Modeling and Control of Voltage Source Converters Connected to the Grid

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by

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Abstract

This thesis deals with the modeling and control of pulse width modulated (PWM) voltage source converters connected to the grid. When voltage source converters are connected to the grid, the power quality and the dynamic performance are affected by the line filter connected between the converter and the grid, and by nonlinearities caused by the switching converter. In the thesis, the dynamic performance and power quality of converters connected to the grid by first-order L-filters and third-order LCL-filters are focused on. For each line filter, predictive vector control principles that allow the independent control of the active and the reactive powers are developed and verified by measurements. It is shown that a similar dynamic performance can be obtained with both line filters. To obtain a high power quality, it is advantageous to use LCL-filters. The thesis also deals with different nonlinearities and their influence on dynamic performance. Measured small-signal frequency responses are compared with Bode diagrams obtained from linear analytical models. The analytical models are created by using a technique based on state space equations that is developed in the thesis. In the thesis, switching frequencies and sampling frequencies from 5 to 7 kHz are used. At such switching and sampling frequencies, nonlinearities and cross coupling caused by the uniform PWM method and the coordinate transformation in the control system do not have any significant effect on the small-signal frequency responses for frequencies below 1 kHz. It is, however, shown that a more ideal performance of the vector control system can be obtained by compensating for errors due to the non-ideal commutation caused by the blanking time and on-state voltage drops across the non-ideal IGBT valves of the converter. Moreover, measurements have shown that non-modeled losses in the line filter inductors have an impact on the small-signal frequency responses. When voltage source converters are used as active filters, the performance of the active filters are affected by phase shifts in the current control system, and also by the cross coupling between the control of the active and the reactive currents. Different principles for compensating for the phase shifts have been evaluated. Measurements show that it is possible to compensate for the phase shifts and thereby obtain efficient active filters also at moderate switching frequencies such as 5 to 7 kHz.

Keywords: Power electronics, Voltage source converter, IGBT valve, Line filter, L-filter, LCL-filter, Pulse width modulation, switching frequency, sampling frequency, small-signal, frequency response, transfer function, active filtering.

Preface

This thesis is an important part of the result of my time at the Department of Electric Power Engineering at Chalmers University of Technology. It has been a privilege for me to work in an area related to the proper utilization of energy resources. Hopefully, the results will contribute to an improved base for design of power converters, or, at least, to future relevant research projects.....

First of all, I would like to thank Sydkraft ABs Research Foundation for fully financing the project. I also had the support from the reference group constituted by Bengt Ahlman and Peter Bäckström at AB Sydkraft and Professor Mats Alaküla at the department of Industrial Electrical Engineering and Automation, Lund Institute of Technology. I would also like to thank Professor Alaküla and Anders Carlsson at the same department for making it possible to utilize the DSP system developed at their department. I would also like to thank my industrial adviser, Tommy Lejonberg, previously with ABB and now Technical Director of Elmo Industries for useful comments on the research.

At our department at Chalmers, first of all, I would like to thank Dr. Jan Svensson for all cooperation during the preparation of the experimental system and the measurements in the laboratory, as well as several of the publications in the thesis. Furthermore, I would like to thank Professor Jorma Luomi for his support throughout the project and for his excellent comments, proof reading, and general advice during the preparation of the thesis. I also want to thank my supervisors during the first part of the project; Dr. Tore Svensson and Professor Kjeld Thorborg.

I also want to express my sincere gratitude to my dear parents Aina and Bengt for their encouragement and support.

Finally, thank you Karin for your love and never ending support.

List of Appended Publications

A M. Lindgren, J. Svensson, "Connecting Fast Switching Voltage Source Converters to the Grid — Harmonic Distortion and its Reduction," *IEEE Stockholm Power Tech Conference*, Stockholm, Sweden, June 18-22, 1995, Proceedings of Power Electronics, p. 191-195.

B M. Lindgren, "Analysis and Simulation of Digitally Controlled Grid-connected PWM-converters using the Space-vector Average Approximation, " *IEEE 5th Workshop on Computers in Power Electronics*, Portland, USA, August 11-14, 1996, p. 157-162.

