Grid-connected converters are widely used in energy production and consumption. The grid voltage is typically measured, and hence voltage sensors are required for control of these converters. However, it is beneficial from system cost and reliability point of views that the converters are able to estimate the grid voltage and they can operate without the grid-voltage sensors. This thesis deals with estimation methods for grid-voltage sensorless control of the converters equipped with an LCL filter. For the grid-voltage estimation, an adaptive observer is proposed, analyzed, and experimentally tested.
Estimation Methods for Grid-Voltage Sensorless Control of Converters Equipped with an LCL Filter

Jarno Kukkola

A doctoral dissertation completed for the degree of Doctor of Science (Technology) to be defended, with the permission of the Aalto University School of Electrical Engineering, at a public examination held at the lecture hall T2 of the Computer Science building on 20 January 2017 at 12.

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Grid-connected three-phase power electronic converters, equipped with an inductor-capacitor-inductor (LCL) filter, are widely used in energy production and consumption. The LCL filter effectively attenuates the switching harmonics of the converter; however, a drawback of the filter is its resonating behavior. The resonance can be damped by means of converter control, and the damping becomes easier and more effective, if the states of a dynamic full-order LCL-filter model are known (measured or estimated). The grid voltage is typically measured for converter control, but replacing voltage sensors with estimation may reduce system costs at low power ratings. Alternatively, a voltage estimation in parallel with the measurement increases system reliability. This thesis proposes estimation methods for grid-voltage sensorless control of converters equipped with an LCL filter. Only the converter AC currents and the DC-link voltage are measured for the control system. An adaptive observer is proposed for a combined state and grid-voltage estimation based on the full-order LCL-filter model. For unbalanced grid conditions, the observer is augmented with a disturbance model for the negative-sequence grid-voltage component. The nonlinear estimation-error dynamics of the observer are linearized and theoretically analyzed. The proposed observer is experimentally tested as a part of a grid-voltage sensorless control system, where the estimated states are applied in state-space current control. Based on the linearized dynamics, an analytic design procedure is presented for the observer in the continuous and discrete-time domains. The design procedure retains a link between the observer gains and dynamic performance, thus resulting in symbolic expressions for the gains as a function of the performance specifications and LCL-filter model parameters. The proposed observer can estimate the grid-voltage magnitude, frequency, and angle. In unbalanced grid conditions, the augmented observer can also estimate the negative-sequence component of the grid voltage. Estimated quantities can be used in the converter control system. The analytic design procedure enables the proposed estimation methods to be applied with different converters and LCL filters and further enables automatic tuning of the methods, for example, at the converter startup. The proposed methods can be applied, for example, in active-front-end rectifiers of motor drives or solar inverters.
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Estimointimenetelmiä LCL-suodattimella varustettujen suuntaajien verkkojänniteanturittomana säätöön

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I started as a doctoral candidate in 2013. During the journey toward the doctoral degree, the Electric Drives Group offered an inspiring working environment, and I am grateful to the head of the group, my supervisor, Prof. Marko Hinkkanen for his guidance and support during this work. His professional comments on the work and guidance on the scientific writing have been very valuable. Publications III and IV were written in collaboration with Dr. Kai Zenger and Mr. Jussi Koppinen, and I would like to thank them for good cooperation. I would also like to thank all my former and present colleagues and the department staff for the interesting conversations, help with the research topic and experimental measurements, and a great working environment. I am also grateful to the people working for ABB who provided valuable comments and discussion on this work in the research project meetings.

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Espoo, November 21, 2016,

Jarno Kukkola
Preface

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This thesis consists of an overview and of the following publications which are referred to in the text by their Roman numerals.


V Jarno Kukkola and Marko Hinkkanen. State observer for grid-voltage...


Author’s Contribution

Publication I: “Observer-based state-space current control for a three-phase grid-connected converter equipped with an LCL filter”

The author wrote the paper under the guidance of Prof. Hinkkanen.

Publication II: “State observer for grid-voltage sensorless control of a grid-connected converter equipped with an LCL filter”

The author wrote the paper under the guidance of Prof. Hinkkanen.


The author wrote the paper under the guidance of Prof. Hinkkanen. Dr. Zenger contributed by commenting on the manuscript.

Publication IV: “Parameter estimation of an LCL filter for control of grid converters”

The author performed the measurements and participated in the writing of the paper.

Publication V: “State observer for grid-voltage sensorless control of a converter equipped with an LCL filter: Direct discrete-time design”

The author wrote the paper under the guidance of Prof. Hinkkanen.
Publication VI: “Method for DC-link capacitance identification in voltage-source converters”

The author wrote the paper under the guidance of Prof. Hinkkanen.

Publication VII: “Grid-voltage sensorless control of a converter under unbalanced conditions: On the design of a state observer”

The author wrote the paper under the guidance of Prof. Hinkkanen.
Symbols and Abbreviations

Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>A, Bc, Bg, Cc</td>
<td>matrices of state-space presentation</td>
</tr>
<tr>
<td>C_{dc}</td>
<td>DC-link capacitance</td>
</tr>
<tr>
<td>C_{f}</td>
<td>LCL-filter capacitance</td>
</tr>
<tr>
<td>i_{c}</td>
<td>converter-current vector</td>
</tr>
<tr>
<td>i_{ca}, i_{cb}, i_{cc}</td>
<td>converter phase currents</td>
</tr>
<tr>
<td>i_{cd}, i_{cq}</td>
<td>real and imaginary components of ( i_{c} )</td>
</tr>
<tr>
<td>i_{g}</td>
<td>grid-current vector</td>
</tr>
<tr>
<td>i_{ga}, i_{gb}, i_{gc}</td>
<td>grid phase currents</td>
</tr>
<tr>
<td>j</td>
<td>imaginary unit</td>
</tr>
<tr>
<td>k</td>
<td>discrete-time index</td>
</tr>
<tr>
<td>K</td>
<td>controller gain vector</td>
</tr>
<tr>
<td>k_{1}, k_{2}, k_{3}, k_{4}, k_{1}, k_{w}</td>
<td>controller gains</td>
</tr>
<tr>
<td>k_{i,u}, k_{i,ω}, k_{p,u}, k_{p,ω}</td>
<td>gains of PI-adaptation mechanisms</td>
</tr>
<tr>
<td>K_{o}</td>
<td>observer gain vector</td>
</tr>
<tr>
<td>k_{o1}, k_{o2}, k_{o3}, k_{o4}</td>
<td>elements of the observer gain vector</td>
</tr>
<tr>
<td>L_{f}</td>
<td>inductance of an L filter</td>
</tr>
<tr>
<td>L_{fc}</td>
<td>converter-side inductance of an LCL filter</td>
</tr>
<tr>
<td>L_{fg}</td>
<td>grid-side inductance of an LCL filter</td>
</tr>
<tr>
<td>L_{t} = L_{fc} + L_{fg}</td>
<td>total inductance</td>
</tr>
<tr>
<td>p_{c}</td>
<td>active power of the converter</td>
</tr>
<tr>
<td>p_{dc}</td>
<td>external DC-link power</td>
</tr>
<tr>
<td>p_{g}</td>
<td>active power at the PCC</td>
</tr>
<tr>
<td>p_{ref}</td>
<td>active-power reference</td>
</tr>
<tr>
<td>p_{sw}</td>
<td>semiconductor-side DC-link power</td>
</tr>
</tbody>
</table>
Symbols and Abbreviations

$q_g$ reactive power at the PCC
$q_{\text{ref}}$ reactive-power reference
$R_f$ resistance of an L filter
$s$ Laplace variable
$t$ time
$T_d$ time delay
$T_s$ sampling period
$u_c$ converter-voltage vector
$u'_{c,\text{ref}}$ control-voltage vector
$u_{\text{dc}}$ DC-link voltage
$u_f$ LCL-filter capacitor-voltage vector
$u_g$ grid-voltage vector
$u_g$ magnitude of the grid-voltage vector
$u_{g_a}, u_{g_b}, u_{g_c}$ grid phase voltages
$W$ DC-link capacitor energy
$x$ state vector
$x_I$ integral state

$\gamma = e^{-j\omega_g T_s}$ complex auxiliary variable
$\Gamma_c, \Gamma_g$ input vectors
$\vartheta_g$ grid-voltage angle
$\Phi$ system matrix
$\phi_a, \phi_b, \phi_c$ phase shifts of the grid voltages
$\omega_g$ angular frequency of the grid voltage
$\omega_p$ angular resonance frequency of an LCL filter

Complex symbols are marked with boldface. Stationary-reference-frame vectors and matrices are marked with the superscript $s$. Estimated variables are marked with a hat. Estimation-error quantities are marked with a tilde. Positive-sequence components are marked with the subscript $+$, and negative-sequence components with the subscript $-$. Reference values are marked with the subscript $\text{ref}$, and limited reference values are also marked with overline.
### Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>AC</td>
<td>alternating current</td>
</tr>
<tr>
<td>DC</td>
<td>direct current</td>
</tr>
<tr>
<td>DPC</td>
<td>direct power control</td>
</tr>
<tr>
<td>DTC</td>
<td>direct torque control</td>
</tr>
<tr>
<td>LCL</td>
<td>inductor-capacitor-inductor (filter)</td>
</tr>
<tr>
<td>PCC</td>
<td>point of common coupling</td>
</tr>
<tr>
<td>PI</td>
<td>proportional integral</td>
</tr>
<tr>
<td>PLL</td>
<td>phase-locked loop</td>
</tr>
<tr>
<td>PR</td>
<td>proportional resonant</td>
</tr>
<tr>
<td>PWM</td>
<td>pulse-width modulation</td>
</tr>
<tr>
<td>ROGI</td>
<td>reduced-order generalized integrator</td>
</tr>
<tr>
<td>ZOH</td>
<td>zero-order hold</td>
</tr>
</tbody>
</table>
1. Introduction

1.1 Background

Grid-connected power electronic converters are widely applied in energy production and consumption. In energy production, they are used as interfaces between energy sources and the grid. For example, in solar power generation, these converters convert the direct current (DC) produced by solar cells to alternating current (AC) that is supplied to the AC grid. In energy consumption, an application for the grid-connected converter is the active-front-end rectifier of an electric motor drive.

Energy flow through the grid-connected converter is controlled by software algorithms. The structure and functions of the converter control system depend on the application. Typically, there are faster (inner) control loops for current or power control and slower (outer) loops, for instance, for DC voltage control. Nonetheless, a common requirement for all grid-connected converters is to be synchronized with the grid. Proper synchronization is needed in order to enable instantaneous power control. Conventionally, the synchronization has been obtained by measuring the grid voltage and using this information as an input of a phase-locked loop (PLL) or frequency-locked loop. An alternative for these methods is grid-voltage sensorless operation, where the voltage is not measured. The synchronization can be based on an estimated grid voltage or equivalent virtual flux.

The grid-voltage sensorless operation or control has been motivated by improved immunity to electrical noise (Noguchi et al., 1998) and lower costs (Hansen et al., 2000) due to the reduced number of measurement sensors in the system. Furthermore, a grid-voltage sensorless control mode can be implemented in parallel with the conventional operation pos-
sensing the sensors. This increases the reliability of a converter, since the sensorless control can continue operation, regardless of whether there are failures in the voltage measurement sensors, wires, or interfaces. In the case of such an event, these failures do not necessarily mean the end of production, for example, in renewable energy generation. Alternatively, the available grid-voltage estimate may enable the use of low-cost sensors in some applications. A potential application could be a solar inverter of which investment costs are predicted to fall significantly, even to a fifth of current prices by 2050 (Fraunhofer ISE, 2015).

