Robust Design of Active Damping with Current Estimator for Single-Phase Grid-Tied Inverters

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Abstract—This paper presents an improved control scheme for operation of grid-tied inverters in order to attain IEEE 1547.2-2008 harmonic limit standard even with a wide range of grid parameters. This control structure includes an active damping with a new current estimator and a proportional resonant (PR) controller with harmonic compensation (HC). In order to guarantee the inverter operation with a wide range of grid parameters the damping and the controller gains are designed considering robustness characteristics. Moreover, it proposes a new estimator for the current filter capacitor for operation with a wide range filter resonance frequency. The grid synchronization mechanism uses the MSOGI-FLL technique. Experimental results are carried out and presented to validate the control structure design and the proposed current estimation.

Index Terms—Grid-tied inverters, Robustness, Active damping, Current estimator, Power Quality.

I. INTRODUCTION

In the recent years, the distributed generation has shown a significant growth, especially with photovoltaic (PV) systems and wind farms [1], [2]. This kind of generation has been applied to meet the increasing energy demand and to reduce the environmental impact of electric power generation based on fossil fuel [3], [4]. Distributed generation can be defined as the electric power generation within the distribution network or with the consumer [5]. This distributed generation systems are typically grid connected and can be a single-phase or three-phase system depending on the application installed power. Low power (<10kW) generation units are becoming more and more applied and they are usually connected to the grid with a single-phase voltage-source inverter (VSI) at low voltage (110V to 220V) [3].

The grid-tied voltage-source inverter (VSI) has been widely adopted in distributed generation systems in a large power range applications [6]. This inverters (VSI) are typically connected with a low-pass filter to the grid in order to limit current harmonics which are caused by pulse width modulation (PWM) [6], [7]. Moreover, the transformerless grid-tied inverter with higher order low-pass filters have become ordinary because of costs and volume considerations [8].

Distributed generation systems can be connected at different grid points, including different types of local loads and different grid characteristics, as strong or weak impedances. Moreover, the regulations for grid connection have become increasingly restrictive. The standard IEEE 1547.2-2008 [9] places limits of amplitude for each harmonic component as well as a limit for the total harmonic distortion (THD). In order to attain the harmonic requirements, the current control loop can use the stationary frame proportional resonant (PR) controller with selective harmonic compensation (HC) [10]–[13]. Furthermore, a state of art technique to deal with grid frequency oscillations is to use the Second Order General Integrator (SOGI) Phase Locked Loop (PLL) with a Frequency Locked Loop (FLL) [10], [14]–[19].

The improvement of system stability is addressed by several recent works with active damping approaches [20]–[22], even with sensorless schemes [23]–[25]. In [26], the PR parameters are designed using direct pole placement but active damping is not considered. In [27], the design of PR gains is determined by the transient performance. However, no active damping is used and grid parameters variations are not considered. In [28], robust design of active damping is proposed but it deals only with PI controllers in synchronous frame. The design of a stationary frame PR with harmonic compensation (HC) and active damping considering grid parameters variation is lacking in the literature and would be of interest for grid-tied inverters operating at a wide range of grid parameters. Moreover, several sensorless active damping methods have been proposed in the literature, but most of them call for complex methods for grid parameters estimation, such the resonance frequency can be found and attenuated. Other works call for observer techniques, as in [29]. In this case the robustness of the active damping with a Luenberger observer is analyzed at several resonance frequencies. The authors show that the observer reduces system robustness due to grid inductance variations, especially at high resonance frequencies, as in strong grids.

The main contribution of this paper is to design the stationary frame PR with harmonic compensation (HC) and active damping in order to meet a wide range of grid parameters. Additionally, it proposes an active damping implementation with a new current estimation algorithm with simple design and implementation. The estimation scheme can operate at either high (strong grid) or low (weak grid) resonances frequencies. Furthermore, a control strategy with a connection filter is proposed in order to attain the harmonic limits of the IEEE 1547.2-2008. At the same time, the inverter control strategy withstands for a wide range of grid parameters, such
as frequency, resistance and inductance.

This paper is arranged in six sections. The second II details the inverter control structure. Section III presents the proposed active damping with the new current estimation scheme. The controller parameters design is carried out in section IV. Section V presents the experimental results and section VI presents the final remarks.