C J. Svensson, M. Lindgren. "Influence of Non-linearities on the Frequency Response of a Grid-Connected Vector-Controlled VSC," to appear in *IEEE Transactions on Industrial Electronics*.

D J. Svensson, M. Lindgren, "Vector Current Controlled Grid Connected Voltage Source Converter — Influence of Non-linearities on the Performance," *IEEE Power Electronics Specialists Conference (PESC 98*), Fukuoka, Japan, May 17-22, 1998, Vol. 1, p. 531-537.

E M. Lindgren, J. Svensson, "A Method to Compensate for Errors Caused by the Blanking Time in PWM Systems with High Ripple Currents," *1998 IEEE Nordic Workshop on Power and Industrial Electronics* (*NORPIE/98*), Espoo, Finland, August 26-27, 1998, p. 147-152.

F M. Lindgren, "Feed forward — Time Efficient Control of a Voltage Source Converter Connected to the Grid by Lowpass Filters," *IEEE Power Electronics Specialists Conference (PESC 95)*, Atlanta, USA, June 18-22, 1995, Vol. 2, p. 1028-1032.

G M. Lindgren, J. Svensson, "Control of a Voltage-source Converter Connected to the Grid through a LCL-filter — Application to Active Filtering," *IEEE Power Electronics Specialists Conference (PESC 98*), Fukuoka, Japan, May 17-22, 1998, Vol. 1, p. 229-235.

List of Symbols

Symbol	Description	Unit
$\mathbf{A}$	system matrix of continuous state space model	
$\mathbf{B}$	input matrix of continuous state space model	
$\mathbf{C}$	output matrix	
C_f	capacitance of filter capacitor	[F]
C_{dc}	capacitance of dc link capacitor	[F]
$\mathbf{D}$	initial state matrix of state space model	
$\mathbf{F}$	system matrix of discrete state space model	
f_{res}	resonance frequency	[Hz]
f_s	sampling frequency	[Hz]
f_{sw}	switching frequency of a semiconductor valve	[Hz]
f_{mod}	frequency of carrier-wave	[Hz]
$\mathbf{G}$	input matrix of discrete discrete state space model	
i_{load}	current of dc load	[A]
$\underline{i}$	line current vector	
$\underline{i}_1$	converter current vector	
$\underline{i}_2$	line current vector when the LCL-filter is used	
I_{base}	base current	[A]
K_p	proportional gain	
K_{p1}	proportional gain in inner loop	
K_{p2}	proportional gain in outer loop	
K_{pd}	gain to obtain dead beat control of inductor current	
K_{pd2}	gain to obtain dead beat control of capacitor voltage	
L_1	inductance of filter inductor connected to the converter	[H]
L_2	inductance of filter inductor connected to the grid	[H]
L_{grid}	inductance of grid	[H]
M	ratio between output voltage and maximum output voltage	
P	ratio between switching frequency and line frequency	
R	resistance	[Ω]
R_1	resistance of filter inductor connected to converter	[Ω]
R_2	resistance of filter inductor connected to the grid	[Ω]
R_{onD}	resistance of diode of IGBT	[Ω]
R_{onT}	resistance of transistor of IGBT	[Ω]
$\underline{Sw}$	switching vector of converter output voltage	
s	Laplace operator	
T_{blank}	blanking time	[s]
T_s	sampling time	[s]
t_k	time at sample k	[s]
$\underline{u}_1$	converter output voltage vector	
$\underline{u}_2$	grid voltage vector	
$\underline{u}_c$	capacitor voltage vector	
$\mathbf{u}$	input vector of state space model	
U_{base}	base voltage	[V]
u_{dc}	dc link voltage	[V]
U_{D0}	forward voltage across diode of IGBT at zero current	[V]
U_{T0}	forward voltage across transistor of IGBT at zero current	[V]
$\mathbf{x}$	state vector of state space model	
X_C	reactance of filter capacitor	[p.u.]
X_L	reactance	[p.u.]
X_{L1}	reactance of filter inductor connected to converter	[p.u.]
X_{L2}	reactance of filter inductor connected to grid	[p.u.]

