop poles coincide with feedback gain $k$, expressed in (17).

$$k = \frac{3\sqrt{6}}{4} \frac{L_1}{C}$$ (17)

If $a>8$, $L_2>8L_1$, and the root locus is shown in Fig. 2(c). One closed-loop pole is real and tends to $-\infty$ when $k \to \infty$. The other two closed-loop poles have negative real part and tend to the zeros located at $\pm j(\sqrt{a/\alpha})\sqrt{1/(L_1C)}$ in the imaginary axis as $k \to \infty$. However, for the subset of gains $[k_1, k_2] \in \mathbb{R}$, all closed-loop poles become negative real. The gains $k_1$ and $k_2$ are expressed in (18).

$$\begin{cases} k_1 = \frac{\sqrt{2a(a-2)+\sqrt{a(a-8)^2}}}{2a} \frac{3a+\sqrt{a(a-8)}}{a+\sqrt{a(a-8)}} \frac{L_1}{\sqrt{C}} \\ k_2 = \frac{\sqrt{2a(a-2)-\sqrt{a(a-8)^2}}}{2a} \frac{3a-\sqrt{a(a-8)}}{a-\sqrt{a(a-8)}} \frac{L_1}{\sqrt{C}} \end{cases}$$ (18)

There exists an entry/exit to/from the real line of the complex conjugated closed-loop poles, occurring at $s_{k_1}$ and $s_{k_2}$, respectively, expressed in (19).

$$\begin{cases} s_{k_1} = -\frac{\sqrt{2a(a-2)+\sqrt{a(a-8)^2}}}{2a\sqrt{L_1C}} \\ s_{k_2} = -\frac{\sqrt{2a(a-2)-\sqrt{a(a-8)^2}}}{2a\sqrt{L_1C}} \end{cases}$$ (19)

From this analysis, it can be concluded that, for any of the possible cases with $k>0$, the asymptotic stability of the origin is guaranteed since closed-loop poles have negative real parts. However, the transient response depends on different ratios $\alpha$ and feedback gains $k$. Therefore, the inverter-side current feedback (11) adds active damping by moving the closed-loop poles to the left-side of the complex plane as the gain parameter $k$ increases. For a relatively small $k$, the response is dominated by the real closed-loop pole closest to the origin, where the response is well damped. For a relatively large value of $k$, the response behavior is dominated by complex conjugated closed-loop poles, which dictates an oscillatory behavior of the closed-loop system. In either case, an increase in gain $k>0$, in a given range, causes the reduction of the imaginary part of the complex conjugated closed-loop poles in each of the cases depicted in Fig. 2. As a consequence, a more damped behavior is expected, with a damping coefficient depending on the system parameters.

Therefore, with the inverter-side current feedback (11), and taking into consideration the inverter voltage reference $e^{\text{ref}}$, the inverter voltage $e=u+e^{\text{ref}}$ is given by (20).

$$e = -k_i^1 + e^{\text{ref}}$$ (20)

Under the above assumptions, the feedback controller (20) renders the closed-loop system stable. The computation of $e$ requires $i_1^{\text{ref}}$ and $e^{\text{ref}}$, where $e^{\text{ref}}$ can be seen as a decoupling term. The next subsection presents a method to obtain estimates of $i_1^{\text{ref}}$ and $e^{\text{ref}}$, in the case where the grid voltage and $i_1^{\text{ref}}$ are pure sinusoidal signals, i.e., contain fundamental component only. The case with harmonic distortion in the grid voltage requires a harmonic compensation mechanism, which is presented right after the next subsection.

B. Definition and Estimation of Admissible Reference

Implementation of the above controller (20) requires the reference $i_1^{\text{ref}}$ and the decoupling term $e^{\text{ref}}$ to accomplish the control objective, i.e., to assure that $i_1$ is a pure sinusoidal signal with the same frequency and phase of the fundamental component of $v_S$. This subsection presents the estimation for these two references. For the ease of presentation, this subsection considers the case of a pure sinusoidal $v_S$, which is composed of fundamental component only. It is proved that the estimates $\hat{i}_1^{\text{ref}}$ and $\hat{e}^{\text{ref}}$ converge towards their reference $i_1^{\text{ref}}$ and $e^{\text{ref}}$, respectively, in the steady state. However, in the presence of harmonic distortion in $v_S$, the reference estimations exhibit a bounded convergence error.