II. INVERTER CONTROL STRUCTURE

The regulations for grid connection have become increasingly restrictive [30], as with IEEE 519-1992 [31] and IEEE 1547.2-2008 [9], which set a limit for each harmonic component, as well as a total harmonic distortion (THD) threshold. The IEEE 1547.2-2008 establishes the maximum amplitude of each current harmonic and the THD must be lower than 5%. These requirements include higher order harmonics, as the ones at inverter switching frequency. In order to attain to standards in whole range of harmonics this paper applies full bridge PWM inverter with a LCL-LC filter. The additional LC term to the LCL filter is included to attenuate harmonic components at the inverter switching frequency. The grid tied inverter is shown in Fig. 1 and the respective parameters are given in Table I.

![Voltage Source Inverter (VSI) and LCL-LC Filter](image)

Fig. 1. Single-phase LCL-LC grid-tied inverter circuit.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1</td>
<td>LCL-LC Inductance L1</td>
<td>400.7 µH</td>
</tr>
<tr>
<td>L2</td>
<td>LCL-LC Inductance L2</td>
<td>128.81 µH</td>
</tr>
<tr>
<td>C_f</td>
<td>LCL-LC Capacitor C_f</td>
<td>3.35 µF</td>
</tr>
<tr>
<td>C_r</td>
<td>LCL-LC Capacitor C_r</td>
<td>668.53 nF</td>
</tr>
<tr>
<td>L_r</td>
<td>LCL-LC Inductance L_r</td>
<td>94 µH</td>
</tr>
<tr>
<td>R_1</td>
<td>LCL-LC Series Resistance of L_1</td>
<td>40.47 mΩ</td>
</tr>
<tr>
<td>R_2</td>
<td>LCL-LC Series Resistance of L_2</td>
<td>14.9 mΩ</td>
</tr>
<tr>
<td>R_f</td>
<td>LCL-LC Series Resistance of C_f</td>
<td>16.01 mΩ</td>
</tr>
<tr>
<td>R_r</td>
<td>LCL-LC Series Resistance of C_r, L_r</td>
<td>12.86 mΩ</td>
</tr>
<tr>
<td>L_grid</td>
<td>Grid Inductance (weak/strong grid)</td>
<td>2.94 mH/1 µH</td>
</tr>
<tr>
<td>R_grid</td>
<td>Grid Series Resistance (weak/strong grid)</td>
<td>2.46/0.40 Ω</td>
</tr>
<tr>
<td>f_grid</td>
<td>Grid Frequency</td>
<td>60 Hz</td>
</tr>
<tr>
<td>V_grid</td>
<td>Grid Voltage</td>
<td>127 V rms</td>
</tr>
<tr>
<td>f_inv</td>
<td>Switching frequency</td>
<td>20.040 Hz</td>
</tr>
</tbody>
</table>

Equations (1-3) show the grid tied inverter state space model, including the LCL-LC filter and grid parameters.

\[
\dot{x} = Ax + Bu, \quad (1)
\]

with \( x = \begin{bmatrix} i_{L1} & i_{inv} & V_{ef} & V_{cr} & i_{Lr} & i_{grid} \end{bmatrix}^T \) and \( u = \begin{bmatrix} V_u & V_{grid} \end{bmatrix} \). The matrices \( A \) and \( B \) are given by

\[
A = \begin{bmatrix}
    c_1 & c_2 & -1/L_1 & 0 & e_2 & 0 \\
    c_3 & c_4 & 1/L_2 & 0 & -e_3 & e_5 \\
    c_6 & -c_6 & 0 & 0 & -c_6 & 0 \\
    0 & 0 & 0 & 0 & 1/C_r & 0 \\
    -c_7 & 1/L_r & -1/L_r & c_8 & 0 & 0 \\
    0 & c_9 & 0 & 0 & 0 & c_{10}
\end{bmatrix}, \quad (2)
\]

\[
B = \begin{bmatrix}
    1/L_1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & -1/L_{grid} & 0 \\
    0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0
\end{bmatrix}^T. \quad (3)
\]

In (2) and (3), \( c_1 = (-R_1 - R_f)/L_1, \( c_2 = R_f/L_1, \( c_3 = R_f/L_2, \( c_4 = (-R_2 - R_f - R_{load})/L_2, \( c_5 = R_{load}/L_2, \( c_6 = 1/C_f, \( c_7 = R_f/L_r, \( c_8 = (-R_f - R_r)/L_r, \( c_9 = R_{load}/L_{grid} \) and \( c_{10} = (-R_{load} - R_{grid})/L_{grid}. \)

The proposed control structure for the grid-tied inverter is presented in Fig. 2. In the sequence, the paper discusses each module in the diagram and also the robust design of key controller gains for a wide range of grid impedances.