Recently, an LCL filter between the converter and the grid has become a popular option for filtering out the switching harmonics. A factor behind this trend is the higher switching harmonics attenuation of the LCL filter in comparison with the conventional L filter. This enables a physically smaller and cheaper filter at low switching frequencies such as 4...8 kHz (Jalili and Bernet, 2009). However, a disadvantage of the LCL filter is its resonating behavior. This has to be considered when designing control methods for converters equipped with an LCL filter, especially if the resonance is not damped using an additional damping resistor.

The control methods for converters equipped with an LCL filter are similar to the methods developed for converters equipped with an L filter, but in the case of the LCL filter, additional functions have been included in the control software in order to damp the LCL-filter resonance. These functions are also known as active damping methods (for example, Blasko and Kaura, 1997; Dahono, 2002; Dannehl et al., 2011). For dealing with the dynamics of the LCL filter, including the resonance, a comprehensive approach is to model the dynamics in the state-space form and use this model in state-space current control (Wu and Lehn, 2006; Bolsens et al., 2006; Dannehl et al., 2010; Huerta et al., 2012; Xue et al., 2012). The state-space control offers a flexible framework to tune the closed-loop dynamics, but a drawback of this approach is that the states must be known. In the case of the LCL filter, the measurable states are typically currents and voltages. Instead of measuring these states, they can be estimated using a state observer.

The state observers and state-space current control can also be used in the grid-voltage sensorless control system. Moreover, grid-voltage estimation can be combined with the LCL-filter state estimation in a common estimation unit, as demonstrated by Bolsens et al. (2006), Mariéthoz and Morari (2009), and Ahmed et al. (2009). However, this area is still open to
research. A research challenge is to design the estimation unit such that it does not require a trial-and-error process in the tuning stage. Another interesting research challenge in this area is unbalanced grid voltage conditions, which may occur during grid faults.

1.2 Objective and Outline of the Thesis

The objective of this thesis is to develop state and grid-voltage estimation methods for grid-voltage sensorless control of grid-connected converters equipped with an LCL filter. The estimation methods should have the following properties: they use only the converter current and DC-link voltage measurements such that the methods could be implemented in a standard converter module used, for instance, in active-front-end rectifiers without any additional measurement hardware; the methods are model-based and built in such a way that they can be automatically tuned, if the main parameters of an LCL-filter model are known. Since the grid connection is assumed in this thesis, the methods for stand-alone, grid-forming, operation are beyond the scope of this thesis.

The thesis consists of this summary and seven publications. A mathematical model for the grid-connected converter equipped with an LCL filter is described in Chapter 2. Chapter 3 reviews grid-voltage estimation methods proposed in the literature. Chapter 4 presents the control functions and the experimental test setup used in the publications. The summaries of the publications and the contributions of this thesis are presented in Chapter 5. Chapter 6 concludes the thesis.
2. System Model

2.1 Space Vectors

Complex space vectors are applied to the modeling of the grid converter system. Complex symbols are marked in boldface. For example, the space vector of the converter current is $\mathbf{i}_c = i_{ca} + j i_{cb}$, where $i_{ca}$ and $i_{cb}$ are the components of the vector. The superscript $s$ denotes the stationary reference frame. With amplitude-invariant scaling, the space vector is defined as

$$\mathbf{i}_c^s = \frac{2}{3} \left( i_{ca} + e^{j2\pi/3} i_{cb} + e^{j4\pi/3} i_{cc} \right)$$

(2.1)

where $i_{ca}$, $i_{cb}$, and $i_{cc}$ are the instantaneous phase currents. It is assumed that there is no path (no neutral connection) for zero-sequence currents in the grid converter system; in other words, the sum of the phase currents is zero. Therefore, the zero-sequence component is neglected and space-vector models are used to describe the behavior of the system.

The space vectors are transformed between different reference frames. For example, the transformation from the stationary reference frame to a synchronous reference frame is given by

$$\mathbf{i}_c = e^{-j\vartheta} \mathbf{i}_c^s$$

(2.2)

where $\vartheta$ is the angle of the rotating reference frame with respect to the stationary reference frame. The space-vector components in the synchronous reference frame are denoted by subscripts $d$ and $q$ instead of $\alpha$ and $\beta$. 
System Model

\[ u_{dc} \]

\[ L_{fc} \]

\[ C_f \]

\[ u_{ga} \]

\[ u_{gb} \]

\[ i_{ca} \]

\[ i_{cb} \]

\[ u_{gc} \]

\[ i_{ga} \]

\[ i_{gb} \]

\[ i_{gc} \]

\[ (a) \]

\[ u_s^c \]

\[ u_s^g \]

\[ L_{fg} \]

\[ C_f \]

\[ i_s^c \]

\[ i_s^g \]

\[ (b) \]

\[ x_s^d \]

\[ A_s \]

\[ B_c \]

\[ B_g \]

Figure 2.1. (a) Converter equipped with an LCL filter. (b) Space-vector circuit model of the LCL filter between the converter and the grid.

2.2 Converter

The three-phase circuit of the grid-connected converter equipped with an LCL filter is shown in Figure 2.1(a). The converter AC voltage is pulse-width modulated. For the control design, switching-cycle-averaged converter voltage is assumed. The converter voltage vector is \( u_c^e \). The converter voltage can be calculated from its reference or from the switching signals and DC-link voltage.

2.3 LCL Filter

The converter-side inductances \( L_{fc} \), capacitances \( C_f \), and grid-side inductances \( L_{fg} \) are assumed to be symmetrical. The space-vector model of the LCL filter is shown in Figure 2.1(b). The grid-voltage vector is \( u_g^e \). Damping resistors in the LCL filter (Peña-Alzola et al., 2013; Beres et al., 2016) are not used in this thesis. The dynamics of the LCL filter are written in the state-space form as

\[
\frac{dx^s}{dt} = \begin{bmatrix}
0 & -\frac{1}{L_{fc}} & 0 \\
\frac{1}{C_f} & 0 & -\frac{1}{C_f} \\
0 & \frac{1}{L_{fg}} & 0
\end{bmatrix} x^s + \begin{bmatrix}
\frac{1}{L_{fc}} \\
0 \\
0
\end{bmatrix} u_c^e + \begin{bmatrix}
0 \\
0 \\
-\frac{1}{L_{fg}}
\end{bmatrix} u_g^e
\]

(2.3)

where \( x^s = [i_c^e, i_f^e, i_g^e]^T \) is the state vector, \( A_s \) is the system matrix, and \( B_c \) and \( B_g \) are the input vectors for the converter and grid voltages, respectively. The states are: converter-current vector \( i_c^e \), capacitor-voltage vector \( u_f^e \), and grid-current vector \( i_g^e \). The transfer function from the converter voltage to the converter current is

\[
Y^s(s) = \frac{i_c^e(s)}{u_c^e(s)} = C_c(sI - A_s)^{-1}B_c = \frac{s^2 + \omega_d^2}{L_{fc} s(s^2 + \omega_d^2)}
\]

(2.4)
where $C_c = [1, 0, 0]$ and

$$
\omega_p = \sqrt{\frac{L_{LC} + L_{fg}}{L_{LC} L_{fg} C_f}} \quad \text{and} \quad \omega_z = \sqrt{\frac{1}{L_{fg} C_f}} \quad (2.5)
$$

are the resonance frequency and the anti-resonance frequency of the LCL filter, respectively.

Generally, series resistances or parasitic elements of the circuit components could be included in the LCL filter model in order to improve modeling accuracy. In this thesis, neither these properties nor nonlinearities of the components are included in the model. From a feedback control point of view, the accuracy of the presented model is sufficient as indicated, for example, in Publications III and V.

In the publications, control and estimation algorithms operate in a synchronous reference frame. The reference frame is tied to the actual grid-voltage vector in Publications I, III, and IV or the estimated grid-voltage vector in Publications II and V. In unbalanced grid voltage conditions, the coordinate system is tied to the estimated positive-sequence component of the grid-voltage vector in Publications V and VII.

In the synchronous reference frame, which is fixed to the actual grid-voltage vector, the converter-current dynamics are

$$
\frac{dx}{dt} = \begin{bmatrix}
-\frac{j \omega_g}{L_{LC}} & -\frac{1}{L_{LC}} & 0 \\
\frac{1}{C_f} & -\frac{j \omega_g}{C_f} & -\frac{1}{C_f} \\
0 & 0 & -\frac{1}{C_{fg}}
\end{bmatrix} x + \begin{bmatrix}
\frac{1}{L_{LC}} \\
0 \\
0
\end{bmatrix} u_c + \begin{bmatrix}
0 \\
0 \\
-\frac{1}{C_{fg}}
\end{bmatrix} u_g
$$

$$
i_c = \begin{bmatrix}
1 \\
0 \\
0
\end{bmatrix} x
$$

where $\omega_g$ is the angular frequency of the grid-voltage vector and $C_c$ is the output vector for the converter current. In the coordinate system fixed to the estimated grid-voltage vector, the corresponding model is obtained replacing $\omega_g$ with $\hat{\omega}_g$ in (2.6).

### 2.4 Grid

Figure 2.2 shows a grid model. The converter is connected to the grid at the point of common coupling (PCC). Generally, the grid behind the PCC is different in different geographical locations, and the grid impedance is frequency and time dependent (Jessen et al., 2015). Therefore, exact modeling of the grid is challenging. From a control design point of view, the
grid can be modeled as a voltage source or a voltage source $e_g^s$ with a series connected impedance $Z_g$. Even though these models are approximations, they are practical in control design for their simplicity.

In this thesis, the control and estimation algorithms are first designed assuming $Z_g = 0$ and $u_g^s = e_g^s$. When the designed algorithms are validated, the robustness against different conditions $Z_g \neq 0$ is examined. In Publications I, III, and V, simulations and numerical analysis were used in the validation.

The phase voltages of the grid are

$$
\begin{align*}
    u_{ga} &= \hat{u}_{ga} \cos(\omega_g t + \phi_a) \\
    u_{gb} &= \hat{u}_{gb} \cos(\omega_g t - 2\pi/3 + \phi_b) \\
    u_{gc} &= \hat{u}_{gc} \cos(\omega_g t - 4\pi/3 + \phi_c)
\end{align*}
$$

(2.7)

where $\omega_g$ is the angular frequency of the voltages, $\phi_a$, $\phi_b$, and $\phi_c$ are the phase shifts and peak values of the voltages are marked with a hat. Ideally, the phase voltages are balanced, which means $\phi_a = \phi_b = \phi_c$ and $\hat{u}_{ga} = \hat{u}_{gb} = \hat{u}_{gc} = u_g$, where $u_g$ is the magnitude of the grid voltages. When the a-phase is selected as the reference, meaning $\phi_a = 0$, the corresponding grid-voltage vector in the balanced conditions is

$$
\mathbf{u}_g^s = e^{j\vartheta_g} u_g
$$

(2.8)

where the angle is

$$
\vartheta_g = \int \omega_g dt
$$

(2.9)

Moreover, the magnitude $|\mathbf{u}_g^s| = u_g$ and the rotation speed $d\vartheta_g/dt = \omega_g$ of the voltage vector are constant in the balanced conditions.

