Design of LCL and LLCL filters for single-phase grid connected converters

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Abstract: Compared with the traditional L filters, the high order LCL and LLCL filters for grid integration of voltage source converters offer better switching harmonic attenuation even with smaller passive elements. However, the inherent resonance of the high order filters complicates the control of the converter and also the design of filter parameters, especially at presence of digital delays and grid impedance variations. This study proposes a simple design procedure for the LCL and the LLCL filters of a delay-based stabilised converter that satisfies the predefined constraints on the converter current ripple, grid current harmonics and the reactive power of the capacitor and at the same time, ensures adequate stability of the simple single loop current control against the resonance. Moreover, the effect of grid impedance on the filter performance and controller stability is taken into account, thus the proposed method is robust against a wide range of grid impedance variations. Experimental results for a 3 kW test rig under steady state and transient conditions confirm the effectiveness of the proposed filter design algorithm.

1 Introduction

The rapidly increasing energy demand and growing environmental concerns cause further attention to renewable energy sources (RESs). A current controlled voltage source converter (VSC) is an industrially accepted solution to interface the RES to power grid [1]. The quality of the current injected by VSCs to the utility grid is a main concern today. In order to prevent the harmonics and switching noises of the VSCs from penetrating the utility grid, some kind of passive low-pass filters, such as L, LCL, LLCL or other higher order filters are used in the output stage of the converter. The harmonic attenuation capability of LCL and LLCL filters are significantly higher than the traditional L filters, even with smaller passive elements, which offers lower volume, cost and losses [2]. Although, in spite of the prominent advantages of high order filters, the inherent resonance nature of them challenges the stability of the current control loop that may even lead to instability under certain circumstances. To deal with the resonance problem and stabilise the closed loop current control system, several methods are presented in recent studies [3–17]. Passive damping, using an additional resistor in the filter structure, is the simplest and most common method, which suffers from major drawbacks, such as power losses, reduced attenuation at switching frequencies and deteriorated performance when connected to a weak grid [3–5].

To ensure stability against the resonance and at the same time avoid additional losses, the active damping methods are proposed. In these methods, the resonance damping is achieved by modifying the control algorithm, which requires the addition of either some kind of digital filters to the control loop or extra sensors to the circuit [6–17]. Filter based methods include two distinct controllers in the current control loop. While the main controller regulates the low-frequency dynamics, the added one, which can be seen as a filter, attenuates the high-frequency resonances [6–9]. The grid impedance variations can weaken the damping effect and turn the stabilised system to instability. Active damping with extra control variables, such as the capacitor voltage [10, 11] or the capacitor current [12–17] are proposed. These techniques offer adequate damping in a selective way, at the cost of added circuit cost and control complexity, especially with considering digital implementation delays [16, 17]. An interesting technique to yield stability over a certain range of resonance to sampling frequency ratios, in a simple single-loop current control system is the delay-based stabilisation [15, 18–21]. In contrast to active damping solutions proposed yet, no digital filter is inserted in the current control loop and no extra variable, except the controlled current is sensed or controlled. In these approaches, different factors that influence the stability, considering the effect of the digital control delay are identified first. Then, the sampling and/or the resonance frequency are carefully selected such that adequate stability margins are obtained for the single-loop control with a certain delay presented in the loop. For instance, considering the 1.5 \( T_p \) delay, the grid current control algorithm, the stability of these methods is highly affected by grid impedance variations. The change of grid impedance can push the resonance frequency outside the stable region.

Regardless of the adopted stabilisation technique, designing the parameters of the high-order filters (LCL and LLCL) is also an important as well as complicated issue. In most of the previous works, the parameter design procedure is a trial and error method with the goal of size and cost reduction [22–27]. In this way, the LCL filter parameters are determined so that the converter-side inductor limits the converter current ripple, the capacitor value is simply chosen based on the reactive power limits and the grid-side inductor is decided to meet the grid current harmonic attenuation in accordance to IEEE-1547.2-2008 standard [28]. A similar approach is also proposed for higher order filters, such as the LLCL [24], the LTCL [25] and the LCL-TC filters [26].