If the grid voltage $v_S$ has no harmonic distortion, the grid-side current reference can be computed in terms of $v_S$, and the desired active power to be supplied to the grid $P_{\text{ref}}$. However, in general conditions, $v_S$ has additional harmonic components, and therefore, its fundamental component may be used instead. Let $v_{s_1}$ be the steady-state fundamental component of the grid voltage, and $\phi_{s_1}$ be its square-phase component, that is, $\phi_{s_1}$ coincides with $v_{s_1}$ except for a phase shift difference of 90°. Notice that $v_{s_1}$ and $\phi_{s_1}$ form a basis for the set of periodic signals at the fundamental grid frequency. Therefore, the system references $i_1^{\text{ref}}$, $i_1^{\text{ref}}$, $v_C^{\text{ref}}$ and $e^{\text{ref}}$ in steady state can be constructed as linear combinations of $v_{s_1}$ and $\phi_{s_1}$.

To guarantee the fulfillment of the control objective, $i_S$ must be proportional to $v_{s_1}$, whose amplitude is fixed by the
desired active power to be supplied. Let \( g = P_{\text{ref}}/V_{\text{s, RMS}} \) be the apparent conductance of the system, and \( V_{\text{s, RMS}} \) the root-mean-square (RMS) value of the grid voltage. Then the grid-current reference can be proposed as (21).

\[
i_{\text{ref}} = g v_{x1}
\]  

(21)

Let \( \mathbf{x}^{\text{ref}} = [i_{\text{ref}}, i_{\phi_{\text{ref}}}, v_{\phi_{\text{ref}}}^\text{g}]^T \), then the remaining system references are imposed by the system dynamics (1)-(3), and can be expressed by (22).

\[
\mathbf{x}^{\text{ref}} = \begin{bmatrix} g (1 - \omega_s^2 L_C) \omega_s C & \omega_s C \\
g & 0 \\
1 & g \omega_s L_2 \\
\end{bmatrix} \begin{bmatrix} v_{x1} \\
\phi_{x1} \\
\end{bmatrix}
\]

(22)

Moreover, the admissible steady-state inverter voltage is given by (23).

\[
e^{\phi_{\text{ref}}} = (1 - \omega_s^2 L_C) v_{x1} + g \omega_s (L_1 + L_2 - \omega_s^2 L_1 L_2) \phi_{x1}
\]

(23)

Notice that \( v_{x1} \) and \( \phi_{x1} \) satisfy (24).

\[
\begin{bmatrix} \frac{dv_{x1}}{dt} \\
\frac{d\phi_{x1}}{dt} \\
\end{bmatrix} = \begin{bmatrix} 0 & \omega_s \\
-\omega_s & 0 \\
\end{bmatrix} \begin{bmatrix} v_{x1} \\
\phi_{x1} \\
\end{bmatrix}
\]

(24)

The time-derivative of the system state reference (22) is given by (25).

\[
\frac{dx^{\text{ref}}}{dt} = \begin{bmatrix} -\omega_s C & g \omega_s (1 - \omega_s^2 L_C) \\
n & g \omega_s \\
-\omega_s L_2 & \omega_s \\
\end{bmatrix} \begin{bmatrix} v_{x1} \\
\phi_{x1} \\
\end{bmatrix}
\]

(25)

Based on (22) and the above time-derivative, it can be verified that references in (22) satisfy the system dynamics (1)-(3), and thus, they are admissible trajectories.

The following linear, uniformly stable estimator for the fundamental component of the grid voltage \( \hat{v}_{x1} \) and its square-phase companion \( \hat{\phi}_{x1} \), is used based on the linear estimator presented in [24], [29], [30].

\[
\begin{bmatrix} \frac{dv_{x1}}{dt} \\
\frac{d\phi_{x1}}{dt} \\
\end{bmatrix} = \lambda (v_{x1} - \hat{v}_{x1}) + \omega_s \phi_{x1}
\]

(26)

\[
\frac{dv_{x1}}{dt} = -\omega_s \hat{v}_{x1}
\]

(27)

The estimation gain \( \lambda > 0 \) modifies the rate of the estimation. Estimates of \( \hat{i}_{\text{ref}}^{\text{ref}} \) and of \( \hat{e}^{\phi_{\text{ref}}} \) can be computed as follows:

\[
\hat{i}_{\text{ref}}^{\text{ref}} = g (1 - \omega_s^2 L_C) \hat{v}_{x1} + \omega_s C \phi_{x1}
\]

