PWM Rectifier with LCL-Filter using different Current Control Structures

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«Voltage Source Converter (VSC)», «Converter Control», «Active Damping»

Abstract

The control of a PWM rectifier with LCL-filter using a minimum number of sensors is analyzed. In addition to the DC-link voltage either the converter or line current is measured. Two different ways of current control are shown, analyzed and compared by simulations as well as experimental investigations. Main focus is spent on active damping of the LCL filter resonance and on robustness against line inductance variations.

I. Introduction

PWM rectifiers are mostly used in regenerative energy systems and in adjustable speed drives where regenerative braking is required. They offer control of the power factor as well as the DC-link voltage while emitting less current harmonics to the grid compared to passive diode rectifier bridges. Besides L-filters, LCL-filters are used for the grid connection. LCL-filters give advantages in costs and dynamic as smaller inductors can be used compared to L-filters in order to achieve the necessary damping of the switching harmonics. As a drawback the filters tend to oscillations with the filter resonance frequency. One way of damping this resonance is the use of resistors in series with the filter capacitors but this causes considerable power losses. Another way is to damp the resonance actively by control algorithms without increased power losses. The approaches presented in the literature differ in signals used for the control, number of sensors, control complexity and performance.

In [1] and [2] the control of the converter current is shown using only the converter currents. The damping is achieved by adding to the PI controller digital filters of second [2] or fourth [1] order in the open loop that attenuate the resonance frequency. By that the number of sensors is not increased by the active damping method. A similar behavior of these digital filters in the open loop can be achieved by feedback of the voltage across the filter capacitors with lead elements [3, 4] and add it to the PI controller output. This requires additional sensors for the voltage measurement. Instead of using a lead element a high pass filter is used in the filter capacitor voltage feedback path in [5]. A digital converter current control is designed in [6] by using the converter current and filter capacitor voltage measurement.

In [7] the line current control using only the line current as feedback signal for a PI controller is shown to be stable even without any kind of additional damping, but line current distortions are clearly visible. In [8, 9] an additional feedback of the filter capacitor current is used for damping the resonance to enhance the control performance. It should be noted, that there are current control approaches using all three states of the filter that is line and converter currents as well as the filter capacitor voltages [10, 11, 12]. From an industrial point of view it is desired to have a minimum number of sensors and therefore these approaches are not further considered in this paper.
Besides the choice of the control structure, including the necessary sensors, the controller tuning and filter design as well as the switching and sampling frequency have big impact on the system behavior concerning stability, transient and steady state behavior, robustness against distortions and parameter variations. In the mentioned papers mostly different system settings are used and therefore a comparison of the different structures based on the literature and the choice of the most suitable one for a certain application is difficult to do.

For the analysis in this paper two current control structures are selected based on the industrial requirement of minimum number of sensors. On the one hand the control structure using only the converter current [2] and on the other hand using only the line current [7] are selected. The scope of this paper is to show the design, analysis and comparison of these two different current control structures using the same PWM rectifier with LCL-filter. The influence of different system parameters on the control performances is analyzed.

In this paper the analyzed system is described in chapter II. The control system with two different current control structures is presented in chapter III. Chapter IV shows simulation results and the experimental results will be given in chapter V of the final paper. The paper will be closed by a comparison of the two investigated control structures and a comparison.

II. System Description

The analyzed system is shown in Figure 1 and the nominal system data in Table I. The PWM rectifier is connected to the grid by an LCL-filter and a transformer. The transformer inductance and the mains side filter inductance are combined in L_m for modeling and simulation purpose. The PWM rectifier is loaded by a resistor. To investigate two different current control techniques different measurement signals will be used. The DC-link voltage will be measured in both cases and in addition either the line or the converter currents. For synchronizing purpose the line voltage is also measured. The control algorithms are implemented using a dSpace 1106 board.