In combination with the delay-based stabilisation methods, it is evidently difficult to find the best compromise between various, sometimes conflicting, requirements of output current ripple, harmonic attenuation, reactive power limits, size of components and the stability margins. Consequently, this paper proposes a simple design procedure for the LCL and the LLCL filters of a
delay-based stabilised converter that ensures stability of the closed loop system, and at the same time meets the reactive power, current ripple and harmonic limits requirements. The effect of grid impedance variations on the system stability is also considered, which ensures robust operation even with a weak grid.

The rest of the paper is organised as follows: Section 2 formulates the Bode plots of LCL and LLCL filters in both low and high frequency ranges for the LCL filter and the LLCL filters is presented in Section 5. The experimental results are given in Section 6 and Section 7 concludes the paper.

### 2 Harmonics in the output of the single phase VSC

The power stage of the grid-connected single phase VSC is shown in Fig. 1. The switching harmonics of the VSC output voltage \( v \) appear in the grid current. As mentioned before, to adequately suppress the switching harmonics in the current, the VSC is connected to the grid through a high-order low-pass LCL or LLCL filter.

The most common modulation method for the single phase VSC is the unipolar asymmetrical (double-update) regular sampled pulse-width modulation (PWM), for which the output voltage of the VSC is expressed as \[29\] (see (1))

where \( q = 2m + (2n − 1)M_t \) and \( V_{dc} \) is the dc link voltage, \( M_L \) and \( M_r \) are the amplitude and the frequency modulation indices, respectively, \( \omega_0 \) and \( \omega_0 \) are the carrier and the fundamental frequencies, respectively, \( (M_L = \omega_0/\omega_0) \) and \( J_1(x) \) is the integral of Bessel function, which is expressed as \[J_1(x) = \int_0^x \cos(n\pi - \sin(\pi x)) \, dx\].

In the asymmetrical unipolar PWM, the sampling frequency is \( \omega_0 = 2\omega_c \) and obviously, the first sideband is located at \( 2\omega_c \).

From (1), the amplitude of the high frequency switching components (i.e. sideband harmonics) can be derived as (2). The two first sidebands are plotted in Fig. 2.

\[Y_{sh,n}(m,n,\omega_c) = \frac{4V_{dc}}{\pi} \sum_{m=1}^{\infty} \frac{1}{q} J_{2n-1}\left(\frac{\pi}{2} M_r\right) \cos((m + n - 1)\pi) \times \cos((2m\omega_c + (2n - 1)\omega_0)t)\]  

(2)

The lower the frequency of the sideband, the greater the harmonic amplitude; that is, the most dominant harmonics, as demonstrated in Fig. 2, are located around the first and the second sidebands.

### 3 LCL and LLCL filters characteristics

Both LCL and LLCL filters are shown in Fig. 1. The transfer function of the converter voltage \( v \) to the converter and grid currents \( (i_1 \) and \( i_2) \) for the LLCL filter are formulated as \[24\]

\[Y_{11}(s) = i_1(s) = \frac{C_f(L_1 + L_2) s^3 + 1}{C_f L_1 L_2 s^2 + 1} \]  

\[Y_{22}(s) = i_2(s) = \frac{C_f(L_1 + L_2) s^3 + 1}{C_f L_1 L_2 s^2 + 1} \]  

(3)

where \( L_2 = L_2 + L_4 \) and \( L_2 \) represents the equivalent inductance of the grid. Apparent from (3), the resonance frequency of the LLCL filter is \( \omega_0 = (L_1 + L_2)/(2\pi C_f) \) and the filter attenuation for \( i_1 \) and \( i_2 \) currents, respectively. Fig. 3 shows the Bode plots of \( Y_{21} \) for the LCL and LLCL filters.