$\mathbf{y}$	output vector of state space model	
Z	impedance	[p.u.]
Z_{base}	base impedance	[Ω]
Z_{cable}	wave impedance of cable	[Ω]
Z_{conv}	wave impedance of converter	[Ω]
Z_{load}	wave impedance of load	[Ω]
Z_1	impedance of filter inductor connected to converter	[Ω]
Z_2	impedance of filter inductor connected to grid	[Ω]
z	z operator	
Δu_{blank}	error in average phase voltage due to blanking time	
Δu	compensation voltage for delay time	[V]
θ	angle of grid-flux vector	[rad]
θ_1	average angle of grid-flux vector of sampling interval k+1, k+2	[rad]
ω_{g}	angular frequency of grid voltage vector	[rad/s]
ω_{s}	angular frequency of grid voltage vector	[rad/s]
$\Gamma_{\mathrm{con}v}$	converter reflexion coefficient of converter	
Γ_{load}	reflexion coefficient of load	

Superscripts

*	reference
$\alpha\beta$	vector in $\alpha\beta$-reference frame
dq	vector in dq-reference frame

Subscripts

a	phase a
b	phase b
c	phase c
dc	dc link
d	vector component in d-direction
q	vector component in q-direction
$d0$	vector component in d-direction in steady state
$q0$	vector component in q-direction in steady state
α	vector component in α–direction
β	vector component in β–direction
L	L-filter
LCL	LCL-filter

Table of Contents

Part II. Appended Publications

Chapter 1

Introduction

1.1. Power Converters and Power Quality

In this thesis, switching power converters for industrial applications are investigated. As a result of tremendous developments in the semiconductor industry, switching pulse width modulation (PWM) converters are becoming cheaper and available at increased power levels. Today, PWM converters are widely used in industrial applications, such as adjustable-speed drives (ASDs) and uninterruptible power supplies (UPSs). This is advantageous since the use of power electronics results in equipment with higher performance and lower energy consumption. The development is, however, not only positive since almost all power electronic circuits behave as nonlinear loads. Therefore, harmonic currents are injected into the grid.

The nonlinear loading of the grid increases continuously since the standards for the emission of harmonic currents still allow a high total harmonic distortion. As a consequence, the power quality is reduced in several ways. Usually a poor power quality refers to the presence of disturbances and harmonic distortion in the grid voltage. Harmonic distortion and disturbances may cause malfunction of equipment, such as computers and ASDs. Harmonic currents can also cause fires due to overheating of neutral conductors. The problem of harmonics can be solved either by reducing the harmonic distortion, or by increasing immunity to harmonic distortion and designing the system to provide efficient and safe operation even with a large amount of harmonics present.

In industry, the major part of electricity is used to supply electrical machines. Due to the increased prices for electricity and reduced cost of power converters, more and more machines are being replaced with ASDs. In these drives, a converter from ac to dc voltage (i.e., a rectifier) is connected to the grid. Today, line commutated rectifiers using diodes or thyristors are used. Such a rectifier is a nonlinear load that generates harmonic currents.

As a result of the development of new semiconductors and control computers, it would be possible to use PWM rectifiers and, thereby, obtain sinusoidal line currents. Today, PWM converters are mainly used as converters from dc voltage to ac voltage (i.e., inverters) in ASDs and in UPSs. Compared with line-commutated rectifiers, significantly higher power quality and increased dynamic performance can be obtained with PWM rectifiers. These rectifiers are not usually used today since existing harmonic standards can be fulfilled by using cheap line-commutated rectifiers. This situation may, however, change as soon as customers start to request grid friendly rectifiers. Such a development can take place due to new standards that reduce the permitted emission of harmonic currents, or due to a reduced cost for grid friendly rectifiers.

1.2. Line Interference and Control

To obtain a high efficiency and a high power rating from a PWM converter, the current and the voltage across the semiconductor valves should not be high simultaneously. This is solved by using PWM. At each switching of a valve, however, a high instantaneous power dissipation occurs in the valve. Therefore, the switching times must be short to obtain high efficiency and rated power. As a consequence of very low switching times of below 1 μs, high voltage derivatives occur at each switching of a valve. These voltage derivatives give rise to disturbances at high frequencies. Both radiated and conducted emission may occur. The disturbances caused by switching are, consequently, affected by the switching frequency. If the switching frequency is reduced, the power dissipation in the valves at each switching may be increased and the switching speed can, thus, be reduced compared to operation at a higher switching frequency. Therefore, it is advantageous to operate at moderate switching frequencies. In high-power applications, the switching frequency has to be below a few kHz with the valves available today.