(28)

\[
\hat{e}^{\phi_{\text{ref}}} = (1 - \omega_s^2 L_C) \hat{v}_{x1} + g \omega_s (L_1 + L_2 - \omega_s^2 L_1 L_2) \phi_{x1}
\]

(29)

If reactive power injection is required, references (28) and (29) can be modified as follows. Let \( Q_{\text{ref}} \) be the desired reactive power references, and define \( h = Q_{\text{ref}}/V_{\text{s, RMS}}^2 \). The magnitude of the apparent admittance of the system is defined as \( \sqrt{g^2 + h^2} \). Then the required grid current reference in (21) must be modified to \( \hat{i}_{\text{ref}} = g v_{x1} + h \phi_{x1} \). Following an analogous procedure as the one used to obtain the references imposed by the system dynamics (22) and the admissible steady-state inverter voltage (23), the following references can be proposed.

\[
\hat{i}_{\text{ref}} = \hat{i}_{\text{ref}}^{\text{ref}} + h (1 - \omega_s^2 L_C) \phi_{x1}
\]

(30)

\[
\hat{e}^{\phi_{\text{ref}}} = \hat{e}^{\phi_{\text{ref}}} - h \omega_s (L_1 + L_2 - \omega_s^2 L_1 L_2) \hat{v}_{x1}
\]

(31)

Then, the expressions (30) and (31) are the estimated references for the inverter-side inductor current and the steady-state inverter voltage, respectively, when the reactive power reference \( Q_{\text{ref}} \) is not zero.

In fact, the estimator in (26) and (27) is linear. Its eigenvalues have negative real part, and it is bounded-input, bounded-output stable. Therefore, if the grid voltage is bounded and periodic, the estimations \( \hat{i}_{\text{ref}}^{\text{ref}} \) and \( \hat{e}^{\phi_{\text{ref}}} \) are also bounded and periodic.

Assuming that \( v_s \) has no harmonic contents, the estimations \( \hat{i}_{\text{ref}}^{\text{ref}} \) and \( \hat{e}^{\phi_{\text{ref}}} \) in (28) and (29) converge exponentially to \( i_{\text{ref}}^{\text{ref}} \), in the first scalar equation of (22), and \( e^{\phi_{\text{ref}}} \) in (23). However, if \( v_s \) has considerable harmonic contents, there is an steady-state error between estimated references and actual references. In the remaining part of this subsection, bounds are obtained for these errors. The usual grid voltage harmonics for single-phase electrical systems are at odd multiples of the fundamental frequency. Then the Fourier series representation of the grid voltage in steady-state is expressed as \( v_s = \sum_{n=1}^{N} V_n \sin((2n-1)\omega_s t) \). Notice that \( \hat{I}_{\phi, n}^{\text{ref}}(s) = G_1(s) \psi_n(s) \) and \( \hat{E}^{\phi_{\text{ref}}}(s) = G_2(s) \psi_n(s) \), where \( G_1(s) \) and \( G_2(s) \) are given in (32).

\[
\begin{align*}
G_1(s) &= \frac{\lambda (1 - \omega_s^2 L_C) s - \lambda \omega_s^2 C}{s^2 + \lambda s + \omega_s^2} \\
G_2(s) &= \frac{\lambda (1 - \omega_s^2 L_C) s - \lambda \omega_s \omega_s (L_1 + L_2 - \omega_s^2 L_1 L_2) C}{s^2 + \lambda s + \omega_s^2}
\end{align*}
\]

(32)

The magnitudes of these transfer functions at each consecutive multiple \( n \) of the harmonic frequency, i.e., \( M_1(n) = |G_1(j(2n-1)\omega_s)| \) and \( M_2(n) = |G_2(j(2n-1)\omega_s)| \) are given by (33), where \( n \in \mathbb{N}^+ \).

\[
\begin{align*}
M_1(n) &= \frac{\lambda \sqrt{g^2 (1 - \omega_s^2 L_C)^2}}{\sqrt{\lambda^2 (2n-1)^2 + 16n^2 (2n-1)^2 \omega_s^2}} \\
M_2(n) &= \frac{\lambda \sqrt{g^2 \omega_s^4 (L_1 + L_2 - \omega_s^2 L_1 L_2 L_2)^2}}{\sqrt{\lambda^2 (2n-1)^2 + 16n^2 (2n-1)^2 \omega_s^2}}
\end{align*}
\]