As it can be seen in Fig. 3 and presented in (4), the LCL filter offers a high attenuation at high frequencies, where the gain rolls-off at -60 dB/decade, while this rate is limited to -20 dB/decade for the LLCL case. However, the LLCL filter has a pair of complex-conjugate zeros in its transfer function located at the first sideband frequency (as the dominant switching harmonic). Therefore, the LLCL filter can significantly attenuate the first sideband harmonics. Although the high frequency roll-off rate of LLCL is the same as the \( L \) filter (with \( L = L_1 + L_2 \)), but the attenuation is higher with a factor of \( 1/(1 + L_2/L_1) \).

\[Y_{21,LF,LLCL} = \left| Y_{21}(s) \right|_{s = \omega_0} \approx \frac{1}{L_1 L_2 C_f \omega_0^2} \]  

\[Y_{21,HP,LLCL} = \left| Y_{21}(s) \right|_{s = \omega_0} \approx \frac{1}{L_1 + L_2/2 \left(1 + L_2/L_1 \right) \omega_0} \]  

(4)

In the low frequency range, as calculated in (5) and shown in Fig. 3, the characteristics of both LCL and LLCL filters are identical and the same as a simple \( L \) filter with \( L = L_1 + L_2 \). Therefore, the design of the current controller, regardless of the type of the filter, follows a
similar procedure.

\[ Y_{21,LF} = \frac{1}{(L_1 + L'_2)\omega} \approx \frac{1}{L_1\omega}, \quad (5) \]

Furthermore, the filters also present similar characteristics at high frequencies in terms of the converter current ripple attenuation, as presented in (6). As a result, regardless of the type of the filter, the converter current ripple is imposed only by the \( L_1 \) value.

\[ Y_{11,HF} = \frac{L'_2 + L_f}{[L_1 L'_2 + (L_1 + L'_2)L_1]\omega} \approx \frac{1}{L_1\omega}, \quad (6) \]

From the stability point of view, Bode plots of Fig. 3 show a sharp phase transition through \(-\pi\) along with a high magnitude peak at the resonance frequency that readily leads to instability.

4 Single-loop control of grid current with delay-based stabilisation method

The proportional-resonant (PR) controller is commonly used to eliminate the steady-state error in the stationary reference frame. The open loop transfer function of the system, in presence of a delay \( T_d \), can be derived as

\[ G_{ol}(s) = G_{pr}(s) Y_{21}(s) e^{-T_ds} \quad (7) \]

where \( G_{pr} \) is the transfer function of the PR controller. Open loop Bode plots for both LCL and LLCL filters are shown in Fig. 4.

4.1 Stabilisation of the single-loop controller

A digitally implemented control system has several delays, mainly originated from the sampling, analogue-to-digital conversion and computations. All these can be modelled with one sampling period \( (T_s) \) delay. Taking into account the \( 0.5T_s \) PWM transport delay of the converter system, a total delay of \( T_d = 1.5T_s \) exists in the control loop [30]. The delay introduces a linear phase lag and changes the phase characteristic of open loop system. In delay-based stabilisation method, the resonance frequency (compared with the sampling one) is selected low enough to take benefit of this phase lag. This way, the phase lag originated from the loop delay causes the sharp phase change occurs below \(-\pi\). If this sudden phase fall ends above \(-3\pi\), then the single loop system remains stable. Based on this analysis, the stable range of resonance frequency for \( f_s/6 < f_{res} < f_s/2 \) is already obtained as \( f_{res} < f_s/2 \) [18, 20].

Hence, Fig. 4 is plotted with the assumption that the resonance frequency is located in the stable range of \( f_{res} < f_s/2 \); therefore, the sharp phase change lies between \(-\pi\) and \(-3\pi\). In this situation, three phase margins can be recognised. At the low frequency range, the phase margin \( PM_1 \) characterises the current control loop performance and is determined in the controller design stage, which is presented in the next subsection. Indeed, the dynamic performance of the closed loop system is imposed by the stability margin \( PM_1 \), while stability margins \( PM_2 \) and \( PM_3 \) have negligible effects on the dynamic performance and only must be positive to ensure the stability. In practice, \( PM_2 \) and \( PM_3 \) should be selected large enough to remain positive in presence of parameter uncertainties of the filter circuit and ensure stability against the
resonance under all circumstances. The latter issue is fully attended in Section 5.