Due to the PWM principle, harmonics occur even when PWM converters are used. To obtain a converter system with sinusoidal line currents and a low emission of disturbances due to the switching, a line filter is connected between the converter and the grid. The selection of the line filter is important since it affects both the power quality and the dynamic performance. To obtain both low cost and high dynamic performance, a line filter consisting of a series inductor in each phase can be used. This filter is called an L-filter. To obtain sinusoidal line current with an L-filter, the switching frequency of the converter, or the inductance of the line filter, has to be high. An alternative solution for obtaining sinusoidal line currents, is to use low-pass filters, such as LC- or LCL-filters. Such filters can be designed to have a very high attenuation of harmonics due to the PWM even at moderate switching frequencies.

In most applications for grid connected PWM converters, a control system that can control the active and the reactive current independently is appropriate. At low switching frequencies, controllers based on stationary models of the line filter, such as power angle controllers are often used. At high switching frequencies, however, the ability to obtain a high bandwidth by using vector control should by utilized. In a vector control system, the active and reactive current and, consequently, the active and the reactive power, are controlled by separate controllers. As a result, a high bandwidth and a low cross coupling between the control of the active and the reactive currents can be obtained. The final performance of the control system depends on the line filter, on the switching frequency, and on the control principle. In some applications, such as the active filtering of current harmonics, the phase-shifts and the cross coupling between the control of the active and the reactive currents should be predicted in order to obtain the desired performance.

In the design of control systems for PWM converters, analytical models are important for the prediction of dynamic performance with different control laws and system parameters. Using linear time-invariant (LTI) models is most attractive since it is much more complex to design control systems for nonlinear and time-variant systems.

1.3. The Present Thesis

In applications for grid-connected voltage source converters, obtaining a high power quality and dynamic performance is important. In addition, independent control of the active and the reactive powers should be obtained.

An important goal of the thesis is to compare the power quality and dynamic performance of vector controlled PWM converters connected to the grid by an L-filter and an LCL-filter. The dynamic performance and power quality are affected by nonlinearities in the converter system. The thesis investigates different nonlinearities and their influence on dynamic performance and power quality. To reduce the complexity of the system design, linear analytical models should be used. Methods of analytical modeling and the accuracy of linear analytical models are studied by comparing frequency responses obtained from analytical models and measured small-signal frequency responses. In the active filtering of harmonic currents, phase shifts between the reference currents and the active and reactive currents, and cross coupling between the control of the active and the reactive currents reduce the performance. Methods to predict and compensate for phase-shifts are dealt with and the influence of cross coupling are displayed.

The thesis consists of two parts. In Chapters 2 and 3 of Part I, different phenomena that influence the controllability and power quality are described, and references to previous publications in the area of research are given. Furthermore, the objectives and some of the results of the publications appended in Part II are described. Chapter 2 gives an introduction to vector control of PWM converters and describes different nonlinearities. Chapter 3 deals with analytical modeling. In Chapter 4, the objectives and results of the publications in Part II of the thesis are summarized and Chapter 5 concludes the thesis work. Part II of the thesis consists of the appended publications.

Chapter 2

Control System Design and Influence of Nonlinearities

In this chapter, an introduction to vector control of grid-connected PWM converters is presented. The operating principle of a vector controlled PWM converter connected to a three-phase grid is described, as well as different parts of the converter system that introduce nonlinearities and consequently affect the performance of the converter system.

In the literature, control principles usually applied for linear systems, such as PI control [1][2], state feedback control [3][4], as well as control principles intended for nonlinear systems are proposed. Another most frequently proposed control structure is to use comparators that control the phase currents, or the active and reactive currents of a vector control system, within specified tolerance bands [5]-[7]. Such a controller, along with phase current estimators, can be used to obtain sensorless current control [7].