![Single-Phase Inverter Control Diagram](image)

Fig. 2. Single-Phase Inverter Control Diagram.

A. Phase-Locked-Loop (MSOGI-FLL)

The control algorithm starts with the phase-locked-loop module. The PLL is an algorithm used to obtain the phase and frequency of a signal, in this case, the PCC voltage \( V_{PCC} \). This work applies a resonant phase-locked-loop (18), (19), which provides two outputs with the same magnitude but displaced each other by 90°, i.e., the first \( (V_α) \) with the same phase as the input and the second \( (V_β) \) with a +90° phase. This PLL scheme is used since the current controller runs in the \( αβ \) frame.

The PLL block in Fig. 2 is based on the Modified Second Order General Integrator Phase Locked Loop (MSOGI-PLL), see (10), (14)-(17). The SOGI-PLL is reported to have good performance under grid voltage abnormalities like sag, swell, harmonics, unbalance, frequency variations, except DC offset.
[18]. Therefore, the modified SOGI-FLL is implemented to estimate the DC offset voltage. Thereby, an improved stability to the frequency estimation (FLL) is obtained as an input offset can lead to frequency oscillations or instability. In a real-world application, this input offset may occur due to measurement but usually comes from an instrumentation error. Besides that, the DC estimation guarantees that the current reference don’t have a DC component which could lead to distribution transformer saturation.

The MSOGI-FLL equations are given by:

\[
V_{\alpha}(s)/V_{PCC}(s) = K_c \omega s^2/\Delta, \tag{4}
\]

\[
V_{\beta}(s)/V_{PCC}(s) = K_c \omega s/\Delta, \tag{5}
\]

\[
V_{offset}(s)/V_{PCC}(s) = K_{os} \omega (s^2 + \omega^2)/\Delta, \tag{6}
\]

\[
\omega = \omega_0 + \omega_{in}/s. \tag{7}
\]

where \(\Delta = s^3 + (K_c + K_{os})\omega s^2 + \omega^2 s + K_{os} \omega^3\) and \(\omega_{in} = \omega K_{ic}gamma(V_\alpha + V_{offset} - V_{PCC})V_{\beta}/(V_\alpha^2 + V_{\beta}^2)\). In (4)-(7) the following gains are applied: \(K_c = 1, K_{os} = 0.27\) and \(gamma = 46\).

### B. Reference Generator

The PLL feeds the Reference Generator in Fig. 2 with \(V_\alpha\) and \(V_\beta\), which are synchronized with the electrical grid. In this stage, the inverter output current reference \(i_{\alpha}^*\) is generated using also external inputs \(P^*\) and \(Q^*\), the active and reactive power references, and \(P_{dc}\), the power variation due to the DC link voltage variation. These variables are used to generate the current reference with appropriate phase, frequency and amplitude. The reference generator algorithm proposed here is based in the three-phase case [32] and adapted for the single-phase case. Thus, the current reference \(i_{\alpha}^*\) is calculated as:

\[
i_{\alpha}^* = \frac{V_\alpha(P^* - P_{dc}) - V_\beta Q^*}{V_\alpha^2 + V_{\beta}^2}. \tag{8}
\]

Equation (8) still depends on \(P_{dc}\). The DC power \(P_{dc}\) can be obtained from the voltage deviation in the DC bus. The DC Controller block in Fig. 2 shows that \(P_{dc}\) is obtained from \(V_{DC}^*\) and \(V_{DC}\). The DC Controller block is composed essentially by a PI controller and a low pass filter. The input is the error for the DC bus reference voltage \(e = V_{DC}^* - V_{DC}\), where \(V_{DC}\) is the DC bus voltage measurement passing by a digital low pass filter. The low pass filter is tunned with cutoff frequency lower than the grid frequency to reduce the voltage ripple in the DC bus measurement.