(33)

Therefore, the Fourier representation of the steady-state estimated references are given by (34).

\[
\begin{align*}
\hat{I}_{\phi, n}^{\text{ref}} &= \sum_{n=1}^{N} V_n \sin((2n-1)\omega_s t - \psi_{2n-1}) \\
\hat{E}^{\phi_{\text{ref}}} &= \sum_{n=1}^{N} V_n \sin((2n-1)\omega_s t - \theta_{2n-1})
\end{align*}
\]

(34)

where \( \psi_{2n-1} = \varphi_{2n-1} - \angle G_1(j(2n-1)\omega_s) \) and \( \theta_{2n-1} = \varphi_{2n-1} - \angle G_2(j(2n-1)\omega_s) \).
\[ \angle G_1(j(2n-1)\omega_3). \text{ Hence, the error of each estimated reference, } \Delta i_{i1}^\text{ref} = i_{i1}^\text{ref} - i_{i1}^\text{err} \text{ and } \Delta e^\text{ref} = e^\text{ref} - e^\text{err}, \text{ satisfy (35) and (36),} \]
\[ |\Delta i_{i1}^\text{ref}| \leq \frac{2\omega_3 C \sum_{n=2}^{\infty} V_{m,2n-1}}{n(n-1)} + \frac{2\omega_3 L_1 + L_2 - \omega_3 L_1 L_2 C}{4 \sum_{n=2}^{\infty} mn(n-1)} \]
\[ |\Delta e^\text{ref}| \leq \frac{2\omega_3 L_1 + L_2 - \omega_3 L_1 L_2 C}{4 \sum_{n=2}^{\infty} mn(n-1)} \]  
(35)

If the harmonic components of \( v_s \) are zero from a given \( n \) and above, as occurs in practical applications, or if the magnitude of the harmonics is of the order of \( 1/n \), errors are bounded since the sums in the inequalities (35) and (36) are convergent. The approximation error becomes zero whenever the grid voltage is a pure sinusoidal.

C. Harmonic Compensation Mechanism

The harmonic distortion of \( v_s \) propagates to the controller and estimator. A solution to compensate the harmonic distortion is to use a resonant controller structure following the principle of the internal model [26]. This includes the model of the perturbation to be compensated in the stable feedback path. Hence, asymptotic tracking is achieved. This control approach has been successfully used in several power electronics applications [24], [25].

Considering the estimated references \( \tilde{i}_{i1}^\text{ref} \) and \( \tilde{e}^\text{ref} \), the controller (20) can be expressed as follows:
\[ e = -k_i + ki_{i1}^\text{ref} + \tilde{e}^\text{ref} - ki_{i1} + k \left( i_{i1}^\text{ref} - \Delta i_{i1}^\text{ref} \right) + \left( e^\text{ref} - \Delta e^\text{ref} \right) - ki_{i1} + ki_{i1}^\text{ref} + e^\text{ref} + \Phi \]  
(37)

where \( \Phi = -k \Delta i_{i1} - \Delta e^\text{ref} \) is a periodic bounded perturbation whose Fourier series representation includes the odd harmonics in the grid voltage. Therefore, a resonant control structure may be included at the feedback path to compensate \( \Phi \), which consists of a set of resonant filters tuned at odd multiples of the fundamental grid frequency, i.e., frequencies defined by \( (2n-1)\omega_s, \forall n \in \{1, 2, \cdots, N\} \), where \( N \) can be fixed as the number of the first more significant harmonics in the grid voltage. By considering the resonant structure (37), the following expression for the controller is obtained:
\[ e = -k_i + ki_{i1}^\text{ref} + e^\text{ref} + \Phi - \sum_{n=1}^{N} \gamma_{2n-1} \zeta_{2n-1} \]  
(38)

The last term in (38) can be implemented in the form of a transfer function as follows:
\[ \sum_{n=1}^{N} \gamma_{2n-1} \zeta_{2n-1} (s) = \sum_{n=1}^{N} \frac{\gamma_{2n-1} s}{s^2 + (2n-1)^2 \omega_s^2} \]  
(39)

where \( \gamma_{2n-1}, \forall n \in \{1, 2, \cdots, N\} \) is a constant gain.