Among the control principles intended for nonlinear systems, neural network controllers and fuzzy logic controllers [8]-[10], are becoming popular. The control principle of neural networks is based on the learning of the control system based on the consequences of different operations. When a fuzzy logic controller is used, the operation of the converter is determined by predetermined rules. By using such a controller, the operation of the converter can be programmed to obtain high performance by using a physical knowledge of the system. As an example, it is possible to implement a vector control system without measuring the line voltages by using fuzzy logic control [9]. In some publications, other control principles for nonlinear systems, such as sliding mode control [11] and Lyaponov based control [12], are proposed.

When hysteresis controllers such as tolerance band and fuzzy controllers are used, neither parameter variation in the line filter nor errors caused by the PWM converter or by non-symmetrical or distorted grid voltages have a significant impact on the performance of the converter system. However, when hysteresis controllers are used, it is usually difficult to predict the harmonic spectra of the line currents since the switching frequency of the converter varies with the operating conditions and the amplitude of the phase voltages of the grid.

In the thesis, the objective is to investigate the performance of linear control systems that can be modeled by linear analytical models. Nonlinear controllers such as hysteresis controllers are, consequently, not studied. In the thesis, predictive vector controllers are used. Using such a controller is appropriate for obtaining both high dynamic performance and phase currents with a specified harmonic distortion. For a predictive vector controller, the dynamic performance depends on the linearity of the PWM converter, on the line filter, and on the control principle used.

2.1. The Operating Principle of a Vector Controlled PWM Converter

A principal scheme of a PWM converter with a predictive vector control system is displayed in Fig. 2.1. The control system is based on an inner loop and an outer loop. The inner loop controls the power between the dc link and the grid, while the reference current of the inner loop is provided by the outer loop and depends on the application. In Fig. 2.1, the outer loop is designed to control the dc voltage of the load. To obtain sinusoidal line currents, a line filter is connected between the converter and the grid. The line filter attenuates the voltage harmonics caused by the PWM converter and is also necessary in the control of the power between the dc link and the grid.

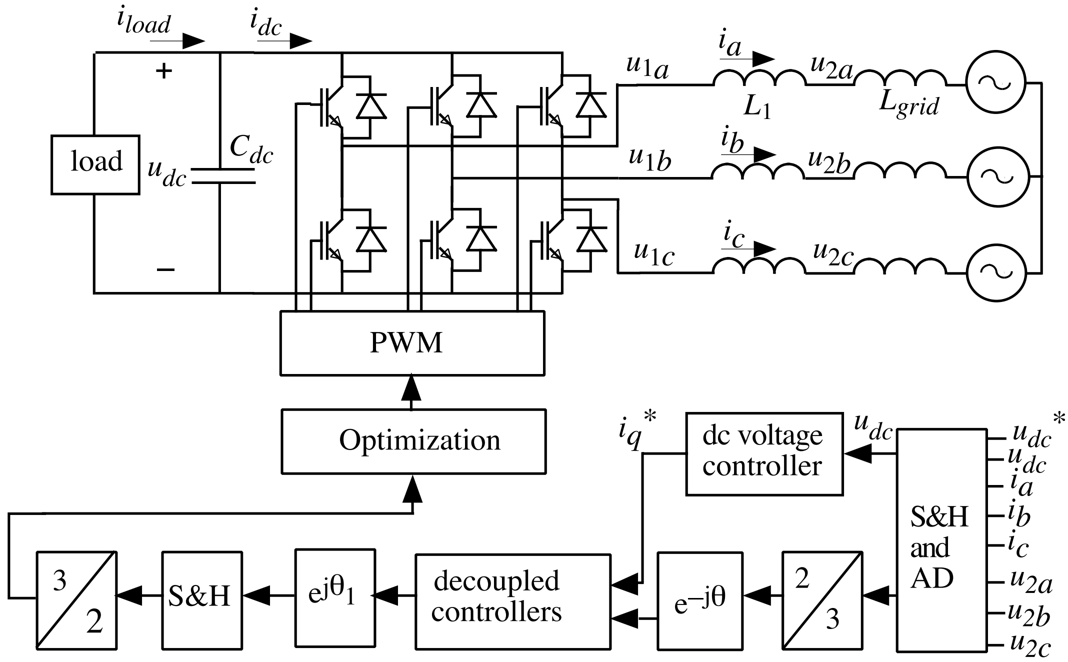

Fig. 2.1. System layout vector controlled PWM converter connected to a three-phase grid.