### C. Current Controller

The PR+HC block in Fig. 2 is the current controller and it applies a proportional resonant (PR) controller [10]–[12] over the current error \(i_{\alpha} - i\), where \(i\) is the measured inverter filter output current. The ideal \(G_{PRi}(s)\) and the damped \(G_{PRd}(s)\) PR transfer function [13] are given by:

\[
G_{PRi}(s) = K_P + \frac{2K_{ic}s}{s^2 + \omega^2}. \tag{9}
\]

\[
G_{PRd}(s) \approx K_P + \frac{2K_{ic}\omega s}{s^2 + 2\omega_{ic}s + \omega^2}. \tag{10}
\]

Most works applies the damped PR, since it allows to adjust \(\omega_c\) to reduce the controller sensitivity to small frequency oscillations. On the other hand, it presents lower peak gain but still high enough to smoothly lead the steady-state error to zero. Since the MSOGI-FLL allows fast grid frequency estimation, in this paper the ideal PR controller with grid frequency feedback is adopted. This scheme allows frequency oscillations while having a high gain to obtain fast convergence of the steady state error to zero. Furthermore, resonant controllers at harmonic frequencies \((h\omega = 3\omega, 5\omega, 7\omega, ...)\) are included for selective harmonic compensation:

\[
G_h(s) = \sum_{h=3,5,7...} \frac{2K_{ih}s}{s^2 + (h\omega)^2}. \tag{11}
\]

### III. ACTIVE DAMPING WITH CURRENT ESTIMATION

Active damping is a control technique used for damping the resonance in the LCL filter. Thus, the active damping improves the system stability [20]–[22] and allows to operate within a range of grid parameters with a proper controller gains design. The key advantage over passive damping is that no power losses are introduced. Passive and active damping implementation can be seen in Fig. 3 (a) and (b) respectively.

The virtual resistor method is a known simple method of active damping but usually requires a current measurement. This measurement requires the design of an anti-aliasing filter with cutoff frequency higher than the highest resonance frequency. This filter design can proven to be difficult with strong grids and high resonance frequencies. Others works have proposed to achieve the same LCL filter resonance damping without a current measurement [23]–[25]. However, most of them resort to complex methods for grid parameters estimation or rely on high order observers. To overcome this issue, in this paper a simpler active damping implementation by the LCL capacitor current estimation \(i_c\) is proposed.

In Fig. 4 is shown the filter model proposed for current estimation. This model can be applied with both the LCL and LCL-LC filter. An equivalent capacitance of the LCL-LC filter must be calculated by adding the capacitors \(C_r\) and \(C_f\) [33]. The virtual resistor \(K_{ic}\) in this model introduces a damping factor such that it is possible to use the estimator with high resonance frequencies as in strong grids.

Choosing \(K_{ic}\) greater than \(L_1\) series resistance introduces an error to the model but it allows operation in a range of electrical grid parameters. More details of the advantages and drawbacks of using \(K_{ic}\) will be shown using computational analysis in Section IV.
Thus, as $K_{ic}$ increases the stability margin is also increased. On the other hand, when $K_{ic}$ is chosen to be a different value from filter series resistance a small error is introduced in the current estimation. This effect occurs especially for weak grids with low resonance frequency, as shown in Fig. 5. If $K_{ic} = 12$ is chosen a good stability margin is obtained for strong grids despite introducing a small estimation error with weak grids, see Fig. 5.

![Fig. 5. $K_{ic}$ effect on $i_c$ bode diagram.](image)

![Fig. 6. $K_{ic}$ effect on $G_{ic}(z)$ poles.](image)

The performance of the proposed current estimator is first verified using computational simulations. The main objectives are analyzed: the ability of ($\hat{i}_c$) to follows the real capacitor current ($i_c$); and, the estimator based active damping improves system stability. The grid tied inverter parameters used in simulations are presented in Table I and the grid parameters are $R_{grid} = 2, 46\Omega$ and $L_{grid} = 2, 94mH$, from a typical weak grid. Furthermore, no local load is connected in the PCC and the controller parameters are $K_p = 2, 5$, $K_i = 3000$, $K_m = 2$. In this case the active damping was activated only after 81 ms. From 81 ms to 120 ms $K_{ic} = 12$ is set and after 120 ms $K_{ic}$ is changed to 0.

Thus, as $K_{ic}$ increases the stability margin is also increased.