When the converter is operating under a grid fault, the phase voltages of the grid may be unbalanced, which generally means that the magnitudes $\hat{u}_{ga}$, $\hat{u}_{gb}$, and $\hat{u}_{gc}$ or the phase angles $\phi_a$, $\phi_b$, and $\phi_c$ are unequal. In the unbalanced grid conditions, the magnitude $|\mathbf{u}_g^s|$ and the rotation speed $d\vartheta_g/dt$ of the vector are no longer constant, but they are pulsating (Teodorescu et al., 2011). However, in these conditions, the grid-voltage vector can be written using positive and negative-sequence voltage components ($u_{g+}^s$).
and $u^s_{g-}$ as follows

$$
\mathbf{u}^s_g = \mathbf{u}^s_{g+} + \mathbf{u}^s_{g-} = e^{j\varphi_{g+}} u_{g+} + e^{j\varphi_{g-}} u_{g-}
$$

(2.10)

where $u_{g+}$ and $u_{g-}$ are the magnitudes and $\varphi_{g+}$ and $\varphi_{g-}$ are the angles of the positive and negative-sequence components, respectively. In steady-state conditions, the magnitudes $u_{g+}$ and $u_{g-}$ are constant, although $|u^s_g|$ is pulsating. The positive-sequence angle is

$$
\varphi_{g+} = \int \omega_{g+} dt = \int \omega_g dt
$$

(2.11)

where the rotation speed $\omega_{g+}$ of the positive-sequence component equals the angular frequency $\omega_g$ of the phase voltages. The negative-sequence component rotates in the opposite direction, and its angle is $\varphi_{g-} = -\varphi_{g+} + \phi_{g-}$, where $\phi_{g-}$ is the phase shift with respect to the positive-sequence component.

In addition to the unbalance, the real grid voltages may contain some harmonics at frequencies multiples of the fundamental frequency (EN 50160, 2010). Positive and negative-sequence harmonics can be modeled with space vectors in a similar way to the fundamental frequency components, and the superposition principle can be applied to build a vector for distorted grid voltages.

2.5 DC link

The DC link of the converter is modeled as an ideal capacitor. The equivalent circuit of the model is shown in Figure 2.3, which also illustrates the power flow through the converter. The external power flowing into the DC link is $p_{dc} = u_{dc} i_{dc}$, where $u_{dc}$ is the DC-link voltage and $i_{dc}$ is the external direct current. The power and current on the DC side of the semiconductor bridge are $p_{sw}$ and $i_{sw}$, respectively. The dynamics of the DC-link voltage are

$$
C_{dc} \frac{du_{dc}}{dt} = i_{dc} - i_{sw}
$$

(2.12)

where $C_{dc}$ is the DC-link capacitance. Multiplying both sides of (2.12) by $u_{dc}$ and assuming a constant capacitance, the capacitor energy dynamics
are obtained
\[
\frac{d}{dt} \left( \frac{1}{2} C_{dc} u_{dc}^2 \right) = p_{dc} - p_{sw}
\] (2.13)

where
\[
W = \frac{1}{2} C_{dc} u_{dc}^2
\] (2.14)
is the energy stored in the capacitor. Although there are some losses in the converter, the power \(p_{sw}\) approximately equals the converter power \(p_c\).

2.6 Grid-Side Powers

The active and reactive powers at the PCC are of interest. The complex power at the PCC is \(s_g = 3/2 \cdot u_s^g \cdot i_s^g\), where \(^*\) denotes the complex conjugate. The active and reactive powers are
\[
p_g = \frac{3}{2} \text{Re}\{u_s^g i_s^g^*\} = \frac{3}{2} \text{Re}\{u_g i_g^*\}
\]
\[
q_g = \frac{3}{2} \text{Im}\{u_s^g i_s^g^*\} = \frac{3}{2} \text{Im}\{u_g i_g^*\}
\] (2.15)

Other powers are similarly calculated.

2.7 Discrete-Time Model

For direct digital design of control and estimation algorithms, the LCL filter is modeled in the discrete-time domain as presented in Publication III. The zero-order hold (ZOH) is assumed for the converter voltage \(u_c^c\) in the stationary reference frame in order to model the hold related to pulse-width modulation (PWM). The assumption means that \(u_c^c\) remains constant between the sampling instants. The sampling is synchronized with the PWM. For the grid voltage \(u_g\), the ZOH is assumed in the synchronous reference frame, where \(u_g\) is a DC quantity in balanced conditions. A hold-equivalent discrete-time model for (2.6) is
\[
x(k + 1) = \Phi x(k) + \Gamma_c u_c(k) + \Gamma_g u_g(k)
\]
\[
i_c(k) = C_c x(k)
\] (2.16)
where \( k \) is the discrete-time index. In the model, the system matrices are

\[
\Phi = e^{A T_s} = \gamma \begin{bmatrix}
L_{tg} + L_{tg} \cos(\omega_p T_s) & -\sin(\omega_p T_s) \\
\omega_p L_{tg} & \omega_p L_{tg} \\
L_{tg} \frac{\sin(\omega_p T_s)}{\omega_p L_{tg}} & \cos(\omega_p T_s) \\
\omega_p L_{tg} & \omega_p L_{tg} \\
L_{tg} \frac{\sin(\omega_p T_s)}{\omega_p L_{tg}} & \frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg}} \\
\frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg}} & \frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg}}
\end{bmatrix} \\
\begin{bmatrix}
-L_{tg} \sin(\omega_p T_s) \\
\omega_p L_{tg} \\
L_{tg} \frac{\sin(\omega_p T_s)}{\omega_p L_{tg}} \\
\omega_p L_{tg} \\
L_{tg} \frac{\sin(\omega_p T_s)}{\omega_p L_{tg}} \\
\frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg}}
\end{bmatrix}
\]

\( \Gamma_c = \left( \int_0^{T_s} e^{A \tau} e^{-j \omega_g (T_s - \tau)} d\tau \right) B_c = \gamma \begin{bmatrix}
\frac{T_s}{L_t} + \frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} & \frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{T_s}{L_t} - \frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{T_s}{L_t} + \frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{L_{tg} \sin(\omega_p T_s)}{\omega_p L_{tg} L_t}
\end{bmatrix} \gamma
\]

\( \Gamma_g = \left( \int_0^{T_s} e^{A \tau} d\tau \right) B_g = \gamma \begin{bmatrix}
\frac{T_s}{L_t} + \frac{\omega_p \sin(\omega_p T_s) + \omega_p^2 \cos(\omega_p T_s) - \delta - e^{j \omega_g T_s} \omega_p^2}{\omega_p L_{tg} L_t} \\
\frac{\omega_p \sin(\omega_p T_s) + \omega_p^2 \cos(\omega_p T_s) - \delta - e^{j \omega_g T_s} \omega_p^2}{\omega_p L_{tg} L_t} \\
\frac{T_s}{L_t} - \frac{\omega_p \sin(\omega_p T_s) + \omega_p^2 \cos(\omega_p T_s) - \delta - e^{j \omega_g T_s} \omega_p^2}{\omega_p L_{tg} L_t} \\
\frac{\omega_p \sin(\omega_p T_s) + \omega_p^2 \cos(\omega_p T_s) - \delta - e^{j \omega_g T_s} \omega_p^2}{\omega_p L_{tg} L_t}
\end{bmatrix} \gamma
\]

where \( T_s \) is the sampling period, \( \gamma = e^{-j \omega_g T_s}, \delta = \omega_p^2 - \omega^2, \) and \( L_t = L_{tg} + L_{tg} \).

Generally, a hold-equivalent discrete-time model can be derived starting from the solution \( x(t) = f(x(0), t, u_c, u_g) \) to the system of differential equations (2.6), which is an integral equation (Franklin et al., 1997). In the literature, different ZOH-equivalent discrete-time models have been presented, for example, by Wu and Lehn (2006), Mariéthoz and Morari (2009), Dannehl et al. (2010), Xue et al. (2012), Huerta et al. (2012), Nishida et al. (2014), and Miskovic et al. (2014). The different models originate from different assumptions used in the modeling. If the ZOH for the grid voltage \( u_g^n \) is assumed in the stationary reference frame (Xue et al., 2012; Nishida et al., 2014; Miskovic et al., 2014), the input vector \( \Gamma_g \) becomes different

\[
\Gamma_g' = \left( \int_0^{T_s} e^{A \tau} e^{-j \omega_g (T_s - \tau)} d\tau \right) B_g = \gamma \begin{bmatrix}
\frac{T_s}{L_t} + \frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} & \frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{T_s}{L_t} - \frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{T_s}{L_t} + \frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t} \\
\frac{\omega_p \sin(\omega_p T_s)}{\omega_p L_{tg} L_t}
\end{bmatrix} \gamma
\]

On the other hand, if the ZOH for the converter voltage is assumed in the synchronous reference frame as, for instance, by Wu and Lehn (2006), and Huerta et al. (2012), the input vector \( \Gamma_c \) becomes different

\[
\Gamma_c' = \left( \int_0^{T_s} e^{A \tau} d\tau \right) B_c
\]

However, the hold of the converter-voltage vector and the rotation of the grid-voltage vector have been modeled more accurately in (2.17) compared to (2.18) and (2.19). The discrete-time model (2.16) in the stationary reference frame is

\[
x(k + 1) = \Phi^c x^c(k) + \Gamma_c u_c^c(k) + \Gamma_g u_g^c(k)
\]

\[
\hat{e}_c^c(k) = C_c x^c(k)
\]
where $\Phi_s = \gamma^{-1} \Phi_s$, $\Gamma_c^s = \gamma^{-1} \Phi_s$, $\Gamma_g^s = \gamma^{-1} \Gamma_g$.

In the grid-voltage sensorless control methods by Ahmed et al. (2009, 2011a), and Mohamed et al. (2012), the approximative Forward Euler discretization method was used to calculate a discrete-time state-space model. This means that the derivative in a continuous-time state-space model is replaced with the approximation $[x(k+1) - x(k)]/T_s$ (Franklin et al., 1997). In this case, simpler discrete-time models are obtained, but the accuracy of the models is reduced in comparison with the hold-equivalent models.
3. Grid-Voltage Estimation

This chapter reviews grid-voltage estimation methods proposed in earlier studies for grid-voltage sensorless control. Since the estimation methods operate within the framework of sensorless control, the framework is first briefly introduced. The basic grid-voltage estimation principles have been originally introduced in the case of the grid-connected converter equipped with a conventional L filter. Therefore, in order to provide background, the methods developed for this case are briefly discussed, followed by a review of the estimation methods developed for the converters equipped with an LCL filter, which is the scope of this thesis. Finally, the proposed estimation methods are described.

3.1 Grid-Voltage Estimation as a Part of the Control System

Since the 1990s, different control methods have been proposed in order to enable the converter to operate without grid-voltage, or more generally, AC-side voltage sensors (for example, Takeshita et al., 1994; Ito et al., 1994; Manninen, 1995; Noguchi et al., 1998; Ohnuki et al., 1999); nevertheless, some other sensors are needed in the control system. Typically, these grid-voltage sensorless control methods use measurements from the DC-link voltage and converter phase currents. However, some other instrumentation solutions have been proposed. For example, the control method by Lee and Lim (2002) is based on the DC-side current and voltage measurements, while the phase currents of the converter are reconstructed from the measured direct current and switching states. In the method by Ohnuki et al. (1999), only the converter phase currents are measured and the DC-link voltage is estimated together with the grid voltage.

In this thesis, the measurements from the DC-link voltage and con-
Grid-Voltage Estimation

Figure 3.1. Control functions of a grid-voltage sensorless converter.

Converter phase currents are assumed. They are useful in overcurrent and overvoltage protection, and sensors for these measurements can be integrated inside of a converter module. Moreover, the same converter modules can also be used in motion-sensorless variable-speed drives without any hardware modifications.