4.2 Tuning the controller parameters

As mentioned before, the characteristics of both LCL and LLCL filters at the low frequency range are identical and the same as a simple L filter with $L = L_1 + L_2$. Therefore, the PR controller for the current loop can be designed in the same way as for a system with the L filter. The tuning of PR constants is performed according to directions of [15]. Considering the negligible effect of the resonant terms of the PR controller on the gain crossover frequency, $\omega_{gc1}$ (for PM1), then $\omega_{gc1}$ can be calculated as

$$\omega_{gc1} = \frac{(\pi/2 - PM_{1d})}{T_d} \tag{8}$$

where $PM_{1d}$ is the desired value for the $PM_1$. To achieve the unity loop gain at the crossover frequency, the proportional gain is set to

$$k_p = \omega_{gc1}(L_1 + L_2) \tag{9}$$

The resonant gain is selected so that a negligible phase contribution at $\omega_{gc1}$ is achieved, i.e.

$$k_r = 0.02k_p\omega_{gc1} \tag{10}$$

5 LCL and LLCL filters design

Previous works have already proposed some criteria for selecting the parameters of high-order filters based on the limits on the current ripple of the inductors and the reactive power of the capacitor while the stability against the resonance is achieved with a passive or active damping method [22–26]. In this paper, the tuning of filter parameters is conducted in conjunction with the delay-based stabilisation technique. Indeed, the proposed step-by-step filter design algorithm satisfies the predefined constraints on the converter current ripple, grid current harmonics and the reactive power of the capacitor and at the same time, ensures adequate stability of the simple single loop current controller. Moreover, the effect of grid impedance on the filter performance and controller stability is taken into account, thus the proposed method is robust against a wide range of grid impedance variations.

In the following, the limits on the converter current ripple, the grid current harmonics and the reactive power of the filter capacitor are presented to simultaneously meet all the constraints.

5.1 Limits on the converter current ripple, the grid current harmonics and the reactive power of the filter capacitor

To design the parameters of an LCL or LLCL filter, some limits on current ripples and harmonics and reactive power should be considered, which are already discussed in [22, 24, 25]:

i. At the fundamental frequency, both filters act as a shunt capacitor that supplies reactive power. The reactive power by capacitor must be limited to a per cent, we call it $x_1$, of the rated power (generally $x_1 = 2.5$ to 5%).

ii. The converter side current ripple mainly determines the current rating of converter switches and is decided to be a per cent, we call it $x_2$, of the peak rated current of the converter $I_{F,\text{rated}}$ (generally $x_2 = 15$ to 40% [24]). In both cases, the maximum current ripple can be derived as [24]

$$\Delta i_{1,\text{max}} = \frac{V_d T_d}{8L_1} \leq x_1 I_{F,\text{rated}} \tag{11}$$

iii. The IEEE-1547.2-2008 standard recommend that all harmonics higher than 35th in the grid current must be lower than a per cent, $x_3 = 0.3\%$, of the rated current. To address the above requirement, as already discussed in Section 3, in the case of an LCL filter, the first sideband and in the case of an LLCL filter, the second sideband must be limited as

$$\begin{align*}
&\left\{I_{2,\text{sb1, max}} = \frac{V_{21,\text{HF},LCL} V_{\text{shl, max}}}{\omega_{res}} \leq x_1 I_{F,\text{rated}} \right. \\
&\left. I_{2,\text{sb2, max}} = \frac{V_{21,\text{HF},LLCL} V_{\text{shl, max}}}{\omega_{res}} \leq x_1 I_{F,\text{rated}} \right. \tag{12}
\end{align*}$$

where the $V_{\text{shl, max}}$ and $V_{\text{shl, max}}$ are the maximum values of all sideband amplitudes in the expected range of variations for $M_d$, which can be replaced from (2) or readily from Fig. 2.