In PWM converters for ac applications, vector control systems can be utilized to obtain independent control of the active and the reactive powers. One of the most advantageous characteristic of a vector control system is that vectors of ac currents and voltages occur as constant vectors in steady state. Therefore, static errors in the control system can be avoided by using PI controllers. The theory for modeling three-phase systems by using vector theory was originally developed to analyze transient phenomena in ac machines [13][14]. Later on, vector theory was applied to the analysis and control of grid-connected converters [15]. Today, vector theory is widely used in the control of three-phase PWM converters.

In a vector control system, three-phase currents and voltages are described as vectors in a complex reference frame, usually called the $\alpha\beta$-frame. In addition, for a grid-connected converter, a rotating reference frame that is synchronized to the grid-flux vector is introduced; this frame is usually called the dq-frame. Since the dq-frame is synchronized to the positive-sequence fundamental part of the grid-flux vector, positive-sequence voltages and currents of the fundamental frequency occur as constant vectors in the dq-frame in steady state. In systems with a significant voltage harmonic distortion and non-symmetrical grid voltages, the control system should separate the positive-sequence voltages from the negative-sequence voltages [16], and take into account the influence of the voltage harmonics of the grid voltages [17].

In the literature, no standard definition seems to be used and the *d*-axis is used to represent the direction of the grid-voltage vector, as well as the grid-flux vector. In this thesis, the grid-voltage vector occurs in the *q*-direction, and the grid-flux vector in the *d*-direction. The angle between the α-axis of the $\alpha\beta$-frame and the *d*-axis of the *dq*-frame is used in the transformations between the $\alpha\beta$- and the *dq*-frame. In Fig. 2.1, the angle between the α-axis of the $\alpha\beta$-frame and the *d*-axis of the *dq*-frame is denoted by θ and by θ_1. θ is the angle at the time instant of the sample k, t_k. The value of the angle θ is obtained by using a synchronization method [18].

As described in Fig. 2.1, the currents and voltages used in the control system are sampled and AD converted in the block S&H and AD. In the measurements in the thesis, all currents and voltages are sampled synchronously at the same instants of time. Sampling the phase-currents synchronously is important, since a significant current ripple occurs in the phase-currents. The AD converted currents and voltages are transformed into *d*- and *q*-components in the *dq*-frame, where individual controllers for the active and the reactive powers are used.

When high sampling frequencies are used, a time delay of one sampling interval occurs in the control loop due to the AD conversion and the calculation time in the digital control system. At low switching frequencies, such as 1 kHz, it is possible to avoid the delay of one sampling interval since each sampling interval starts with the output of the same voltage vector for a long time. Normally, high sampling frequencies are used, and a time delay of one sampling interval occurs in the feedback loop. The voltage vector required by the control system during the sampling interval from sample *k*+1 to sample *k*+2 is, consequently, based on the sampled currents and voltages at the sample *k*. The reference voltages for the phase voltages of the VSC are obtained by transforming the reference voltage in the *dq*-frame back into the three-phase system. To obtain the *d*- and *q*-voltages required by the controllers on average during the sampling interval from *k*+1 to *k*+2, the angle θ_1 in the transformation from the *dq*-frame to the $\alpha\beta$-frame, is the average angle of the grid-flux vector during the sampling interval from *k*+1 to *k*+2.

A three-phase voltage is transformed into a vector in the $\alpha\beta$-frame by the transformation

$$\underline{u}_1^{\alpha\beta}(u_a,u_b,u_c) = \sqrt{\frac{2}{3}}\left(u_a + u_b e^{j\frac{2\pi}{3}} + u_c e^{-j\frac{2\pi}{3}}\right) \tag{2.1}$$

In the transformation, the factor $\sqrt{2/3}$ is introduced to obtain a power invariant transformation. Thus, the same power in the two-axis system and in the three-phase system. In the thesis, power invariant transformations are used exclusively. In the literature, amplitude invariant transformations are also commonly used. When amplitude invariant transformations are used, the phase currents and voltages can be directly obtained as the projections of current vectors and voltage vectors on the reference axes of the three phases, at the angles 0, $2\pi/3$ rad and $-2\pi/3$ rad.