Figure 3.1 shows the typical control functions of the grid-voltage sensorless converters from a voltage-oriented control point of view. The functions enable: controlled energy transfer between the DC and AC sides, the regulated DC-link voltage, and the controlled reactive power and current quality on the AC side. Although the control goals are common, the interfaces between the functions may be different. For example, in direct torque control (DTC) (Manninen, 1995) or direct power control (DPC) (Noguchi et al., 1998), a separate pulse-width modulator is not used. Furthermore, in virtual-flux based control methods, an equivalent flux of the grid voltage (Malinowski et al., 2001) or converter voltage (Pöllänen et al., 2003) is estimated providing the reference frame for the power control.

In addition to the functions presented in Figure 3.1, the control system may include other application-dependent functions and receive inputs from other parts of the converter software or from a user. Additional functions can be, amongst others, identification algorithms for circuit parameters of the converter system as presented in Publications IV and VI.

A grid-voltage estimator operates as a part of a grid-voltage sensorless control system. A typical structure of the sensorless control system is illustrated in Figure 3.2. It has cascaded controllers for the DC-link voltage and the converter-current vector. The reference $i_{c,\text{ref}}$ for current control is calculated from the power references $p_{\text{ref}}$ and $q_{\text{ref}}$, using either the nominal grid-voltage magnitude and frequency or the estimated values $\hat{u}_g$ and
Figure 3.2. Grid-voltage sensorless control system, where the control functions are implemented in the estimated grid-voltage reference frame.

\( \hat{\omega}_g \). The active-power reference \( p_{\text{ref}} \) is obtained from the voltage controller, which regulates the DC-link voltage to the given value \( u_{\text{dc},\text{ref}} \). In Figure 3.2, the control and grid-voltage estimation algorithms operate in the synchronous reference frame that is fixed to the estimated grid-voltage angle \( \hat{\vartheta}_g \). Alternatively, the control and estimation algorithms can be implemented in the stationary (\( \alpha\beta \)) reference frame (Suul et al., 2012a). Furthermore, instead of space vectors, phase quantities (abc reference frame) have been used in the grid-voltage estimation and control, for instance, by Mariéthoz and Morari (2009) and Ahmed et al. (2011a).

The sensorless control system used in this thesis follows the structure presented in Figure 3.2. While the state and grid-voltage estimation methods are presented in this chapter, the other functions used in this thesis are described in more detail in Chapter 4.

### 3.2 Introduction to Grid-Voltage Estimation Methods

#### 3.2.1 Inductor Voltage Equation

The first grid-voltage estimation methods were proposed for grid-connected converters equipped with a conventional L filter that is ideally a pure inductance \( L_f \) between the converter and the grid. The voltage equation of this filter is

\[
L_i \frac{d q^s_c}{dt} = -R_i q^s_c + u^s_c - u^s_g
\]
where $R_t$ represents the series resistance of the inductor and is often neglected. From (3.1), a grid-voltage estimate can be directly calculated, if the converter voltage and currents are known (Ohnuki et al., 1999; Kennel et al., 2003). A practical drawback of this approach is the differentiation of the current, which is needed for calculation of the inductor voltage. The differentiation amplifies the possible measurement noise in the current as reported by Hansen et al. (2000).

### 3.2.2 Power-Based Estimation

For the converters equipped with an L filter, a grid-voltage estimation method based on instantaneous power calculation was proposed by Noguchi et al. (1998). First, they estimated the grid-side powers $\hat{p}_g$ and $\hat{q}_g$, which are functions of the inductance, converter currents, their derivatives, switching states, and DC-link voltage. Then, a grid-voltage estimate was calculated from the estimated powers and measured converter currents using the instantaneous power theory. The method of Noguchi et al. enables estimation of the harmonic components of the grid voltage together with the fundamental component, but the current derivatives are needed in the calculations, which potentially causes noise in the estimated quantities.

In the DPC and grid-voltage estimation method by Noguchi et al. (1998), the currents are sampled and the powers are estimated several times during a switching state. This is avoided in voltage-oriented control using a space vector modulation (Hansen et al., 2000), where the currents are sampled in the zero-vector states. Moreover, Hansen et al. applied the power-based voltage estimation principle by Noguchi et al. to the inductor voltage with the grid-voltage estimate being calculated as a sum of the converter and inductor voltages. To reduce the noise due to the differentiation, Hansen et al. proposed a simple low-pass filter.

### 3.2.3 Virtual-Flux Models

Mathematically, a grid-connected converter equipped with an L filter and an electric motor drive have many similarities. The inductors combined with the grid-voltage source can be treated as a virtual AC motor seen from the converter terminals (Duarte et al., 1999). The virtual motor has a virtual flux, and the virtual-flux approach has been applied in grid-voltage sensorless control by various authors, such as Manninen (1995),
Malinowski et al. (2001), Pöllänen (2003), and Suul (2012). In this approach, the grid voltage is indirectly estimated through a virtual flux of the grid-voltage vector. The virtual flux is defined as an integral of the voltage, in other words, the equivalent flux of the grid voltage can be directly calculated from the integral of (3.1).

In the virtual-flux estimation, current derivatives are not needed. This decreases sensitivity against noise (Malinowski et al., 2001). However, the integrator cumulates even a small DC offset in the integrated signals causing drift, which is a problem in the practical use of the integrator-based virtual-flux estimator. Different alternatives have been proposed for the pure integrator in the virtual-flux estimation. A simple alternative is a low-pass filter (Malinowski, 2001; Pöllänen, 2003), with several other methods having been reviewed by Pöllänen (2003) and Suul (2012).

### 3.2.4 Control-System Dependent Methods

Instead of building a separate estimator for the grid voltage, the proportional integral (PI) controllers typically used in control of the converter current and DC-link voltage have been shown to provide intermediate variables for grid-voltage frequency and phase estimation (Ohnishi and Fujii, 1997; Barrass and Cade, 1999; Kwon et al., 1999; Agirman and Blasko, 2003). For example, an angle error signal for the frequency and angle estimation has been obtained from the control error of the reactive-power-producing current component (q-component) (Kwon et al., 1999), or its integral in a current controller (Agirman and Blasko, 2003). If estimators depend on the control functions, their general applicability may be limited. However, in converter current control, integrators are generally used to cancel out the steady-state errors caused by disturbances. Since the grid voltage is a disturbance for the control system, the output signals of these integrators can be applied to the disturbance (grid voltage) estimation, as also reported by Liserre et al. (2006).

### 3.2.5 State Observers

“Missing state-variable information, not available for measurement, can be suitably approximated by an observer" as concluded by Luenberger (1971). A simple state observer could be a copy of a system model, such as (2.6) or (3.1). However, the grid voltage would be an input to this observer, which means that the voltage should be measured or estimated.
Different observers that combine the state estimation of an L or LCL filter model with the grid-voltage estimation have been proposed in the literature. These observers could be classified in various ways, but in this thesis, they are divided into two groups. Here, the first group is called observers augmented with a disturbance model, which includes the observers presented, for example, by Song et al. (2003) and Bolsens et al. (2006). Typically, a disturbance model describing grid-voltage dynamics is included in the model matrices of these observers. The second group is here called adaptive observers, which includes the observers presented, for instance, by Takeshita et al. (1994), and Mariéthoz and Morari (2009), as well as the observers proposed in Publications II and V. In this group, parameters or variables of the observer are updated based on estimation or adaptation laws that take an estimation error signal as an input.

The general structures of these observers are illustrated in Figure 3.3. Depending on the selection of the coordinate system (abc, αβ, dq) and estimated variables, different implementations for these observers have been proposed. Observers for the converters equipped an L filter are reviewed in Sections 3.2.6 and 3.2.7. The observers for the converters equipped with an LCL filter are reviewed in Section 3.3.2 and the observers developed in this thesis are presented in Sections 3.4, 3.5, and 3.7.

### 3.2.6 Observers Augmented with a Disturbance Model

The grid voltage can be considered as a disturbance in the control system. When the constant magnitude \( u_g \) (balanced voltages) is assumed, a dynamic model for the grid voltage is

\[
\frac{d}{dt} u^s_g = \frac{d}{dt} \left( e^{j\varphi_g} u_g \right) = j\omega_g u^s_g
\]
Moreover, if the unbalanced conditions are considered and the voltage is modeled as in (2.10), the dynamics can be expressed as (Song et al., 2003)

\[
\frac{d}{dt} u_s^g = j\omega_g u_s^g+ - j\omega_g u_s^g
\]  

(3.3)

Furthermore, Song et al. (2003) proposed that the grid voltage can be estimated using a state observer that is based on the system model (3.1) combined with the disturbance model (3.3). The observer is

\[
\frac{d}{dt} \begin{bmatrix}
\hat{i}_s^c \\
\hat{u}_s^g \\
\end{bmatrix} = \begin{bmatrix}
-R_f & -1/L_f \\
-j\omega_g & 0 \\
0 & -j\omega_g \\
\end{bmatrix} \begin{bmatrix}
\hat{i}_s^c \\
\hat{u}_s^g \\
\end{bmatrix} + \begin{bmatrix}
1/L_f \\
0 \\
0 \\
\end{bmatrix} u_c + K_o(i_c - \hat{i}_c)
\]  

(3.4)

where \(K_o\) is the gain and the estimated values are marked with a hat. The observer enables simultaneous estimation of the positive and negative-sequence components of the voltage. However, the grid-voltage angular frequency \(\omega_g\) is a parameter needed for the disturbance model; hence, the method of Song et al. is not directly frequency-adaptive and assumes constant frequency. In addition to the positive and negative-sequence components, the disturbance observer structure can be augmented for the grid-voltage harmonics (Lee et al., 2009).

A different grid-voltage estimator based on disturbance estimation was proposed by Ito et al. (1994). They built an observer for the converter-current component caused by the disturbances, including the grid voltage, and an estimate for the grid voltage was obtained scaling the estimated current component with a constant gain.

### 3.2.7 Adaptive Observers

An observer for the converter current can be directly derived from (3.1). Let the observer be implemented in the synchronous reference frame, whose real (d) axis is aligned with the estimated grid voltage, i.e., \(u_g = \hat{u}_g + j0\), where \(\hat{u}_g\) is the magnitude of the voltage (\(\hat{u}_g = \hat{u}_s^g = e^{j\hat{\varphi}_g}\)). The observer is

\[
\frac{d\hat{i}_c}{dt} = - \left( \frac{R_f}{L_f} + j\omega_g \right) \hat{i}_c + \frac{1}{L_f} u_c - \frac{1}{L_f} \hat{u}_g + k_o(i_c - \hat{i}_c)
\]  

(3.5)

where the estimated values are marked with a hat and \(k_o\) is the observer gain multiplying the estimation error \(\hat{i}_c = i_c - \hat{i}_c\). As the equation shows, the estimated grid-voltage magnitude, frequency, and angle are needed for the current estimation. Nevertheless, the estimation error \(\hat{i}_c\) includes information on the grid voltage.
Using an estimator similar to (3.5) and $k_o = 0$, Takeshita et al. (1994), and later Lee and Lim (2002), showed that the grid-voltage magnitude $\hat{u}_g$ and angle $\hat{\vartheta}_g$ can be estimated from the current estimation error $\tilde{i}_c$. In the method of Takeshita et al. (1994), the $d$ and $q$ components of the estimation error $\tilde{i}_c = \tilde{i}_{cd} + j\tilde{i}_{cq}$ are inputs for PI regulators that provide the estimated magnitude and angle of the grid voltage as outputs

$$
\hat{u}_g = k_{p,u} \tilde{i}_{cd} + k_{i,u} \int \tilde{i}_{cd} dt, \quad \hat{\vartheta}_g = k_{p,\theta} \tilde{i}_{cq} + k_{i,\theta} \int \tilde{i}_{cq} dt
$$

(3.6)

Takeshita et al. (1994) and Lee and Lim (2002) calculated the estimated grid-voltage angular frequency as $\hat{\omega}_g = d\hat{\vartheta}_g/dt$. A similar adaptive estimation approach was also proposed by Agirman and Blasko (2003). They presented an observer similar to (3.5), but only for the active-power-producing component of the current, and the estimate for the magnitude $\hat{u}_g$ was obtained from an integral of the estimation error of this current component. Agirman and Blasko estimated $\hat{\omega}_g$ and $\hat{\vartheta}_g$ using a separate control-system dependent method (cf. Section 3.2.4).