5.2 Constraints for stability against the resonance

However by setting the resonance frequency in accordance to $f_{res} < f_{\text{crossover}}$ (sharp phase change occurs between $-\pi$ and $-\pi$), the stability of single-loop current regulator is achieved, but sufficient damping at resonance frequency, in terms of the phase margins $PM_2$ and $PM_3$, must be provided to ensure positive margins under the expected range of filter parameter uncertainties. In a weak grid, the grid impedance variations lead to a wide uncertain range of the resonance frequency that must be also explored. Therefore, according to the grid impedance condition, the expected range of variation of resonance frequency should be first determined, then the filter parameters must be chosen such that the resonance frequency range always remains inside the stable range of $f_{res} < f_{\text{crossover}}$ with appropriate margins.

In the case of a PR controller, where the phase change due to the resonance network just affects the low frequencies (obvious in Fig. 4), then the open-loop system phase in vicinity of the resonance frequency can be assumed as

$$\Delta G_{o}(j\omega) = -\frac{\pi/2 - \omega T_d}{3\pi/2 - \omega T_d} \omega < \omega_{res} \tag{13}$$

As it can be seen in Fig. 4, the gain crossover frequencies for $PM_2$ and $PM_3$ are slightly lower and higher than the resonance frequency, respectively. With a good approximation, both gain crossover frequencies can be replaced with the resonance frequency. Therefore, the criterion that the sharp phase change lies in the stable range of $-\pi$ to $-\pi$, with some margins as we already called $PM_2$ and $PM_3$, respectively, imposes that

$$\begin{align*}
&\frac{-\pi/2 - \omega_{res} T_d}{3\pi/2 - \omega_{res} T_d} = -\pi - PM_2 \\
&\frac{-3\pi/2 - \omega_{res} T_d}{3\pi/2 - \omega_{res} T_d} = -3\pi + PM_3
\end{align*} \tag{14}$$

Hence, the appropriate resonance frequency range that meets the desired phase margins $PM_{2d}$ and $PM_{3d}$ can be concluded as

$$\frac{\pi/2 + PM_{2d}}{T_d} \leq \omega_{res} \leq \frac{3\pi/2 - PM_{3d}}{T_d} \tag{15}$$

It is worth mentioning that if one replace $PM_{2d} = PM_{3d} = 0$ and $T_d = 1.5T_s$, then (15) simplifies to $f_{res} < f_{\text{crossover}}$.

On the other hand, the possible variation range for the resonance frequency with variation of $L_g$ between zero and infinity can be
5.3 Choosing the proper phase margins

As mentioned before, the values of the PM2d and PM3d must be selected properly to avoid instability due to uncertainties and drifts of filter parameters. A change in the filter parameters, may lead to a decrease of the minimum or an increase of the maximum value of the resonance frequency, already predicted by (16), which may push it outside the stable range. Considering (16), the worst case for $\omega_{res,\min}$ occurs when all filter parameters have positive uncertainty (i.e. the actual values be higher than the designed ones), and for $\omega_{res,\max}$ it occurs when all filter parameters have negative uncertainty (i.e. the actual values be smaller than the designed ones). Assuming that the actual values of filter parameters, with a subscript ‘a’, are related to the designed values as $L_a = k_L L$ and $C_a = k_C C$, then, the actual resonance frequency can be derived as

$$\omega_{res,a} = \frac{\omega_{res}}{\sqrt{k_L k_C}}. \quad (19)$$

For the given values of $k_L$ and $k_C$, with the goal of achieving the zero phase margin for the worst case of parameter uncertainties, the values of the PM2d and PM3d can be derived using (15), (17), and (18) as

$$\begin{align*}
\text{PM2d} &= (\sqrt{k_L k_C} - 1)\pi / 2, \\
\text{PM3d} &= 3(1 - \sqrt{k_L k_C})\pi / 2. \quad (20)
\end{align*}$$