The phase voltages of the VSC are defined by

$$u_{1a} = v_{1a} - v_0 \tag{2.2}$$

$$u_{1b} = v_{1b} - v_0 \tag{2.3}$$

$$u_{1c} = v_{1c} - v_0 \tag{2.4}$$

where v_0 is the potential in the neutral point, and v_{1a}, v_{1b} and v_{1c} are the phase potentials of phase a, phase b, and phase c, respectively. Thus, the voltage vector of the VSC can be written as

$$\underline{u}_1^{\alpha\beta}(u_a, u_b, u_c) = \sqrt{\frac{2}{3}}\left[v_{1a} + v_{1b}e^{j\frac{2\pi}{3}} + v_{1c}e^{-j\frac{2\pi}{3}} - v_0\left(1 + e^{j\frac{2\pi}{3}} + e^{-j\frac{2\pi}{3}}\right)\right] \tag{2.5}$$

Since PWM is used, each phase of the converter is connected to the positive or the negative side of the dc link. In analytical models, the converter can be conveniently modeled by introducing switching states. In these models, the voltage drops across the valves are usually not taken into account. Furthermore, the models assume that the converter switches instantaneously. By using the switching states s_{wa}, s_{wb}, and s_{wc} that can be equal to 1 or –1, the phase potentials of the VSC can be written

$$v_{1a} = s_{wa}\frac{u_{dc}}{2} \tag{2.6}$$

$$v_{1b} = s_{wb}\frac{u_{dc}}{2} \tag{2.7}$$

$$v_{1c} = s_{wc}\frac{u_{dc}}{2} \tag{2.8}$$

Now, the voltage vector of the VSC can be written as

$$\underline{u}_1^{\alpha\beta}(s_{wa}, s_{wb}, s_{wc}) = \sqrt{\frac{2}{3}}\frac{u_{dc}}{2}\left(s_{wa} + s_{wb}e^{\mathrm{j}\frac{2\pi}{3}} + s_{wc}e^{-\mathrm{j}\frac{2\pi}{3}}\right) \tag{2.9}$$

By introducing a switching vector, this can be simplified to

$$\underline{u}_1^{\alpha\beta} = u_{dc}\underline{sw}^{\alpha\beta} \tag{2.10}$$

where the switching vector is defined by

$$\underline{sw}^{\alpha\beta} = \sqrt{\frac{1}{6}}\left(s_{wa} + s_{wb}e^{\mathrm{j}\frac{2\pi}{3}} + s_{wc}e^{-\mathrm{j}\frac{2\pi}{3}}\right) \tag{2.11}$$

The three switches of the converter can be combined in eight ways; the resulting voltage vectors for these combinations are displayed in Fig. 2.2. A vector $\underline{u}(s_{wa}, s_{wb}, s_{wc})$ with switching states $s_{wa} = 1, s_{wb} = -1$, and $s_{wc} = -1$ is denoted by $\underline{u}(100)$.

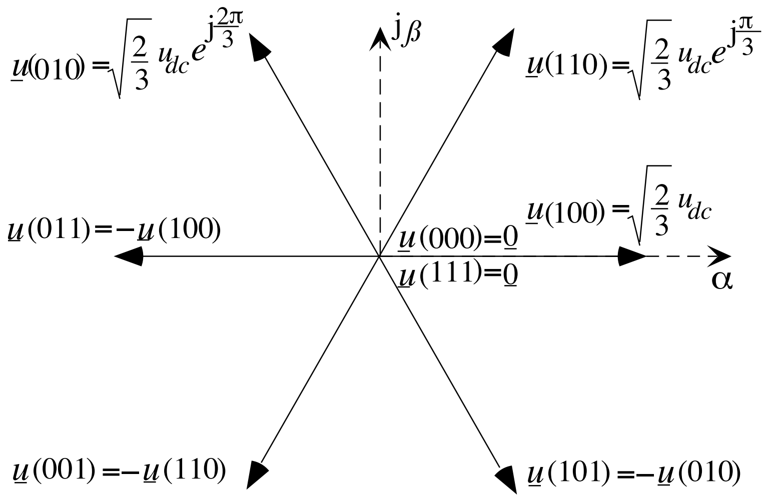

Fig. 2.2. The voltage vectors of the ideal VSC .