Later, Mohamed et al. (2007) stated that an adaptive observer for estimating the grid voltage $u_g$ in a synchronous reference frame is obtained combining an observer based on (3.5) and a steepest descent estimation algorithm. This algorithm resembles the integral action in the estimation laws by Takeshita et al. (1994) and Agirman and Blasko (2003). Moreover, an observer similar to (3.5) and a steepest descent estimation law for the grid voltage have been presented in the framework of neural networks (Mohamed and El-Saadany, 2008).

Instead of implementing an adaptive observer in the synchronous reference frame, Miskovic et al. (2014) showed that a similar performance can be obtained, if the observer (3.5) is implemented in the stationary reference frame and a proportional resonant (PR) controller is used in the grid-voltage estimation mechanism. Moreover, they showed that by adding resonant terms for harmonic frequencies (e.g., 5th, 7th, ...) in the mechanism, the observer is able to estimate the voltage harmonics superimposed on the fundamental component.

### 3.3 Estimation Methods for Converters Equipped with an LCL Filter

The grid-voltage estimation methods presented for the converters equipped with an LCL filter follow concepts similar to those that have been proposed (and reviewed in Section 3.2) for the converters equipped
Grid-Voltage Estimation

with an L filter. However, the dynamic model (2.3) of the LCL filter is more complex compared to the model (3.1) of the L filter. Therefore, the grid-voltage estimation methods for the converters equipped with an LCL filter are either more complicated, or approximations have been used in order to simplify estimation, or only the capacitor voltage \( u_c^{s} \) is estimated instead of the grid voltage \( u_g^{s} \).

### 3.3.1 Considering an LCL Filter as an L Filter

The capacitor voltage \( u_c^{s} \) can be estimated using the voltage equation for \( L_{fc} \),

\[
L_{fc} \frac{d\hat{i}_c^{s}}{dt} = u_c^{s} - u_i^{s}
\]

and applying the methods based on the inductor voltage equation (reviewed in Section 3.2). Capacitor-voltage estimation methods following this approach have been reported by Malinowski and Bernet (2008), Mohamed and El-Saadany (2011), and Miskovic et al. (2016). The method by Malinowski and Bernet (2008) is based on instantaneous power calculation following the method by Hansen et al. (2000). Alternatively, Miskovic et al. (2016) applied their previously introduced adaptive observers (Miskovic et al., 2014) for the capacitor-voltage estimation. The capacitor voltage of an LC filter (presented as LCL filter with a grid inductance) was also estimated by Mohamed and El-Saadany (2011), using an adaptive observer and the L-filter model in the observer.

Alternatively, it has been assumed that the impedance of the LCL-filter capacitor is high at low frequencies and \( i_g^{s} \approx i_c^{s} \); in other words, the LCL filter resembles cascaded inductors \( (L_{fc} + L_{fg}) \). Based on this approximation, Malinowski and Bernet (2008) estimated the voltage drop across the both inductors for calculating a grid-voltage estimate \( \hat{u}_g^{s} \). The steady-state solution for \( u_c^{s} \) at a single frequency \( \omega \) can be calculated, for example, from (2.3). The solution is

\[
\hat{u}_g^{s} = u_c^{s} - j\omega(L_{fc} + L_{fg})\hat{i}_c^{s} - \omega^2 C_f L_{fg} u_c^{s} + j\omega^3 C_f L_{fc} L_{fg} \hat{i}_c^{s}.
\]

The last two terms of the solution indicate the frequency-dependent error in \( \hat{u}_g^{s} \) related to the approximation.

### 3.3.2 Full-Order LCL-Filter Model in the Estimation

**Direct Calculation from the Dynamic Model**

It is possible to solve the grid voltage \( u_g^{s} \) directly from the dynamic LCL-filter model (2.3), if the circuit parameters, converter current, and con-
verter voltage are known. The result would include the third derivatives of the converter current (Gullvik et al., 2007; Hoff and Sułkowski, 2012). The third derivatives of the current in the estimator would make it extremely noise sensitive.

**Virtual-Flux Models**

Current derivatives can be avoided in virtual-flux models that have also been applied in the case of the converter equipped with an LCL filter (Pölätänen et al., 2003; Serpa et al., 2007; Gullvik et al., 2007; Wrona and Malon, 2014; Zhang et al., 2014; Roslan et al., 2015). Estimation of the virtual flux across the LCL-filter capacitor is analogous with the estimation of the grid virtual flux in the case of the L filter. However, in the estimation of the grid virtual flux, the grid-side current $i_{gs}$ of the LCL filter has to be known. In the method of Serpa et al. (2007), the capacitor currents are measured together with the converter currents for calculation of the grid-side currents. Instead of adding an extra set of sensors, Gullvik et al. (2007) estimate the capacitor currents using the second derivative of the virtual flux across the capacitor. Wrona and Malon (2014) proposed an estimation method for the capacitor currents using second-order generalized integrators, but the detailed explanation of the method is missing. This method has also been studied by Roslan et al. (2015). In the method by Zhang et al. (2014), the currents are estimated assuming sinusoidal waveforms and using fundamental-frequency calculations.

The estimation of the grid virtual flux was avoided by Pölätänen et al. (2003). They fixed the synchronous reference frame of the control system to the estimated converter virtual flux enabling the active and reactive power control. In principle, the approximation $i_{gs}^* \approx i_{gs}$ could also be used to achieve an approximative grid virtual flux and to avoid the problem of the grid or capacitor current estimation.

**Disturbance Observer**

In order to avoid extra measurements and without approximating $i_{gs}^* \approx i_{gs}$, the unmeasured states ($u_s$ and $i_{gs}$) can be estimated using state observers. Bolsens et al. (2006) built a state observer augmented with a disturbance model (cf. Figure 3.3(a)). They combined the LCL-filter model with a disturbance model, including the grid-voltage harmonics up to the 7th component. The resulting 14-state model was used for the state and grid-voltage estimation. Detailed characteristics, such as harmonics, of the grid voltage can be estimated with a high-order disturbance model. How-
ever, a high-order observer is difficult to tune. In the estimation, Bolsens et al. used the Kalman filter and tuned it based on trial and error.

The method by Bolsens et al. (2006) was introduced for a single-phase converter. However, the concept can be applied for three-phase converters as presented by Hoffmann et al. (2012) and Fischer et al. (2013), although their motivation was to improve the disturbance rejection of current control, rather than achieve grid-voltage sensorless operation.

**Adaptive Observes**

In the case of the LCL filter, the principle of estimating the state vector and grid voltage with an adaptive observer structure, shown in Figure 3.3(b), has been reported by Ahmed et al. (2009) and Mariéthoz and Morari (2009). In the method of Ahmed et al. (2009), a discretized system model (cf. Section 2.7) for phase quantities is first applied for the state estimation as follows

\[
\begin{bmatrix}
\hat{i}_{ca}(k+1) \\
\hat{u}_{fa}(k+1) \\
\hat{i}_{ga}(k+1)
\end{bmatrix} =
\begin{bmatrix}
1 - \frac{R_{fc}T_s}{L_{fc}} & -\frac{T_s}{L_{fc}} & 0 \\
\frac{T_s}{C_f} & 1 & -\frac{T_s}{C_f} \\
0 & \frac{T_s}{L_{fg}} & 1 - \frac{R_{fg}T_s}{L_{fg}}
\end{bmatrix}
\begin{bmatrix}
\hat{i}_{ca}(k) \\
\hat{u}_{fa}(k) \\
\hat{i}_{ga}(k)
\end{bmatrix} +
\begin{bmatrix}
\frac{T_s}{L_{fc}} & 0 \\
0 & 0 \\
0 & -\frac{T_s}{L_{fg}}
\end{bmatrix}
\begin{bmatrix}
u_{ca}(k) \\
\hat{u}_{fa}(k) \\
\hat{i}_{ga}(k)
\end{bmatrix}
\]

(3.8)

where \( \hat{i}_{ca}, \hat{u}_{fa}, \hat{i}_{ga} \), and \( \hat{u}_{ga} \) are the estimated values for the a-phase converter current, capacitor voltage, grid current and grid voltage, respectively, and \( u_{ca} \) is the a-phase converter voltage. The b and c-phase quantities are estimated similarly. Then, Ahmed et al. estimated the grid voltage adaptively as follows

\[
\hat{u}_{ga}(k+1) = \hat{u}_{ga}(k) - \lambda \frac{T_s}{L_{fg}} [i_{ga}(k) - \hat{i}_{ga}(k)]
\]

(3.9)

where \( \lambda \) is a positive gain multiplying the grid current estimation error. In fact, (3.9) can be interpreted as integration of the estimation error. An upper limit for the gain was provided to satisfy a stability criterion, but a detailed explanation of the gain selection, estimator tuning and convergence was not presented. For the frequency and phase estimation, Ahmed et al. (2009) used a separate PLL. Since Ahmed et al. derived (3.9) using the steepest descend method, as Mohamed et al. (2007) did for the L filter, the estimation algorithm is similar to the algorithm presented by Mohamed et al.

Although the observer by Ahmed et al. (2009) internally estimates the state vector, Ahmed et al. applied another parallel Kalman filter to the estimation of the states that were used in feedback control. Later, Xue
et al. (2012) also applied the estimation principle by Ahmed et al. (2009) and used a parallel Kalman filter for the state estimation. For the gain selection, Xue et al. reported that $\lambda$ should be sufficiently small to avoid instability problems.

In the adaptive estimation method by Mariéthoz and Morari (2009), the estimation law for the grid voltage is similar to (3.9). In the combined grid-voltage and state estimation, Mariéthoz and Morari used the structure of the Kalman filter and a piecewise affine model for the converter and LCL filter instead of the observer (3.8). The feedback gain for the Kalman filter was numerically calculated using a function provided in the Matlab software. Moreover, any other explanation of the gain selection and convergence of the estimated quantities was not presented in their work.

A grid-voltage estimation method based on an adaptive observer was proposed by Mohamed et al. (2012). However, in the method, all the states $x = [i_s^c, u_s^c, i_s^g]^T$ are measured for calculation of the state estimation error $x - \hat{x}$, which is then input for an adaptation algorithm. Since the whole state vector $x$ has to be measured, extra sensors are needed, thus meaning that the method is not truly AC-voltage sensorless.