The relative short circuit voltage of the LCL-filter is mostly given at the fundamental frequency by the total inductance L_T=L_c+L_m and is chosen to be 3% at 7,6 A (5,5 kW). The higher the ratio L_c/L_m is chosen the better the switching frequency components of the converter currents are damped. By that the losses in the converter side inductance are lower. But if a small L_m is chosen the influence of the line impedance variations on the resonance frequency is increased. This can cause stability problems, especially in weak grids. Here the ratio L_c/L_m is set to 2. Figure 2 shows the transfer function |I_L(j\omega)/U(j\omega)| of the LCL-filter with two different filter capacitances (U(j\omega): Converter output voltage). On the one hand it becomes clear that the higher frequency current components at the switching frequency (and multiple of it) get better damped for higher filter capacitances. On the other hand stability problems can arise if the filter capacitances are chosen too high, that is a too low resonance frequency, as will be shown in the following sections. The investigations in this paper will be done with two different filter capacitances (see Table I).
The current control will be done in rotating dq-coordinates aligned to the line voltage space vector. Therefore the mathematical model of the LCL-filter is given directly in dq-coordinates (\(u_{Cf}\): filter capacitor voltage):

\[
L_m \frac{d i_{dq}}{dt} = u_{dq} - u_{Cf}^{dq}
\]

\[
C_f \frac{d u_{Cf}^{dq}}{dt} = i_{L}^{dq} - i_{C}^{dq}
\]

\[
L_c \frac{d i_{C}^{dq}}{dt} = u_{Cf}^{dq} - u_{dq}
\]

The transformation from three phase values in stationary reference frame into space vectors in rotating reference frame can be found in the appendix. Neglecting the losses in the filter and converter the relation between the mains side and DC side active power can be written as: \(u_{DC} i_{DC} = 3 i_{Ld} u_{Ld}/2\). Using this relation the dynamics of the DC-link voltage can be written as follows:

\[
C_{DC} \frac{d u_{DC}}{dt} = \frac{3}{2} i_{Ld} u_{Ld} - i_{Load}
\]

### III. Control Design

The complete control system is shown in Figure 3. The outer DC-link voltage control only consists of a PI controller with an anti-wind up mechanism. The control parameters are tuned according to the symmetrical optimum [14].

The two different structures of current control are shown in Figure 4 and will be explained in more detail in the following sections. They differ in the measured currents. Both are using dq-coordinates aligned to the line voltage vector which is measured directly. The angle of the line voltage fundamental is determined by a PLL algorithm that is executed with a frequency of 20 kHz. The control is executed with the switching frequency (5 kHz). In both structures a decoupling between the d and q component and an anti-wind up mechanism as well is used. The time constant of the PI control integrators are chosen to compensate the slowest time constant (\(T_I = R_T/L_T\); \(R_T\) models the parasitic resistance of \(L_T\)). The proportional gain of the PI controller are set to \(K_p = -K*L_T*f_{sw}/2\). The parameter \(K\) is used for fine tuning and for visualizing purpose as well (\(K = 1\) corresponds to the modulus optimum [14]).

#### A. Line Current Feedback

The line current control structure is shown in Figure 4 (left). Note that besides the PI controller there is no additional damping of the resonance. Figure 5 (left) shows the root locus of the closed loop using an LCL-filter with a high resonance frequency (\(C_f=16\ \mu F\)). The direction of the poles with increasing proportional gain is illustrated by arrows. It can be seen that there is a range of proportional gains for

### Table I: System Nominal Data

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Line voltage (line-line rms)</td>
<td>400 V</td>
</tr>
<tr>
<td>Line frequency (f_L=\omega_L/(2\pi))</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Filter inductance (L_c) (converter side)</td>
<td>2 mH</td>
</tr>
<tr>
<td>Filter inductance (L_m) (mains side)</td>
<td>1 mH</td>
</tr>
<tr>
<td>Filter capacitances (C_f)</td>
<td>16/24(\mu F)</td>
</tr>
<tr>
<td>DC-link capacitance (C_{DC})</td>
<td>2200 (\mu F)</td>
</tr>
<tr>
<td>DC-link voltage (U_{DC})</td>
<td>650 V</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>5 kHz</td>
</tr>
</tbody>
</table>
which the system is stable, that is all roots inside the unity circle. The originally unstable resonance poles are attracted inside the unity circle with increasing $K$. The proportional gain is limited by the fact that the low frequency poles finally get pushed outside the unity circle. In Figure 5 (left) the pole locations for $K=1$ (modulus optimum) are marked. This parameter design results in a bandwidth of 1.68 kHz and a stability margin of 1.57.