5.4 Proposed design procedure

Two sets of limits and constraints are already presented in Sections 5.1 and 5.2. Satisfying them simultaneously results in a filter with good harmonic attenuation and sufficient stability margins, regardless of the grid impedance conditions. After choosing the limits $x_1$, $x_2$ and $x_3$ and the range of parameters uncertainty ($k_L$ and $k_C$), according to the application requirements, the LCL and LLC filters parameters can be calculated according to the following straightforward step-by-step procedure:

i. For both filters, the capacitor value is initially determined from the reactive power limit as

$$C_t \leq \frac{x_1 P_{\text{rated}}}{\omega_a V_g} \quad (21)$$

where $P_{\text{rated}}$ is the rated power and $V_g$ is the rms value of the grid voltage. The second limit for the capacitor in the case of the LCL filter is obtained by removing $L_1$ from (11) and (17)

$$C_t \leq \frac{8x_2 P_{\text{rated}}}{T_s V_{dc}} \left( \frac{1 - \omega_{res}^2}{\omega_{res}^2} \right). \quad (22)$$

and in the case of the LLC filter and with the extra assumption that the filter zero is exactly located at the first sideband, i.e. $1/(L_1 C_3)^{1/2} = \omega_a$, we have

$$C_t \leq \frac{8x_2 P_{\text{rated}}}{T_s V_{dc}} \left( \frac{1}{\omega_{res}^2} - 1 \right). \quad (23)$$

Hence, the minimum of (21) and (22) determines the $C_t$ value for the LCL and the minimum of (21) and (23) determines the $C_t$ value for the LLC filter.

ii. in the LLC filter, the first sideband is extremely attenuated by selecting the $L_f$ value as

$$L_f = \frac{1}{C_t \omega_a^2}. \quad (24)$$

iii. With the calculated values for $C_t$ and $L_f$, the value of $L_1$ can be determined using (17) as:

$$L_1 = \frac{1}{C_t \omega_{res}^2} - L_f. \quad (25)$$

It should be noted that the ripple limit on the converter side current is considered in selection of $C_t$, therefore with this value for $L_1$ the ripple limit is also satisfied.

iv. The value of $L_2$ should be determined to achieve a high attenuation of grid current harmonics based on the standards and at the same time obtain a guaranteed stability. Therefore, (4) and (12) are combined to provide a limit for the grid current attenuation for the LCL and the LLC filters, presented in (26) and (27), respectively. Since the worst case of harmonic attenuation occurs in a stiff grid with $L_g = 0$, the following results are calculated with the assumption that $L_g = 0$.

$$L_2 \geq \frac{V_{\text{sh2,\max}}}{V_{\text{rated}\times x_1 P_{\text{rated}}}} \quad (26)$$

$$L_2 \geq \frac{V_{\text{sh2,\max}}}{2\omega_{res} x_1 P_{\text{rated}}} - L_1 \quad (27)$$

The values of $V_{\text{sh2,\max}}$ can be determined from (2) or Fig. 2 for the expected range of $M_s$. From the stability point of view, the limit on the value of $L_2$ is achieved using (18) as

$$L_2 \geq \frac{L_1 (1 - L_1 C_3 \omega_{res}^2)}{(L_1 + L_f C_3)^{1/2} \omega_{res}^2 - 1}. \quad (28)$$

Again for the LCL filter, one must replace $L_f = 0$. Hence, the maximum of (26) and (28) determines the $L_2$ value for the LCL.
and the maximum of (27) and (28) determines the $L_2$ value for the LLCL filter.

Fig. 5a summarises the four step design algorithm for the LCL and the LLCL filters.