### 3.4 Proposed Adaptive Observer

An adaptive observer for the state and grid-voltage estimation was presented in Publication II. Contrary to the methods of Ahmed et al. (2009) and Mariéthoz and Morari (2009), the observer does not estimate the phase quantities, but it operates in a synchronous reference frame and estimates space-vector quantities. Moreover, the magnitude $u_g$, frequency $\omega_g$ and angle $\vartheta_g$ of the grid voltage are estimated with the proposed observer. Whereas the adaptive observers by Mohamed and El-Saadany (2011) and Miskovic et al. (2016) are based on the L-filter approximation, the observer in Publication II is based on the full-order LCL-filter model (2.6),

$$\frac{d}{dt}\hat{x} = \hat{A}\hat{x} + \hat{B}_{c}u_{c} + \hat{B}_{g}\hat{u}_{g} + K_{o}\hat{i}_{c}, \quad \hat{i}_{c} = C_{c}\hat{x},$$

where estimated quantities are marked with a hat, $\hat{i}_{c} = i_{c} - \hat{i}_{c}$ is the estimation error of the converter current, $K_{o}$ is the gain for the estimation error feedback, and $\hat{A}$, $\hat{B}_{c}$, and $\hat{B}_{g}$ consist of the estimated system parameters $\hat{L}_{fc}$, $\hat{C}_{l}$, $\hat{L}_{lg}$, and $\hat{\omega}_{g}$. The real and imaginary components of the estimation error $\hat{i}_{c} = \hat{i}_{cd} + j\hat{i}_{cq}$ are the inputs for PI-adaptation mechanisms, which output the estimated grid-voltage magnitude and angular
frequency as follows
\[
\hat{u}_g = k_{p,u} \tilde{i}_{cd} + k_{i,u} \int \tilde{i}_{cd} dt, \quad \hat{\omega}_g = k_{p,\omega} \tilde{i}_{cq} + k_{i,\omega} \int \tilde{i}_{cq} dt \tag{3.11}
\]
where \(k_{p,u}, k_{i,u}, k_{p,\omega},\) and \(k_{i,\omega}\) are the gains. Furthermore, the estimated grid-voltage angle \(\hat{\vartheta}_g\) is obtained according to (2.9)
\[
\hat{\vartheta}_g = \int \hat{\omega}_g dt \tag{3.12}
\]

In order to tune the state observer, symbolic expressions for the gains in \(K_o = [k_{o1}, k_{o2}, k_{o3}]^T\) were derived in Publication I and applied in Publication II. The resulting gains are functions of the system parameters and dynamic performance specifications. The estimation error dynamics of the full-order observer obtained from (2.6) and (3.10) were considered in the derivation. Since the system parameters \((\hat{L}_{fc}, \hat{C}_f, \hat{L}_{fg})\) are needed in the observer, they must be known or estimated. A parameter estimation method was presented in Publication IV.

The estimation error dynamics of the adaptive observer are non-linear with respect to the angle estimation error \(\vartheta_g - \hat{\vartheta}_g\) and magnitude estimation error \(u_g - \hat{u}_g\). Therefore, a dynamic model for the estimation errors was derived and linearized around an operating point in Publication II. Based on the linearized model, symbolic expressions as function of system parameters and dynamic performance specifications (such as the bandwidth of the grid-voltage magnitude estimation) were derived for the gains \(k_{p,u}, k_{i,u}, k_{p,\omega},\) and \(k_{i,\omega}\) in (3.11). Moreover, the linearized model was experimentally validated comparing simulated and measured estimation error responses. The linearized model enables analysis of the estimation error convergence.

### 3.5 Direct Discrete-Time Design for the Adaptive Observer

For the digital implementation, the observers based on the continuous-time models have to be discretized. A simple and commonly used method for the discretization is the forward Euler method. The poles (eigenvalues) of the lossless LCL-filter model (2.6) are on the imaginary axis. If the forward Euler method is used for discretizing the model, the eigenvalues \(s = -j\omega_g, s = j\omega_p - j\omega_g,\) and \(s = -j\omega_p - j\omega_g\) are mapped \((z = 1 + sT_s)\) outside the unit circle in the Z-plane meaning that the discretized system would be unstable. In an observer, a proper selection for the gain \(K_o\) based on the continuous-time design may stabilize the discretized system.
However, the forward Euler method for the discretization may result in an unstable system even though the original continuous-time design is stable (Franklin et al., 1997). A discretization method that maps the stable continuous-time design into the stable discrete-time design is the Tustin method (bilinear transformation). Moreover, the imaginary-axis poles are mapped into the unit-circle poles in this transformation (Franklin et al., 1997). This method was used for the observers in Publications I and II.

An alternative to the continuous-time design and approximative discretization is the direct discrete-time design based on a discrete-time model, which considers the sample-and-hold characteristics of the digital implementation. The hold-equivalent discrete-time model (2.16) for the LCL filter was presented in Publication III and used in a state observer.

An adaptive observer based on this discrete-time model was proposed in Publication V for grid-voltage sensorless control. In the adaptive observer, the state vector is estimated as follows

\[
\hat{x}(k+1) = \Phi \hat{x}(k) + \Gamma_c u_c(k) + \Gamma_g \hat{u}_g(k) + K_o \hat{i}_c(k)
\]

where \( K_o = [k_{o1}, k_{o2}, k_{o3}]^T \) is the observer gain and \( \hat{i}_c = i_c - \hat{i}_c \) is the estimation error of the converter current. The matrices \( \Phi, \Gamma_c, \) and \( \Gamma_g \) consist of estimated system parameters \( \hat{L}_{fc}, \hat{C}_f, \hat{L}_{fg}, \) and \( \hat{\omega}_g \) and have the form of (2.17). An estimation method for the LCL-filter parameters has been presented, for instance, in Publication IV. For the observer tuning, symbolic expressions for the gains in \( K_o = [k_{o1}, k_{o2}, k_{o3}]^T \) were derived in Publication III and applied in Publication V.

In the adaptive observer, the estimation law for the grid-voltage magnitude is

\[
\hat{u}_g(k+1) = \hat{u}_g(k) + k_{i,u} \text{Re} \left\{ \frac{a}{b} e^{j\phi} \hat{i}_c(k) \right\}
\]

where \( k_{i,u} \) is the gain and \( a, b, \) and \( \phi \) are model-dependent constants. These constants were derived in Publication V based on a linearized model for the estimation errors. Furthermore, based on this model, a symbolic expression for \( k_{i,u} \) was presented as a function of the magnitude-estimation bandwidth. The estimation method for the grid-voltage frequency has two steps. First, a naturally filtered estimate \( \hat{\omega}_{gf} \) for the angular frequency is obtained from the discrete-time integral

\[
\hat{\omega}_{gf}(k+1) = \hat{\omega}_{gf}(k) + \frac{k_{i,\omega}}{u_{g0}} \text{Im} \left\{ \frac{a}{b} e^{j\phi} \hat{i}_c(k) \right\}
\]

where \( u_{g0} \) is an operating point quantity. Then, the estimate \( \hat{\omega}_g \) is obtained
as the output of the PI mechanism

$$\hat{\omega}_g(k) = \frac{k_{p,\omega}}{u_{g0}} \text{Im}\left\{\frac{a}{b} e^{j\bar{\phi}} \hat{i}_c(k)\right\} + \hat{\omega}_g(k)$$

(3.16)

The grid-voltage angle estimator is

$$\hat{\vartheta}_g(k + 1) = \hat{\vartheta}_g(k) + T_s \hat{\omega}_g(k)$$

(3.17)

Symbolic expressions for the gains $k_{i,\omega}$ and $k_{p,\omega}$ in the frequency estimators were derived based on the linearized model in Publication V. The resulting expressions are functions of the dynamic performance specifications for the angle and frequency estimation. The linearized model was used to analyze small-signal stability and dynamic properties of the observer. Furthermore, the linearized model was experimentally validated by comparing theoretical and measured estimation error responses.

3.6 Estimation of the Positive-Sequence and Negative-Sequence Components

Grid-connected converters may operate in unbalanced grid voltage conditions. In the grid-voltage estimation, the unbalanced grid voltage is often separated into positive and negative-sequence components (2.10). In the methods (Ahmed et al., 2009; Mariéthoz and Morari, 2009; Mohamed and El-Saadany, 2011) where all the three-phase voltages are estimated, some information about unbalanced grid conditions is directly obtained. Moreover, feeding the estimated phase voltages through a sequence separation block (Ahmed et al., 2011b), estimates for the positive and negative-sequence component of the grid voltage are obtained.

In Publication V, cascaded second-order notch filters were added to the estimation laws (3.14) and (3.16) of the adaptive observer in order to estimate only the positive-sequence grid-voltage component under distorted and unbalanced conditions. Sequence separation (or fundamental-component detection) strategies have been used in the case of virtual-flux estimation algorithms as well, for example, by Malinowski et al. (2003), Serpa et al. (2007), and Kulka (2009). However, the sequence separation cascaded with the estimation algorithm slows down the transient behavior of the estimated quantity as reported by Suul et al. (2012b). To avoid this, Suul et al. proposed a virtual-flux estimation algorithm with inherit sequence separation based on the second-order generalized integrators. Later, this method was extended for the converter equipped with an LCL filter by Wrona and Malon (2014).
Figure 3.4. Structure of the adaptive observer augmented with a disturbance model.

The positive and negative-sequence components are inherently estimated and separated in the observers augmented with the disturbance models for these components, as in the method by Song et al. (2003) (cf. (3.4)). A sequence separation strategy based on a disturbance model was also built in an augmented adaptive observer proposed in Publication VII and is further discussed in Section 3.7 below.

3.7 Proposed Augmented Adaptive Observer

The general observer structures were shown in Figure 3.3. The observers augmented with a suitable disturbance model (cf. (3.4)) are able to estimate the positive and negative-sequence components of the grid voltage in the case of unbalanced grid conditions. However, the grid-voltage angular frequency $\omega_g$ is a parameter needed in the disturbance model. On the other hand, with the presented adaptive observers the frequency and positive-sequence quantities can be estimated, but the negative-sequence component of the grid voltage is not directly obtained. In Publication VII, the structures of the adaptive observer (Figure 3.3(b)) and the observer augmented with a disturbance model (Figure 3.3(a)) were combined and an augmented adaptive observer was proposed. The structure of the proposed observer is illustrated in Figure 3.4. The observer simultaneously estimates the angular frequency $\hat{\omega}_g$, positive-sequence magnitude $\hat{u}_{g+}$, positive-sequence angle $\hat{\vartheta}_{g+}$, and negative-sequence component $\hat{u}_{g-} = \hat{u}_{gd-} + j\hat{u}_{gq-}$ of the grid voltage together with the states $\hat{x}$ of the LCL-filter model.

In the observer, the augmented state vector $x_a$, including the negative-
sequence component \( u_{g-} \), is estimated with the augmented structure

\[
\begin{bmatrix}
\hat{x}(k + 1) \\
\hat{u}_{g-}(k + 1)
\end{bmatrix} =
\begin{bmatrix}
\Phi & \hat{\Gamma}_{g-} \\
0 & e^{-2j\omega_0 T_s}
\end{bmatrix}
\begin{bmatrix}
\hat{x}(k) \\
\hat{u}_{g-}(k)
\end{bmatrix} +
\begin{bmatrix}
\hat{\Gamma}_c \\
0
\end{bmatrix} u_c(k) +
\begin{bmatrix}
\hat{\Gamma}_{g+} \\
0
\end{bmatrix} \hat{u}_{g+}(k)
\]

\[
\begin{bmatrix}
\hat{x}_a \\
\hat{u}_{g-}(k) + \hat{\Gamma}_{g-} 0
\end{bmatrix}
\]

\[
+ K_o [\hat{i}_c(k) - \hat{i}_c(k)]
\]

where \( K_o = [k_{o1}, k_{o2}, k_{o3}, k_{o4}]^T \) is the observer gain. Furthermore, \( \hat{\Gamma}_{g+} \) and \( \hat{\Gamma}_{g-} \) are the input vectors for the positive and negative-sequence components, respectively, which have been presented in more detail in Publication VII. In the structure, the disturbance model for \( u_{g-} \) (under the dashed line) is a discrete-time equivalent to the dynamics of \( u_{g-} \) in (3.3). The estimation laws for \( \hat{u}_{g+}, \hat{\vartheta}_{g+}, \) and \( \hat{\omega}_g \) have forms similar to (3.14), (3.17), and (3.16), respectively. However, the detailed formulation and gain selection for the estimation are based on a linearized model that was derived for the augmented adaptive observer in Publication VII.
4. Control System

The proposed grid-voltage estimation methods operate as a part of a grid-voltage sensorless control system. The structure of the sensorless control system applied in this thesis is illustrated in Figure 3.2 in Section 3.1. This chapter briefly describes the implemented control functions. Then, the experimental setup used to test the methods developed in this thesis is presented.