If the filter is designed with a higher capacitance the resonance frequency is lower and the system stability gets worse, as can be seen in Figure 6 (left). The originally unstable resonance poles are attracted inside the unity circle only for a small range of proportional gains. The design according to the modulus optimum ($K=1$, not marked in Figure 6 (left)) results in a bandwidth of 1.34 kHz and a stability margin of 1.24. Even for nominal system parameters the system is quite close to the stability limit.

In Figure 6 (left) there are marked pole locations for two different proportional gains. Optimal resonance damping is achieved with $K=0.9$. By increasing $K$ the damping gets lower and can cause stability problems if it is chosen too high ($K_{\text{max}}=1.24$). In order to illustrate the stability problem in case of an additional line inductance (still using the same control parameters) the poles for $K=1.15$ are also shown in Figure 6 (left). Without a line inductance the system is still stable but if there is one of $L=500 \mu\text{H}$ the system gain as well as the resonance frequency are changed and the root locus shown in Figure 6 (right) indicates almost instability even for $K=1.15$. (In this case the resonance pole is shifted to approx. 825 Hz.) Figure 5 (right) indicates that there are not such stability problems related to line inductance variations in case of higher filter resonance frequencies.
B. Converter Current Feedback

The current control using the converter current is shown in Figure 4 (right). In order to achieve unity power factor on the mains side it is necessary to compensate the filter phase shift. Therefore the reference of the q-axis is set to: \[ i_q^* = -\omega_L C_f \left( u_d + R_i c_i d^* \right) / (1 - \omega_L^2 L_c C_f). \]

Figure 7 (left) and Figure 8 (left) show the root loci of the closed loops using only PI control without additional active damping (AD) for different resonance frequencies. It can be seen that in any case additional damping is necessary to stabilize the system. Using second order filter as shown in Figure 4 (right), that is adding two poles and zeros in the open loop (see Figure 7 (middle) and Figure 8 (middle)), the system can be stabilized. For the purpose of unity DC gain of the active damping function its proportional gain is set to \[ K_{AD} = (1 - 2 \Re{p_0} + |p_0|^2) / (1 - 2 \Re{z_0} + |z_0|^2). \] The choice of the pole and zero locations is one essential step in the AD design. As can be seen the AD zeros attract the unstable resonance poles. To get the root locus into the unity circle the resonance poles and zeros should lay between the AD poles and zeros. There are stability problems possible if the resonance frequency of the system is decreased, for example by an additional line inductance. Figure 7 (right) and Figure 8 (right) visualize the root loci including additional line inductances of 500 µH that decrease the resonance frequency (corresponds to decrease of the angle of the resonance poles and zeros) using the same control parameters as before. The system gets unstable since the resonance zeros start to interact with the low frequency poles.
The wider the angle in between the AD poles and zeros the more the system resonance may vary without leading to instability. On the other hand the region of possible locations of the AD poles and zeros is limited. Too low angles of the AD poles will make them interacting with the low frequency dynamics, especially for low resonance frequencies. By that the system gets unstable even without parameter uncertainty (see Figure 9). So, only in case of higher resonance frequency (not too close to the low frequency dynamics) the robustness against smaller system parameter variations could be influenced by the position of the active damping poles and zeros by keeping a wide angle in between them. In general, the sensitivity against parameter variations that change the resonance frequency, for example the filter capacitance, is higher in comparison to line current feedback.

In Figure 7 (middle) and Figure 8 (middle) there are marked the pole locations for K=1 (modulus optimum). In case of high resonance frequency a bandwidth of 1,19 kHz and a stability margin of 1,63 is achieved. A bandwidth of 1,00 kHz and a stability margin of 1,6 is achieved if a low resonance frequency is used.

IV. Simulation Results

Simulations are carried out with Matlab/Simulink in order to analyze the system behavior and to confirm the theoretical analysis. The switching of the line side PWM converter is included by using the toolbox PLECS. Simulations are performed with an active power of 5,5 kW. The line voltage is assumed to be ideal, that is no harmonics and unsymmetries. If not stated different K=1 is used.