![Fig. 5. $K_{ic}$ effect on $i_c$ bode diagram.](image)

![Fig. 6. $K_{ic}$ effect on $G_{ic}(z)$ poles.](image)

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Fig. 7 shows the voltage at the connection point ($V_{PCC}$), the inverter current ($i_{inv}$), the estimated current ($\hat{i}_c$) and the measured capacitor current after the anti-aliasing filter. In Fig. 7(b) it can be verified that the capacitor current is increasing. High frequency oscillations increases at PCC voltage up to 80 ms. After 81 ms, the active damping reduces $i_c$ oscillations such the system that was becoming unstable is stabilize. Fig. 7 (c) presents the detailed view when the current estimation (dark) is triggered. The transient of $\hat{i}_c$ is fast even when started at the peak capacitor current and with current oscillations due to filter resonance. From Fig. 7(c) and (d) it is clear that for lower values of $K_{ic}$ the high frequency behavior of $\hat{i}_c$ will be closer to $i_c$. However, it can be verified the active damping works well with the proposed capacitor current estimation for both values of $K_{ic} = \{12, 0\}$ and performs better with higher values of $K_{ic}$. In this case, $\hat{i}_c$ follows the low frequency behavior of $i_c$.

![Fig. 7. Simulation of the capacitor current estimation ($i_c$) for a weak grid.](image)

The current estimation is also evaluated with strong grid parameters. In this case, the active damping is activated at the inverter grid connection. The current estimation is performed with $K_{ic} = 12$ up to 120 ms, when $K_{ic}$ is changed to zero. Fig. 8 shows the estimated current and it can be noted that with $K_{ic} = 12$ the estimator has good performance. When $K_{ic}$ is switched to zero the estimation diverges. Furthermore, from the bode diagram analysis of Fig. 5, it can be noted that for high resonant frequencies the error carried by $K_{ic} = (0, 12)$ is negligible.

![Fig. 8. Simulation of the capacitor current estimation ($i_c$) for a strong grid.](image)

Thereby, the choice $K_{ic} = 12$ introduces a small error in the current estimation for weak grids but does not significantly interfere in the active damping implementation. On the other hand, this choice of $K_{ic}$ allows the capacitor current estimation and active damping implementation with strong grid parameters as with high resonance frequencies. Gaafar et al [29] presents a robustness analysis of the active damping with a Luenberger observer for the capacitor current observer. The author concludes that the observer reduces system robustness due to grid inductance variations, especially at high resonance frequencies in strong grids. Thus, an advantage of the proposed current estimation scheme is that it does not depend on the electrical grid inductance and with a proper $K_{ic}$ choice allows operation with high resonance frequencies in both weak and strong grids.

IV. CONTROLLER PARAMETERS DESIGN

This section details the proposed methodology for the design of the controller parameters. The control structure employs the PR controller with harmonic compensation (HC) and active damping.

The proper adjust of the proportional gain ($K_p$) can improve system dynamics, especially connection transients. Although high controller gains can lead the system to instability, the appropriate design of active damping gain ($K_m$) can allow higher controller gains or a greater range of grid parameters.

In this paper the focus is the design of the controllers gains such that it is possible to operate with a range of grid parameters. The term $L_{2r}$ is the sum of the converter side inductance ($L_2$) and the grid side inductance ($L_{grid}$). The design approach for the PR + HC controller with active damping is carried out delimiting a range of grid parameters such it is possible to obtain the controllers gains with a proper stability margin.

The block diagram in Fig. 9 with a LCL filter and no local load is considered. The LCL filter is chosen to simplify the analysis, since both filters have similar behavior for frequencies outside the range close of switching frequency.

From the block diagram in Fig. 9 the open and closed loop system transfer function are found in equations (21) and (22),
The coefficients $a_i$, $b_i$ and $b'_i$ are shown in equations (23) and (24). Applying the Routh-Hurwitz stability criterion the gain conditions for stability are found as shown in (25), (26) and (27). It is important to be noted that the stability limits found for the PR controller are similar to the stability limits found for the PI controller in synchronous reference, see [28].

\[
K_p > 0, K_m > 0, K_i + K_i < 0, \quad (25)
\]

\[
K_p < \frac{K_m(L_1 + L_{2r})}{L_1}, \quad (26)
\]

\[
K_i + K_i < \frac{(K_pK_m(L_1 + L_{2r}) - K_p^2L_1)}{2CK_m^2L_{2r}}. \quad (27)
\]

The active damping gain ($K_m$) design takes in account that the virtual resistance method causes the reference voltage $V^*$ to be decremented by $K_mV^*$. Thus, limiting $K_m$ to 10% of the nominal voltage [28] results in the upper bound of $K_m = 8$.