4.1 State-Space Current Control

In grid-voltage sensorless control systems, various current control structures have been applied in the case of the converter equipped with an LCL filter. The current controller has been, for example, a PI controller in the synchronous reference frame (Gullvik et al., 2007; Malinowski and Bernet, 2008; Ahmed et al., 2009), a proportional-resonant (PR) controller in the stationary reference frame (Roslan et al., 2015), a model-predictive controller (Mariéthoz and Morari, 2009; Ahmed et al., 2011a), or a state-space controller in the work by Bolsens et al. (2006) and in Publications II, V, and VII. The adaptive observers developed in this thesis estimate the state vector \( \mathbf{x} = [i_c, u_f, i_g]^T \) of the LCL-filter model together with the grid voltage. In order to effectively use the estimation result in control, a state-space control structure is applied for current control in this thesis.

Although the state-space controller has a more complex structure in comparison with the PI or PR controller, the structure enables arbitrary closed-loop pole placement for the system. The closed-loop pole locations can be selected directly based on the poles of the open-loop system and dynamic performance specifications (e.g., bandwidth and damping), which is convenient in the active damping of LCL-filter resonance (Publications I and III). An alternative to direct pole placement is optimization of some
A state-space control structure for current control is illustrated in Figure 4.1. The control voltage is
\[ u'_{c,\text{ref}} = k_1 i_{c,\text{ref}} + k_I x_I - K \hat{x} - k_4 u_c \] (4.1)
where \( k_1 \) is the feedforward gain, \( k_I \) is the gain of the integral state \( x_I \), and the gain \( K = [k_1, k_2, k_3] \) together with the gain \( k_4 \) forms the state feedback. Two different model-based design methods for the state-space controller were presented: The controller was designed based on the continuous-time model (2.6) in Publication I and based on the discrete-time model (2.16) in Publication III. In the discrete-time design, the computational delay of the control algorithm can be taken into account in the discrete-time model resulting in the extra state \( u_c \) and the gain \( k_4 \). In the continuous-time design, the feedback from \( u_c \) does not exist. Moreover, the other gains and the integration algorithm \( I(z) \) become different in these different design approaches as explained in Publication I and Publication III. Since both of the design approaches are based on the LCL-filter model, the model parameters must be known or estimated. In unbalanced or distorted grid conditions, the control law (4.1) can be augmented with reduced-order generalized integrators (ROGIs) (Busada et al., 2012; Harnefors et al., 2016b), as demonstrated in Publications V and VII.

Figure 4.1. State-space control structure for current control. The blocks inside the gray area are related to the coordinate transformations and the control-voltage saturation to the realizable hexagon. The integrator windup is prevented using the back-calculation method (Åström and Hägglund, 1995; Harnefors and Nee, 1998).
Although the control law is almost equal in both design approaches, the discrete-time design process is more complex, because of the more complex model. However, higher dynamic performance and better resonance damping of the LCL filter can be achieved with the controller designed in the discrete-time domain in comparison with the controller designed in the continuous-time domain as demonstrated in Publication III. Moreover, the handling of the computational delay in the system is straightforward in the discrete-time design, which is beneficial especially at lower sampling (switching) frequencies.

Figure 4.1 includes some items that were not explained in the publications in detail. The maximum converter voltage is limited and the voltage reference may exceed the limit in large transients. Therefore, the voltage reference $u_{\text{c,ref}}$ is limited inside the realizable area (hexagon), and the limited voltage $\bar{u}_{\text{c,ref}}$ is the input for the PWM. The converter voltage is not measured, but the voltage is calculated for the controller and observer from the limited reference as $u_{\text{c}}(t) = \bar{u}_{\text{c,ref}}(t - T_d)$, where $T_d$ is the time delay related to the finite computation time and the PWM. In the continuous-time design, the time delay is modeled as $T_d = 1.5 T_s$, where $T_s$ explains the finite computation time and $0.5 T_s$ the sample-and-hold of PWM. In the discrete-time design $T_d = T_s$, because the ZOH is already included in the discrete-time model (2.16). Since $\vartheta_g(t) = \vartheta_g(t - T_d) + \omega_g T_d$, the converter voltage in the synchronous reference frame can be calculated from the reference voltage as

$$u_{\text{c}}(t) = e^{-j \vartheta_g(t)} u_{\text{c}}(t) = e^{-j \vartheta_g(t - T_d)} \bar{u}_{\text{c,ref}}(t - T_d) = \bar{u}_{\text{c,ref}}(t - T_d) \quad (4.2)$$

The reference signals with the prime are defined as $\bar{u}_{\text{c,ref}}' = e^{-j \omega_g T_d} \bar{u}_{\text{c,ref}}$. When the voltage reference is saturated, the integrator windup is prevented (feedback with the gain $k_w$ in the figure) using the back-calculation method (Åström and Hägglund, 1995; Harnefors and Nee, 1998). In the case of ROGIs in the control loop, the windup prevention has been discussed, for example, by Harnefors et al. (2016b).

## 4.2 Reference Calculation

In this thesis, the grid-side powers (2.15) are indirectly controlled by controlling the converter-side current $i_c$. In balanced grid conditions, the grid
current in steady state is

\[ i_g = (i_c - j \omega_g C_f u_g)/(1 - \omega_g^2 C_f L_{fg}) \]  \hspace{1cm} (4.3)

which is obtained by making \( d/dt = 0 \) in (2.6). If the \( d \)-axis of the synchronous reference frame is aligned with the grid-voltage vector, to be precise, \( u_g = u_g + j0 \), the powers as a function of the converter current become

\[ p_g = \frac{3}{2(1 - \omega_g^2 C_f L_{fg})} u_g i_{cd} \]  \hspace{1cm} (4.4)

\[ q_g = \frac{3}{2(1 - \omega_g^2 C_f L_{fg})} (-u_g i_{cq} + \omega_g C_f u_g^2) \]  \hspace{1cm} (4.5)

For power control, the current references \( i_{cd,ref} \) and \( i_{cq,ref} \) are calculated from these power expressions as a function of the power references \( p_{g,ref} \) and \( q_{g,ref} \), respectively.

In unbalanced grid conditions, different control objectives can be set for the converter current and the active and reactive powers. For example, the objective can be balanced converter currents or elimination of \( 2\omega_g \) oscillations from \( p_g \) or \( q_g \). In order to meet these conditions, different strategies for calculating current references have been discussed by Rodriguez et al. (2007), Wang et al. (2011), Teodorescu et al. (2011), and Suul et al. (2012a). In some strategies, references for the positive and negative-sequence current components are calculated using the voltage components \( u_g^+ \) and \( u_g^- \). Estimated positive and negative-sequence virtual-flux components were applied in the reference calculation by Suul et al. (2012a). In addition, the estimated voltage components (\( \hat{u}_g^+ \) and \( \hat{u}_g^- \)) can be applied in the calculation as demonstrated in Publication VII.

### 4.3 DC-link Voltage Control

DC-link voltage control is a part of the grid-voltage sensorless control system. In the publications, the DC-link voltage was controlled indirectly using the capacitor energy \( W \) in (2.14) as an auxiliary state variable. With this selection nonlinear dynamics (2.13), with respect to the DC-link voltage \( u_{dc} \), become linear with respect to \( W \) (Hur et al., 2001; Ottersten, 2003). The error signal derived from the desired energy \( W_{ref} \) and actual energy \( W \)

\[ e_w = W_{ref} - W = \frac{1}{2} C_{dc}(u_{dc,ref}^2 - u_{dc}^2) \]  \hspace{1cm} (4.6)
Control System

is the input for the PI controller

\[ p_{\text{ref}} = k_{p,\text{dc}} e_w + k_{i,\text{dc}} \int e_w \, dt \]  \hspace{1cm} (4.7)

that regulates the DC-link voltage and provides the power reference \( p_{\text{ref}} \) for the current reference calculation as the output. If the current control bandwidth is assumed to be high in comparison with the voltage control bandwidth, it is possible to approximate \( p_{\text{sw}} \approx p_{\text{ref}} \) (cf. Figure 2.3). Then, the transfer function from the desired energy to the actual energy is obtained from (2.13), (4.6), and (4.7)

\[ \frac{W(s)}{W_{\text{ref}}(s)} = \frac{-k_{p,\text{dc}} s - k_{i,\text{dc}}}{s^2 - k_{p,\text{dc}} s - k_{i,\text{dc}}} \]  \hspace{1cm} (4.8)

The pole locations of the transfer function can be determined with the gains, for example, as \( k_{i,\text{dc}} = -\omega_{\text{dc}}^2 \) and \( k_{p,\text{dc}} = -\zeta_{\text{dc}} \omega_{\text{dc}} \), where \( \omega_{\text{dc}} \) and \( \zeta_{\text{dc}} \) are the natural frequency and damping ratio of the resulting pair of poles, respectively. Alternatively, a gain selection from a passivity point of view has been proposed by Harnefors et al. (2016a).

Since the capacitance value \( C_{\text{dc}} \) is needed in the formulation of the error signal (4.6), it has to be either known or estimated. A method for estimating the capacitance value is proposed in Publication VI.

In the practical implementation, the controller was discretized using the Forward Euler method and the output of the controller was limited. The limited reference \( \bar{p}_{\text{ref}} \) is between the minimum and maximum power allowed for the converter. The back-calculation method (Åström and Hägglund, 1995; Harnefors and Nee, 1998) was applied for the integral windup prevention feeding back the difference \( \bar{p}_{\text{ref}} - p_{\text{ref}} \) to the input of the integrator as also demonstrated by Ottersten (2003).

In unbalanced grid conditions, the measured DC-link voltage \( u_{\text{dc}} \) possibly contains \( 2\omega_g \) ripple, which results from the corresponding ripple in the active power \( p_g \), if the ripple is not compensated. Therefore, for control in unbalanced conditions (Publication V and Publication VII), the measured voltage was filtered using a second-order notch filter at \( 2\omega_g \) in order to attenuate the ripple and to obtain average power reference as the output of the controller.

### 4.4 Startup of Sensorless Control

In the startup of grid-voltage sensorless control, the initial grid-voltage angle is unknown and, therefore, it is separately estimated in order to
prevent high inrush currents and to enable convergence of the actual estimation algorithm. In this thesis, the initial angle estimation follows a simple method applying a zero-vector pulse in $u_c$ (Kwon et al., 1999; Barrass and Cade, 1999). The initial magnitude and frequency of the grid voltage can also be estimated or the nominal values for these quantities can be used to initialize the frequency and magnitude estimators.

### 4.5 Experimental Setup

Figure 4.2 shows the photograph and Figure 4.3 the block diagram of the test setup, which were used to experimentally verify the methods presented Publications I...VII. The setup has two back-to-back connected converters equipped with LCL filters. The 12.5-kVA converter under test is connected directly to the 50-Hz electric power distribution system via a 1-MVA 20-kV/400-V transformer or to the 50-kVA three-phase four-quadrant power supply (Regatron TopCon TC.ACS), which is used to emulate grid voltage disturbances. The other 12.5-kVA converter (ABB ACSM1-204AR-016A-4), is used in the power-control mode to feed (or draw) power to the common DC link. This loading converter is connected to the grid via an isolation transformer.