A. Line Current Feedback

Figure 10 shows two steady state spectra using an LCL-filter with a capacitance of $C_f=16 \mu F$. It illustrates the functionality of the filter concerning the switching noise as well as proper damping of resonance. Instead of disturbing, an additional line inductance of 500 $\mu$H contributes to the damping of the switching without increasing the resonance component. Therefore the total harmonic distortion (THD) is further decreased to 1.05 %, see Figure 10 (right).

Figure 11 indicates that the switching components of $i_L$ get better filtered out by using higher capacitances ($C_f=24 \mu F$). With K=0,9 the resonance is well damped and the THD = 1.17% is achieved, see Figure 11 (left). As expected from Figure 6 (left), the resonance components are slightly higher with K=1,15 whereas the THD of 1.81% is still acceptable, see Figure 11(right). According to Figure 6 (right) it is expected that an additional line inductance of 500 $\mu$H leads the system to the stability limit if the control parameters remains unchanged and this is confirmed by simulations, see Figure 11 (right). There is a peak around the frequency of 825 Hz which the resonance poles get shifted to, as expected from Figure 6 (right). The resonance is considerably disturbing the line current as well as the converter current. Additionally to the resonance itself, the converter current contains also additional components in the sideband of the switching frequency at $f_{sw}\pm 825$Hz.

**Figure 10:** Simulated steady state spectra with PI line current control (5.5 kW motor load, $C_f=16 \mu F$).

Left: $L_L=0 \mu H$, right: $L_L=500 \mu H$. 

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B. Converter Current Feedback

In Figure 12 (left and middle) steady state spectra using $C_f=16 \ \mu F$ for different $K$ are shown. It can be seen that increasing $K$ results in only slightly higher resonance damping. A THD of 1.8 % can be achieved which is higher compared to the line current feedback (1.38%). As can be seen in Figure 12 (right), using higher LCL-filter capacitance ($C_f=24 \ \mu F$) results in lower switching noise but even if kept stable the resonance damping is quite small. The THD is considerably increased to 6.6 %. Note that no harmonics or unsymmetries in the line voltage, that would further increase the THD, are included in the simulations.

V. Measurement Results

Experimental investigations are carried out with a PWM rectifier system shown in Figure 1. Current measurements for visualization purpose are done using Tektronic Current probes. Tests are done with an active power of 6 kW with resistive load, see Table I for system settings. The background distortion of the line voltage was measured to THD = 4.4%.

A. Line Current Feedback

Figure 13 (left) shows the line and converter currents using $C_f=16 \ \mu F$ for different $K$ are shown. It can be seen that increasing $K$ results in only slightly higher resonance damping. A THD of 1.8 % can be achieved which is higher compared to the line current feedback (1.38%). As can be seen in Figure 12 (right), using higher LCL-filter capacitance ($C_f=24 \ \mu F$) results in lower switching noise but even if kept stable the resonance damping is quite small. The THD is considerably increased to 6.6 %. Note that no harmonics or unsymmetries in the line voltage, that would further increase the THD, are included in the simulations.
Figure 13: Measured currents with PI line current control (6.0 kW resistive load, $C_f = 16$ $\mu$F, $K=0.5$). Left: Waveforms without harmonic compensator (Ch 1: Line phase current (10A/Div), Ch 2: Converter phase current (10A/Div)), middle: Waveforms with harmonic compensator (Ch 1: Line phase current (10A/Div), Ch 2: Converter phase current (10A/Div)), right: steady state spectra.

Figure 14: Measured currents with PI line current control (6.0 kW resistive load, $C_f = 24$ $\mu$F, $K=0.5$): Steady state spectra. Left: Nominal Case ($L_L=0$ $\mu$H), right: $L_L=500\mu$H.

A considerable fifth harmonic component and a high THD of 15.8%. Therefore a harmonic compensator for reducing the fifth harmonic according to [15] is inserted. In fact the proportional gain of the PI controller has to be reduced; otherwise the harmonics get amplified. The result is shown in Figure 13 (middle). The THD is reduced to 6.77%. In Figure 13 (right) the spectra of the currents are presented. The switching noise of the line current is filtered out well and no resonance components are visible. The difference to the simulated THD of the line current of 1.38% stems from the low frequency harmonics.