The $K_m$ variation from zero to its upper bound is shown in the open loop Bode diagram in Fig. 10. The open loop bode diagram considers that $K_p = 1$ and $K_i = 0$ are used. The inductance $L_{2r}$ is set equal to $L_2$, resulting in a strong condition, since this is the most critical case of filter resonance. The frequency response shows how $K_m$ affects the damping of the filter resonance. From the $K_m$ curve it is possible to choose a gain that offers a good attenuation with a margin to choose $K_p$. It can be noted that the choice of $K_p$ cause an offset in the frequency response (see Fig. 12), i.e., as $K_p$ increases in Fig. 13 the system poles are moved towards outside of the unit circle. In addition, high values of $K_m$ can lead to a THD increase despite maintaining system stability, since $K_mV^*$ is subtracted from $V^*$. Thereby, the chosen curve $K_m = 2$ presents a margin from the upper bound with an attenuation of -8.43 dB at the filter resonance frequency (8.06 kHz) to increase system stability.

Fig. 9. System Model for controller gains design.

Fig. 10. Frequency response of $G_{PRAMc}(s)$ for $K_m$ variation.

Fig. 11. Root locus of $G_{PRAMc}(z)$ with $K_m$ variation.

Fig. 12. Frequency response of $G_{PRAMc}(s)$ for $K_p$ variation.
Thus, it results replacing the parameters $K_p$ can be noted that as $K_p$ increase, the system poles are moved towards outside of the unit circle. The system is unstable for $K_p > 2.6$.

Fig. 14 shows the root locus of closed loop system with $K_p = 2.5$ and with grid inductance variation. This analysis confirms that increasing the grid inductance it moves the system poles towards inside of the unit circle, that allows a greater proportional gain ($K_p$).

The upper bound for $K_i$ and $K_{ih}$ can be found from (27) replacing the parameters $K_p = 2.5$, $K_m = 2$, $L_{grid} = 0$ mH. Thus, it results

$$K_i + K_{ih} < 34582, 6.$$

This relation shows that the upper bound for the resonant gains is not dependent with the harmonic order ($h$) and the sum of all the resonant gains should be lower than the calculated limit. However, this limit is quite large to allow the choice of a $K_i$ which reduces the steady state error convergence.

The effect of integral gain variation is presented in the frequency response shown in Fig. 15. From the frequency response, one can observe that a higher resonant gain leads to a smaller frequency selectivity. By analyzing the curves, it is convenient to choose mid-term gains in order to reduce the steady state error and still maintain the frequency selectivity. Thereby, $K_i = 1000$ results in fast enough response and with a good selectivity. For $K_{ih}$, an even smaller gain can be chosen, since the amplitude of the harmonic components is usually much smaller than the fundamental component.

![Root locus of $G_{PRAmC}(z)$ for $K_m = 2$ and $K_p$ variation.](image1)

![Root locus of $G_{PRAmC}(z)$ for $L_{grid}$ variation.](image2)

To design the proportional gain, equation (26) is applied providing an upper bound depending on the chosen $K_m$ and grid parameters.

$$K_p < 6.43\left|\frac{L_{grid}}{L_m}\right| = 0 \text{ mH}$$
$$K_p < 0.137\left|\frac{L_{grid}}{L_m}\right| = 0.70 \text{ mH}$$
$$K_p < 0.137\left|\frac{L_{grid}}{L_m}\right| = 2.94 \text{ mH}$$

Fig. 12 presents the effect of $K_p$ variation on the frequency response using $K_m = 2$. Fig. 13 shows the root locus for $K_p$ variation in the discretized closed-loop transfer function ($G_{PRAmC}(z)$), with $K_m = 2$, $K_i = 3000$ and $K_{ih} = 1000$. It can be noted that as $K_p$ increase, the system poles are moved towards outside of the unit circle. The system is unstable for $K_p$ greater than 2.6.

From the designed parameters $K_p = 2.5$, $K_m = 2$, $K_i = 3000$ and $K_{ih} = 300$, Fig. 16 presents the response closed-loop system tracking a sinusoidal reference ($i^*$) for different values of $L_{grid}$.