### 5.5 Design example

The proposed step-by-step design procedure for a single phase converter with $V_g = 220$ V rms/50 Hz, $V_{dc} = 388$ V ($M_a = 0.8$), $P_{rated} = 3$ kW and $f_c = 8$ kHz ($f_s = 2f_c$) is performed in the following and the results are summarised in Fig. 5b. Choosing the value of parameter uncertainties depends on the type of used capacitors and inductors. Here, the deviation of ±30% and ±20% is considered for inductances and capacitors, respectively (i.e. $0.7 < k_L < 1.3$ and $0.8 < k_C < 1.2$). Therefore, the proper stability margins can be calculated from (20) as $PM_{2d} = 23^\circ$ and $PM_{3d} = 68^\circ$.

i. With the reactive power limit of $x_1 = 5\%$, the capacitor should be $< 9.86 \mu F$ and with the current ripple limit of $x_2 = 30\%$, the maximum allowed value is 4.3 $\mu F$ for the LCL and 4.1 $\mu F$ for the LLCL filter. In each case, the lower value considering the available capacitors should be chosen. Therefore, the selected values are 4 $\mu F$ for both LCL and the LLCL filter.

ii. With the chosen value for the capacitor of the LLCL filter, the value of $L_2$ is calculated from (24) as 25 $\mu H$.

iii. The value of $L_1$ is calculated from (25) as 570 and 540 $\mu H$ for the LCL and the LLCL filter, respectively.

iv. The limit on the value of $L_2$ is calculated from (26) and (27). With a dc link voltage of 388 V, which means $M_a \geq 0.8$ (considering the grid impedance volt-drop), the maximum value of the converter voltage for the first and the second sidebands is determined from Fig. 2 as $0.32V_{dc}$ and $0.12V_{dc}$, respectively. Thus the value of $L_2$ is calculated as 940 and 150 $\mu H$, for the LCL and the LLCL filter, respectively. Furthermore, the value of $L_2$ is calculated from (28) as 260 and 210 $\mu H$, for the LCL and the LLCL filter, respectively. Hence, the maximum value for $L_2$ is chosen as 940 and 210 $\mu H$, for the LCL and the LLCL filter, respectively.

The designed filter parameters and the resultant limits and constraints are summarised in Table 1. As expected, the achieved total inductance value for the LCL filter is higher than the LLCL filter. For both filters, the $C_f$ value is determined by the stability constraint $PM_{2d}$. The $L_2$ value is determined by the stability constraint $PM_{3d}$ for the LLCL filter and by the harmonics attenuation limit ($x_3$) for the LCL filter.

### 6 Experimental verification

To confirm the theoretical achievements, a 3 kW single phase VSC is constructed with the filter parameters already calculated in the previous section. The control algorithm is implemented on a TMS320F28335 floating point digital signal controller from Texas Instruments.

![Fig. 6 Steady state waveforms and grid current harmonic spectrum with $L_g = 0$](image)

- a LCL filter
- b LLCL filter

Table 1 Designed filter example

<table>
<thead>
<tr>
<th>Filter type</th>
<th>Filter parameters</th>
<th>Design limits and constraints</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$L_1$, $\mu H$</td>
<td>$L_2$, $\mu H$</td>
</tr>
<tr>
<td>LCL</td>
<td>570</td>
<td>940</td>
</tr>
<tr>
<td>LLCL</td>
<td>540</td>
<td>210</td>
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For both LCL and LLCL filters, the current controller parameters are calculated from (8)–(10), with a desired phase margin of $\text{PM}_{1d} = \pi/3$ as: $\omega_{gc1} = 5.58\ \text{krad/s}$, $k_p = 8.4$ (LCL)/4.2 (LLCL) and $k_r = 940$ (LCL)/470 (LLCL). Besides the fundamental component, the set of harmonics to be regulated are selected as $n = 3, 5, 7$ and 9, based on the grid dominant harmonics. The PR controller is discretised by the impulse invariant method [31].