Figure 14 (left) shows the steady state spectra of the currents using an LCL-filter with $C_f = 24$ $\mu$F (including a harmonic compensator). In comparison to Figure 13 (right) there is no relevant difference in damping of the switching noise observable since it was already filtered out well with $C_f = 16$ $\mu$F. The THD is slightly reduced to 6.47%. By increasing the line side inductance by 500 $\mu$H without changing the control parameters no stability problems are observed, see Figure 14 (right). The THD is slightly reduced to 6.3% due to different low frequency harmonic content. Instead of exciting the resonance an increase of the PI controller gain leads to a considerable amplification of the fifth harmonic. Therefore no results with excited resonance are obtained. This issue has to be addressed in more detail in future investigations.

B. Converter Current Feedback

Figure 15 (left) shows the line and converter currents using an LCL-filter with $C_f = 16$ $\mu$F (including a harmonic compensator). In comparison to Figure 13 (right) a slightly higher THD of the line current of 7.5% is achieved. Due to the fact that the harmonic compensator is directly acting on the converter current instead of the line current a lower THD of 15.1% of the converter current is obtained (compared to the line current feedback). Nevertheless, the switching noise of the line current is filtered out well and resonance is well damped.
As can be seen in Figure 15 (right), using a filter with \( C_f = 24 \ \mu F \) results in a slightly higher THD of the line current (7.62\%) but on the other hand the THD of the converter current is reduced to 13.8 \%. This also confirms that the harmonic compensator works better in case of line current feedback since the line side THD is of more importance for the grid connection. Again, the resonance is damped well.

### VI. Comparison

In Table II the properties of both investigated current control structures are summarized. \( I_L \) feedback turned out to be more robust against variations of resonance frequency. Due to the need of placing the active damping poles and zeros information about the resonance frequency is necessary for \( I_c \) control design whereas it is not in case of \( I_L \) feedback. Especially in weak grids the line inductance can have a big impact on system stability in case of \( I_c \) feedback. In general, parameter variations, for example of the filter capacitance, can lead to instability in case of \( I_c \) feedback. As long as the resonance frequency is not lowered too much \( I_L \) feedback remains stable. Since the filter phase shift has to be compensated in case of \( I_c \) feedback using system parameters the outer control loop of power factor is more robust with \( I_L \) feedback.

In practical application considerable distortions due to low frequency (5th) showed up and an additional compensator was necessary. The quality of the controlled currents was enhanced. In case of \( I_L \) feedback the quality of \( I_c \) was more enhanced than that of \( I_L \). Since the main interest is on line side quality it can be concluded that the line current control showed better behavior.

The effort and complexity of control design is higher for converter current feedback because the poles and zeros of the active damping are to be tuned and implemented. From an industrial point of view it is an advantage to have the current sensors inside the converter because they can be used for protection purpose as well. In terms of dynamics \( I_L \) feedback showed a slightly better behavior.

### Conclusion

In this paper two different current control structures for a PWM rectifier with LCL-filter have been investigated. Each structure is using only one set of current sensors. The control design has been
presented and discussed and simulation results as well as experimental test results have been shown and discussed, too. Main focus has been put on active damping of the LCL filter resonance and on robustness against line inductance variations. In case of line current feedback a pure PI controller can be used whereas in case of converter current feedback additional active damping is necessary. In practical application low frequency line voltage harmonics caused problems that were reduced by using additional compensators. The resonance is effectively damped with both control structures. For both structures it turned out that LCL-filter with lower resonance frequencies are more difficult to control. Line current feedback showed a more robust behavior than converter current feedback. Additionally, the controller tuning is more complex in case of converter current feedback due to the need of designing the necessary additional damping.

Appendix

The transformation from three phase values in stationary reference frame $(x_a, x_b, x_c)$ into space vectors in stationary reference frame $(x_α, x_β)$ and into space vectors in rotating reference frame is done with [16]:

$$\begin{align*}
\underline{x}_α & = \frac{2}{3} (x_a + x_β e^{j120°} + x_β e^{-j120°}), \\
\underline{x}_β & = \underline{x}_α e^{-jφ}, \quad \underline{x}_α = \underline{x}_α e^{-jφ},
\end{align*}$$

References