From the system response in Fig. 16 it can be seen that with small grid inductances (strong grids) the system quickly follows the reference. However, for large grid inductances (weak grids) the system operates slower, but still maintains stability, as occurs with $L_{grid} = 15$ mH. The reason for the slower response in weak grids is due to the proportional gain design ($K_p$). The ($K_p$) upper bound is found using strong grid parameters. It results in slower weak grid dynamics in behalf of system stability for a larger range of parameters. Recalculating the $K_p$ threshold for a weak grid results $K_p = 17.32$ and $K_p = 77.5$ for 2.94 mH and 15 mH, respectively.

The proposed system is also evaluated including harmonic compensation for the the 3rd, 5th, 7th and 9th harmonics,
with gains 600, 500, 400 and 300, respectively. The bode diagram with the designed gains and different grid parameters are shown in Fig. 17.

![Bode Diagram](image1.png)

**Fig. 17.** Frequency response of $G_{PRAM,c}(z)$ with the designed gains.

V. EXPERIMENTAL ANALYSIS

The experimental analysis presents the evaluation of the proposed active damping with the capacitor current estimator and the control structure, with the designed controller gains in Section IV. As suggested in [34], the grid connection and disconnection are driven at the zero voltage crossing. The experimental setup is presented in Fig. 18, where the power circuit (Fig. 1) is shown at the top and the DSP board at the bottom. The signal conditioning and gate driver board is plugged under the DSP board.

![Experimental Setup](image2.png)

**Fig. 18.** Experimental setup: 1 kW at 127 V grid-tied inverter.

Fig. 19 (a) shows the PCC voltage $V_{PCC}$ (in black) and the output current $i_{inv}$ (in gray), for the grid connected inverter at normal operation. The grid connection transitory is shown in Fig. 19 (b), where a step active power reference ($P^*$) is applied at the zero voltage crossing (1710ms) after the inverter is connected to grid (1702ms). The grid disconnection is shown in Fig. 19 (c), where the active power reference ($P^*$) is set to zero by half-cycle and the disconnection is accomplished at the zero voltage crossing (3726ms).

The spectral analysis of inverter current ($i_{inv}$) is presented in Fig. 20 (a), from DC up to 1000th harmonic. The total harmonic distortion (THD) is 2.18 %, that is lower than the 5 % limit of IEEE 1547.2-2008. It is important to note that the THD of the electrical grid voltage before the connection was 1.81 %. Fig. 20 (b) presents the spectral analysis from 2nd to 50th harmonic, which are all within the limits of the IEEE 1547.2-2008 standard. Fig. 20 (c) shows that harmonic components close to switching frequency are also comply the standard due the inclusion of LCL-LC filter.

![Spectral Analysis](image3.png)

**Fig. 19.** Grid-tied inverter operation: (a) Steady state, (b) grid connection and (c) disconnection.

![Spectral Analysis](image4.png)

**Fig. 20.** Inverter current spectral analysis: (a) DC up to 1000th harmonic, (b) 2nd up to 50th harmonic, (c) components at the switching frequency.

The grid connection is also carried out with low PCC voltage to evaluate the proposed active damping. Besides that, a resistive load of 17.25 $\Omega$ is connected at the PCC. The proposed active damping only is enabled after 200ms. Fig. 21
shows the PCC voltage $V_{PCC}$ (in grey) and the inverter current $i_{inv}$ (in black). The active damping is implemented using the measured current (Fig. 21 (a)) and the estimated current (Fig. 21 (b)). The experimental evaluation clearly shows that active damping increases system stability and allow a wider range of grid parameters. Moreover, the proposed capacitor current estimation can be applied to eliminate the respective current sensor.

![Fig. 21. Active damping with (a) measured current and (b) estimated current.](image)

### VI. CONCLUSION

This paper presents a robust design of active damping with current estimation to guarantees the good performance of single-phase grid-tied inverters with a range of grid parameters. The proposed control structure is designed to obtain suitable controller gains for strong or weak grids. In this way, higher controller gains can be used since the proposed active damping approach guarantees the stability of the system. Furthermore, a filter capacitor current estimation is proposed with a simple implementation, which allows operation with low and high filter resonance frequencies. The experimental evaluation confirms the appropriate control design since both grid connection and disconnection result in stable operation. Moreover, the THD=2.1% of output current complies with IEEE 1547.2-2008 standard and the active damping with the proposed current estimation stabilizes the grid-tied inverter operation for a wide range of grid parameters.

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