A block diagram of the overall controller is presented in Fig. 3. The inputs of the reference estimation block are \( v_s \) and \( P_{\text{ref}} \), and the generated outputs are the \( \tilde{e}^\text{ref} \) and \( \tilde{i}_{i1}^\text{ref} \) described by (28) and (29), respectively. The parameters \( g, a_1 = 1 - \omega_3 L_1, a_2 = 1 - \omega_3 L_2, a_3 = \omega_3 C \) and \( a_4 = \omega_3 (L_1 + L_2 - \omega_3^2 L_1 L_2 C) \) are involved in the calculation of the estimates.

The inputs of the controller are the inverter-side current, the inverter-side current reference and the estimations coming from the reference estimation block. The controller includes two actions namely proportional and resonant control actions over the inverter-side current error. To avoid issues caused by the infinite gain at resonant peaks, it is proposed to implement the bank of resonant filters (39) as follows:
\[ \sum_{n=1}^{N} \gamma_{2n-1} \frac{(2n-1)\omega_s/Q_{2n-1}}{s^2 + (2n-1)^2 \omega_s^2} \]  
(40)

where the resonant filters have been replaced by selective band-pass filters. These filters have an adjustable limited gain \( \gamma_{2n-1} \), and can be made as selective as required through the quality factor \( Q_{2n-1} \).

IV. EXPERIMENTAL VALIDATION

The controller presented in Fig. 3 is experimentally tested to evaluate its performance in a 1.0 kVA HB prototype built upon insulated gate bipolar transistor (IGBT) modules and with unipolar sinusoidal pulse width modulation (SPWM). The controller is not limited to this particular topology, and can be modified to control other single-phase topologies. The inverter is connected to the local grid, with an RMS voltage of 122 V, \( \omega_s = 120\pi \text{ rad/s} \), through an LCL filter. The power is supplied by a programmable DC voltage source. The experimental setup is shown in Fig. 4. It includes current and voltage sensors, a signal conditioning stage, a power stage and a dSPACE ACE1103 control board where the controller is implemented. It is worth mentioning that the proposed controller can also be implemented in a DSP-based control board. This is possible because the control strategy involves common math operations and second-order transfer functions (for the reference estimation block and for each harmonic to compensate).
The DC-link voltage is fixed to $V_{DC} = 240$ V, the switching frequency is set to $f_s = 8$ kHz, and the power reference is set to $P_{ref} = 700$ W, which is the active power to be supplied to the grid. The LCL filter parameters are $L_1 = 1$ mH, $L_2 = 552$ μH, $C = 8$ μF. The sampling period is 50 μs. Other system and controller parameters are $V_{s,ref} = 127$ V, $k = 6.5$, $\lambda = 250$, $a_1 = 0.9988$, $a_2 = 0.9994$, $a_s = 3.016 \times 10^{-3}$, $a_h = 0.5848$, $y_1 = 96$, $Q_1 = 93$, $y_2 = 93$, $Q_2 = 94$, $y_3 = 99.89$, $Q_5 = 92.37$, $y_7 = 71$, $Q_7 = 92$, $y_8 = 99$, $Q_8 = 88$, $y_9 = 9.54$, $Q_9 = 89$, $y_4 = 21$, $Q_13 = 61$, $y_5 = 65$, $Q_17 = 77$.

The above parameters follow the design rules previously presented and reported in [2], [22], [23]. However, for the sake of completeness, the tuning process is described in more detail as follows. The base capacitance and base inductance are given by $C_b = P_{ref}/(\omega_s V_{s,ref}^2)$ and $L_b = V_{s,ref}^2/(\omega_s P_{ref}) = 61.12$ mH, respectively. The maximum inverter-side current ripple is given by $\Delta i_{l,n} = V_{DC}/(8 L_1 f_s) = 3.75$ A. The sum $L_1 + L_2$ equals 1.552 mH, which is lower than the recommended 10% of $L_b$, i.e., it is lower than 6.11 mH. The filter capacitance $C = 8$ μF does not exceed the recommended 15% of $C_b$, i.e., it is lower than 17.26 μF. This value limits the reactive power demanded by the LCL filter capacitor. The LCL resonance frequency $f_{res} = \sqrt{(L_1 + L_2)/(L_1 L_2 C)}/2\pi$ is located at 2.984 kHz, which lies between ten times the grid frequency $f_s = 60$ Hz and half the effective switching frequency at the inverter output. This is a common practice to avoid resonance issues in LCL-filter-based converter applications [31], [32].