### 6.1 Steady state performance

Fig. 6 shows the steady state performance of the converter with both designed filters and $L_g \approx 0$. Although the grid voltage is highly polluted with a THD $v = 3.8\%$, the injected current is highly sinusoidal with a THDI of about 1.2% (LCL) and 2.1% (LLCL). These results indicate that the filter parameters, as well as the current controller are designed correctly. The spectra of grid current for both filters confirm adequate switching harmonic attenuation according to design constraints. The amplitude of the dominant switching harmonic for the LCL filter (at the first sideband) and the LLCL filter (at the second sideband) is 0.28 and 0.23%, respectively, which both meet the IEEE–1547 standard requirements.

The low frequency spectra (zoomed view of Fig. 6 for $n < 50$), compared with the IEEE 1547 limits are shown in Fig. 7. As can be seen, both filters successfully meet the IEEE 1547 standard limits. The low frequency harmonics appeared in the grid current are mainly due to the grid voltage background harmonics and the non-idealities of the real system, such as the switching dead times and the inductor core non-linearity and saturation.

![Fig. 7](image_url) Low frequency harmonics of grid current compared with IEEE 1547 limits

![Fig. 8](image_url) Steady state waveforms with $L_g = 3.7\ \text{mH}$

- a LCL filter
- b LLCL filter

![Fig. 9](image_url) Dynamic performance in response to step changes in the reference active power from 2 to 3 kW and vice versa with $L_g \approx 0$ and $L_g = 3.7\ \text{mH}$

- a LCL filter
- b LLCL filter
harmonics are attenuated primarily by the inductors in the filter structure. The harmonic compensation network in the current controller can also improve the results. It should be noted that, the higher magnitude of the low frequency harmonics and consequently higher value for THDi of the LCL filter compared with the LCL one is mainly due to the lower total inductance \((L = L_1 + L_2)\) of the LLC filter, which makes the converter performance, in terms of the quality of the currents, more susceptible to non-idealities of the real system. Obviously, this effect decreases in presence of a higher grid impedance value and as it can be seen in Fig. 8, in such a situation, \(L_g = 3.7\) mH, almost similar THDIs are obtained for both LCL and LLC filters. It can be concluded that, for the filters with lower total inductance, which have the advantages of lower cost and size, the low frequency harmonic suppression level is lower and the THDi value is higher. Moreover, Fig. 8 confirms that both converter systems remain stable and keep their excellent performance, even in presence of a very weak grid, where a large range of impedance values and background harmonic distortions are possible.

The ripple content of the converter side current is almost equal to the capacitor current \(i_C\). As it can be seen in Fig. 6, the maximum peak current ripple is about 6 A, which translates to 30% of the peak rated current \(i_{L\text{peak}}\).

### 6.2 Dynamic performance

The dynamic performance of the VSC is investigated in response to step changes in the reference active power from 2 to 3 kW and vice versa, which is shown in Fig. 9. In the case of \(L_g \neq 0\), results confirm the excellent transient response of both LCL and LLCL filtered converters, where the converter current changes almost instantaneously with a negligible oscillation. The transient performance with \(L_g = 3.7\) mH shows that the converter remains stable, however the transient response is slowed down by the effect of the grid impedance. This appreciable performance can be attributed to the proper tuning of the stability margins of the converter system, \(\Delta P_m\) to \(\Delta i_{L\text{peak}}\). It should be noted that, in all transient waveforms any oscillation corresponding to the filter resonance does not appear.

### 7 Conclusion

In this paper, a simple and straightforward design algorithm for the LCL and the LLCL filtered VSCs with a delay-based stabilised single-loop current controller is proposed that takes into account both filtering requirements and stability issues simultaneously. In this approach the filter parameters are designed analytically, and without any iteration or trial and error procedure. The constraints on the filter parameters due to the stability requirements are carefully extracted for the delay-based stabilisation method, considering the effect of grid impedance variations. As a result, the designed converters meet the filter design requirements and are also stable without the need for additional damping resistors in the filter circuit or digital filters and extra measurements in the control loop. The experimental results confirm the theoretical achievements.

### 8 References