The hardware of the converter under test is similar to the load converter (ABB ACSM1-204AR-016A-4), but the original controller board of the converter under test is replaced with an interface board for use with dSPACE systems. The estimation and control algorithms for this converter were implemented in a dSPACE system consisting of DS1006, DS2201, and DS5202 boards. First, the algorithms were programmed in
Matlab/Simulink environment and then compiled and uploaded in the dSPACE system. The PWM interface of the dSPACE system was used to generate the switching signals for the converter hardware. The sampling of the measured quantities was synchronized with the PWM. The phase currents are measured using external closed-loop Hall-effect current transducers (LEM). The DC-link and grid voltages are measured using external closed-loop Hall-effect voltage transducers (LEM).

At 400 V (rms) line-to-line voltage, the rated current and power of the converter under test are 18 A (rms), and 12.5 kVA, respectively. The measured parameter values of its LCL filter (ABB WFU-02) are $L_{fc} = 3.3 \, \text{mH}$, $C_f = 8.8 \, \mu\text{F}$, and $L_{fg} = 3.0 \, \text{mH}$. The measured capacitance of the common DC-link is 1.8 mF.

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1 In Publications I...V inaccurate parameter values ($L_{fc} = 2.94 \, \text{mH}$, $C_f = 10 \, \mu\text{F}$, and $L_{fg} = 1.96 \, \text{mH}$) were used which were received from the manufacturer. Before Publication VII, the fundamental-frequency inductance values were measured at the rated current using a power analyzer (Voltech PM6000) and the capacitance was measured using an RLC meter (Philips PM6304). The experimentally identified values in Publication IV are close to the measured values.
5. Summaries of Publications

5.1 Abstracts

The abstracts of the publications are reprinted in this section.

Publication I

This paper presents a state-space current control method for active damping of the resonance frequency of the LCL filter and setting the dominant dynamics of the converter current through direct pole placement. A state observer is used, whereupon additional sensors are not needed in comparison with the conventional L filter design. The relationship between the system delay and instability caused by the resonance phenomenon is considered. Nyquist diagrams are used to examine the parameter sensitivity of the proposed method. The method is validated with simulations and experiments.

Publication II

This paper proposes a simple grid-voltage sensorless alternative to the conventional PLL synchronization for a grid-connected converter equipped with an LCL filter. The grid-voltage magnitude and angle are estimated based only on the converter current and DC-voltage measurements. Reduced number of sensors decreases costs and amount of sensor wiring. The analytically derived design is validated experimentally.

Publication III

State-space current control enables high dynamic performance of a three-phase grid-connected converter equipped with an LCL filter. In this paper,
observer-based state-space control is designed using direct pole placement in the discrete-time domain and in grid-voltage coordinates. Analytical expressions for the controller and observer gains are derived as functions of the physical system parameters and design specifications. The connection between the physical parameters and the control algorithm enables automatic tuning. Parameter sensitivity of the control method is analyzed. The experimental results show that the resonance of the LCL filter is well damped, and the dynamic performance specified by direct pole placement is obtained for the reference tracking and grid-voltage disturbance rejection.

**Publication IV**

Model-based control techniques are frequently used with grid-connected converters. This paper proposes a method for identifying model parameters of an LCL filter connected to a grid converter for control purposes. In identification, a discrete-time autoregressive moving average with exogenous input (ARMAX) model structure is used and closed-loop current control is considered. The resulting discrete-time model parameters are translated into the continuous-time physical parameters (inductance and capacitance values of the filter) by comparing the estimated discrete-time model with the analytical discrete-time model. Simulation and experimental results show that the proposed method yields good parameter estimates that are suitable for control tuning.

**Publication V**

Synchronization with the power system is an essential part of control of grid-connected converters. This paper proposes a grid-voltage sensorless synchronization and control scheme for a converter equipped with an LCL filter, measuring only the converter currents and the DC voltage. A discrete-time pole-placement design method is used to formulate an adaptive full-order observer for estimation of the frequency, angle, and magnitude of the grid voltage. The proposed discrete-time design method enables straightforward implementation and suits low sampling rates better than its continuous-time counterpart. The analytically derived design is experimentally validated, and the results demonstrate rapid convergence of the estimated quantities. Moreover, the experimental tests show that grid-voltage sensorless operation is possible under balanced and unbal-
anced grid disturbances and distorted grid conditions.

**Publication VI**

A DC-link capacitance estimation method is presented for a three-phase voltage-source converter. The measured phase currents and DC-link voltage of the converter are used in formulation of the input and output signals for the estimator. The estimator is based on the ordinary least squares method. The capacitance can be estimated during the startup of the converter or while the converter is in normal operation. The proposed method is experimentally validated and the results indicate that the capacitance value is estimated with a good accuracy.

**Publication VII**

This paper deals with grid-voltage sensorless synchronization and control under unbalanced grid conditions. A three-phase grid-connected converter equipped with an LCL filter is considered, and no other signals than the converter currents and the DC-link voltage are measured for control. An augmented adaptive state observer is proposed for estimation of the positive- and negative-sequence components of the grid voltage. The proposed observer is tested as a part of a sensorless control system. Experimental results show that the proposed method works well even in highly unbalanced grid conditions.

5.2 **Contribution of the Thesis**

The main contributions of the thesis can be summarized as follows:

- An adaptive observer was proposed for combined state and grid-voltage estimation in order to enable grid-voltage sensorless control of converters equipped with an LCL filter in Publications II and V. The observer is able to estimate the state vector \( x = [i_c, u_f, i_g] \) together with the grid-voltage magnitude \( u_g \), angular frequency \( \omega_g \), and angle \( \vartheta_g \).

- For the adaptive observer, a design procedure based on an LCL-filter model was explained in the continuous-time domain in Publication II and in the discrete-time domain in Publication V.
• Symbolic expressions were derived for the observer gain vector based on direct-pole placement in the continuous and discrete-time domains in Publications I and III, respectively, enabling automatic tuning of the observer.

• The nonlinear estimation error dynamics of the adaptive observer were linearized and analyzed in Publications II and V in the continuous and discrete-time domains, respectively. Based on the linearized model and analysis: estimation laws for the grid voltage were proposed, symbolic expressions for the estimator gains were derived, and the gain selection was discussed.

• An augmented adaptive observer was proposed in Publication VII for grid-voltage sensorless control of converters equipped with an LCL filter in unbalanced grid conditions. The observer estimates the state vector and positive-sequence magnitude $u_{g+}$, angle $\vartheta_{g+}$, and angular frequency $\omega_{g}$ of the grid voltage together with the negative-sequence component $u_{g-}$ of the grid voltage.

• The nonlinear estimation-error dynamics of the augmented adaptive observer were linearized and analyzed providing the basis for the observer tuning and gain selection in Publication VII.

• A grid-voltage sensorless control system using the proposed adaptive observers and a state-space current controller was built, documented, and experimentally tested (Publications I, II, III, V, VII, and Chapter 4).
Grid-voltage estimation plays an important role in grid-voltage sensorless control of a grid-connected converter. This thesis examined grid-voltage estimation methods for this control scheme. Basic principles have been introduced for converters equipped with an L filter, with the majority of the developed methods being ultimately based on the inductor voltage equation. LCL filters are currently increasingly used between the converters and the grid. In the grid-voltage estimation, the LCL filters have been approximated as L filters, which is reasonable only at low frequencies. Alternatively, only the capacitor voltage of the LCL filter has been estimated.

The dynamic behavior of the LCL filter can be taken into account in the estimation, if a full-order model for the LCL filter is used. The grid voltage can be estimated together with the state vector of the full-order model using an observer. The observed state vector enables state-space current control and direct closed-loop pole placement, which is a flexible and convenient way to determine dynamic behavior (e.g., the bandwidth and LCL-filter resonance damping) of the current-control loop.

In the case of the LCL filter, some observers for the combined state and grid-voltage estimation have been presented in the literature. The majority of them are Kalman filters augmented with a disturbance model or combined with an adaptation mechanism. However, a numerical computation software or trial-and-error method has been used in order to tune these observers. In addition, a clear link between the observer gains and the dynamic behavior of the estimated quantities is missing in these methods.

This thesis (Publications II and V) proposed an adaptive observer for the combined state and grid-voltage estimation for grid-voltage sensorless control of converters equipped with an LCL filter. The nonlinear estima-
tion error dynamics of the observer were linearized, and a model-based design procedure for the observer was presented in the continuous-time domain (Publications I and II) and in the discrete-time domain (Publications III and V). Due to the presented design procedure being purely analytical, a link between the observer gains and dynamic behavior is retained. Moreover, symbolic expressions for the gains were derived as a function of the dynamic performance specifications and model parameters. This generalizes the estimation method for different converter systems and enables the automatic tuning of the method, for instance, in the startup of the converter, if the model parameters are known or estimated. A parameter identification method was presented in Publication IV.

During grid faults, the grid phase voltages may be unbalanced. In these conditions, the grid-voltage vector can be divided into the positive and negative-sequence components. These components can be directly estimated using observers with disturbance models. However, the grid-voltage frequency is a parameter needed in the disturbance models.

The adaptive observer estimates the frequency. The structure of the adaptive observer can be augmented with a disturbance model as shown in Publication VII. This results in a frequency-adaptive observer for estimation of the positive and negative-sequence components. Moreover, the estimated components can be used in the converter control system to meet the control objectives set for the unbalanced conditions.

The proposed model-based design procedure requires the information of the LCL-filter parameters (inductance and capacitance values). Moreover, the converter is connected to the grid, whose impedance is time and frequency dependent. Generally, trade-offs between the dynamic performance and robustness against parameter errors have to be accepted in the control design. Since the dynamic performance specifications are inputs for the proposed observer designs, these specifications can be altered in order to obtain the required robustness. The required robustness depends on the application, and this thesis did not address this topic in detail. Hence, it provides a suitable topic for future research. A related topic for future research could be to examine effect of different grid impedances on the estimation methods and sensorless control.

Finally, the analytical design procedure and analyses presented in this thesis are scientifically valuable and may inspire further studies of adaptive observers. Furthermore, the developed methods can be applied in industrial products.


References


Errata

Publication IV

The sign of the term $\gamma^3 z^{-3}$ in the denominators of Equations (5) and (23) is incorrect. The denominators should be $1 + \alpha_1 \gamma z^{-1} - \alpha_1 \gamma^2 z^{-2} - \gamma^3 z^{-3}$ instead of $1 + \alpha_1 \gamma z^{-1} - \alpha_1 \gamma z^{-1} - \alpha_1 \gamma^2 z^{-2} - \gamma^3 z^{-3}$.

The sign of the term $\gamma^3 z^{-3}$ in Equation (9) is incorrect. The correct form of the polynomial $A(z)$ is

$$A(z) = 1 + \alpha_1 \gamma z^{-1} + (\beta_1 k_p \gamma - \alpha_1 \gamma^2) z^{-2} + (\beta_2 k_p \gamma^2 - \gamma^3) z^{-3} + \beta_1 k_p \gamma^3 z^{-4}$$

Publication V

The first symbol inside the brackets on the line following the Equation (44) should be $\hat{L}_k$ instead of $\hat{L}_{lg}$. 

Grid-connected converters are widely used in energy production and consumption. The grid voltage is typically measured, and hence voltage sensors are required for control of these converters. However, it is beneficial from system cost and reliability point of views that the converters are able to estimate the grid voltage and they can operate without the grid-voltage sensors. This thesis deals with estimation methods for grid-voltage sensorless control of the converters equipped with an LCL filter. For the grid-voltage estimation, an adaptive observer is proposed, analyzed, and experimentally tested.

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Estimation Methods for Grid-Voltage Sensorless Control of Converters Equipped with an LCL Filter

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