The experiments include tests where the harmonic compensation block of the controller shown in Fig. 3 is enabled and disabled with the purpose to demonstrate its benefits, respectively. In both cases, only $v_S$ and $i_1$ are measured. Two system responses are presented to evaluate the performance of the proposed controller, the steady-state response at $P_{ref} = 700$ W, and the transient response during step changes of the power reference between 350 W and 700 W.

As observed in Fig. 5(a), if no harmonic compensation is included, both $i_1$ and $i_S$ exhibit a slight harmonic distortion. In contrast, as observed in Fig. 5(b), if the harmonic compensation scheme is activated, both $i_1$ and $i_S$ achieve almost pure sinusoidal waveforms, which corroborates the benefit of the compensation scheme. These figures also show the effectiveness of the LCL filter in reducing the switching harmonics at the grid-side current. Moreover, Fig. 5(b) shows that $i_S^{ref}$ reaches an almost pure sinusoidal waveform with the required phase shift with respect to $v_{S,1}$. This is necessary to guarantee that $i_S$ is in phase with the fundamental component of $v_S$.

Figure 6 shows the steady-state responses of the inverter-side current error $\tilde{i}_1$, the estimated inverter-side current reference $\tilde{i}_1^{ref}$ and the inverter-side current $i_1$ at $P_{ref} = 700$ W. Figure 6(a) shows the responses without harmonic compensa-
tion, while Fig. 6(b) shows the responses with harmonic compensation. It can be observed that $\hat{i}_i^{\text{ref}}$ is appropriately generated in both cases with and without the compensation scheme. As observed in Fig. 6(b), if the compensation mechanism is activated, $i_i$ achieves an almost sinusoidal waveform, and the amplitude of $\hat{i}_i$ is reduced to zero.

Figure 7 depicts the steady-state response of $v_S$ and its fast Fourier transform (FFT) $v_{S,\text{FFT}}$ and $i_s$ and its FFT $i_{s,\text{FFT}}$ at $P_{\text{ref}}=700$ W. It can be observed that $v_S$ is composed of the fifth and seventh harmonic components besides the fundamental component, while $i_s$ contains mainly fundamental component.

Figure 8 shows the steady-state response at $P_{\text{ref}}=700$ W under the overall controller, i.e., including the harmonic compensation mechanism. The figure includes $e$, $\hat{e}_i^{\text{ref}}$, $i_i$, and $\hat{i}_i^{\text{ref}}$. Notice that in the closed-loop operation, both references $\hat{e}_i^{\text{ref}}$ and $\hat{i}_i^{\text{ref}}$ coincide with the actual injected voltage and the measured inverter-side current, respectively, except for a high-frequency switching component in the inverter-side current.

Figure 10 shows the transient responses with power step changes from 350 W to 700 W and back from 700 W to 350 W, with the overall controller. The figure shows $P_{\text{ref}}$, $\hat{i}_S^{\text{ref}}$, $i_S$ and the grid-side current error $i_S^* = i_S - i_S^{\text{ref}}$. Notice that the current transients are relatively short as compared to the fundamental period of the grid voltage. In fact, both transients are very smooth without appreciable overshoots in the current time waveforms.

V. CONCLUSION

This paper presents the design, analysis and experimental validation of a model-based controller for a power inverter connected to the grid through an LCL filter, which has special interest in applications of photovoltaic generation systems among others.
The controller provides active damping and assures a grid-side current in phase with the fundamental component of the grid voltage. Moreover, the controller adopts a reduced number of measurements. In fact, only the grid voltage and the inverter-side current are required for the controller implementation. The controller structure consists of a proportional term and a harmonic compensation scheme over the inverter-side current error. One of the contributions of this work is the estimation of the inverter-side current reference and a decoupling term required for the controller implementation. Analysis has also been presented to show that the error between these estimations and the references turns out to be bounded. Moreover, it is shown that this error could be compensated even in the general case of a grid voltage with harmonic distortion. Therefore, the overall controller proposed in this paper is able to supply power to the grid with a guaranteed clean sinusoidal current signal. Experimental results with a laboratory prototype inverter prove the benefits of the proposed controller. The results show that the controller is able to introduce active damping and achieves lower-than-standard harmonic distortion in the grid-side current.

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