2.2. Influence of PWM Method and Switching Frequency

In the design of a PWM converter system with a predictive current controller, the selection of the modulation principle and the switching frequency are important for obtaining the desired performance. The switching frequency has an impact on the current harmonic distortion, the losses in the line filter and the converter valves, as well as the accuracy and the dynamic performance of the current control system. It is important that the output voltages of the converter are linear functions of the reference voltages. To obtain linear operation, the ratio between the switching frequency and the frequency of the modulated signal, called the frequency ratio, should be sufficiently high [19][20]. Moreover, due to the PWM, harmonic voltages at the converter output occur at frequencies close to multiples of the switching frequency.

If a high switching frequency is used, the harmonics caused by the PWM can be attenuated by the use of a small line filter. Furthermore, a high switching frequency makes it possible to use a high sampling frequency in the digital control system since the maximum feasible sampling frequency is twice the switching frequency.

The main drawback of a high switching frequency is that the losses in the valves become high. To avoid high losses, the semiconductor valves can be switched rapidly. However, rapid switching increases the high frequency distortion introduced by the converter into the grid and also gives rise to high voltage derivatives that may damage the insulation of the line filter connected between the converter and the grid.

An alternative solution for reducing the losses of the semiconductor valves is to use soft switching. In such a converter, usually called a resonant or quasi-resonant converter, the

commutation of the valves takes place at zero current or zero voltage [21]. Such solutions are often proposed for dc-dc converters and also for ac-dc converters [10], [22]–[24].
Another important advantage of resonant converters is that the reduced voltage derivatives of the converter output voltages reduces the high frequency distortion and also allow the use of long cables between the converter and the line filter or the machine [23].

By using a randomized PWM method, a more continuous spectrum than with a regular PWM method using a constant sampling frequency is obtained [25]. Such methods are often utilized to reduce the audible noise in ASDs. One drawback of randomized methods is that the design of the current control system becomes more complex [26]. Randomized methods are mainly proposed for inverters in ASDs.

In digitally controlled PWM converters with predictive controllers, space vector modulation (SVM) or carrier-wave based PWM is used. For SVM, the time instants at which each switching is to be performed are evaluated analytically. In a carrier-wave based method, reference voltages are compared with a triangular or sawtooth wave, and the commutations occur at the intersection of the references and the wave. When three-phase voltages are compared with a single carrier wave, a sub-oscillating carrier-wave PWM method is obtained [27]. Thus, a sub-oscillating method denotes a PWM method for three-phase systems. When a sub-oscillating method is used, an optimized PWM method should be implemented to fully utilize the dc voltage [27][28]. An important advantage of SVM is that the number of switchings during each sampling interval can be reduced in comparison with a carrier-wave based method [29].

The modulation method has an impact on the size of the line filter. The influence of different modulation methods on the line filter can be studied by using a frequency analysis of the modulation methods [30]. For each PWM method, it is important to synchronize the sampling of the phase currents to the time instants corresponding to the average values of the phase currents. When this is performed accurately, the current harmonic distortion introduced by the PWM converter is not present in the sampled phase currents and consequently not in the *d*- and *q*-currents either. Sampling at the accurate time instants is most important in systems with high ripple currents due to high current derivatives [31]. For a sub-oscillating PWM method with a triangular carrier-wave, the time instants corresponding to the average phase currents occur at the maximum and the minimum of the triangular carrier wave.

When a sub-oscillating method is used, a reference value for each phase voltage is compared with a common carrier wave, as illustrated in Fig. 2.3. In the figure, the sampling frequency is equal to the switching frequency. The voltage vectors $\underline{u}(000)$and $\underline{u}(111)$ result in the same output voltage and consequently, three different voltage vectors are active during one sampling interval.

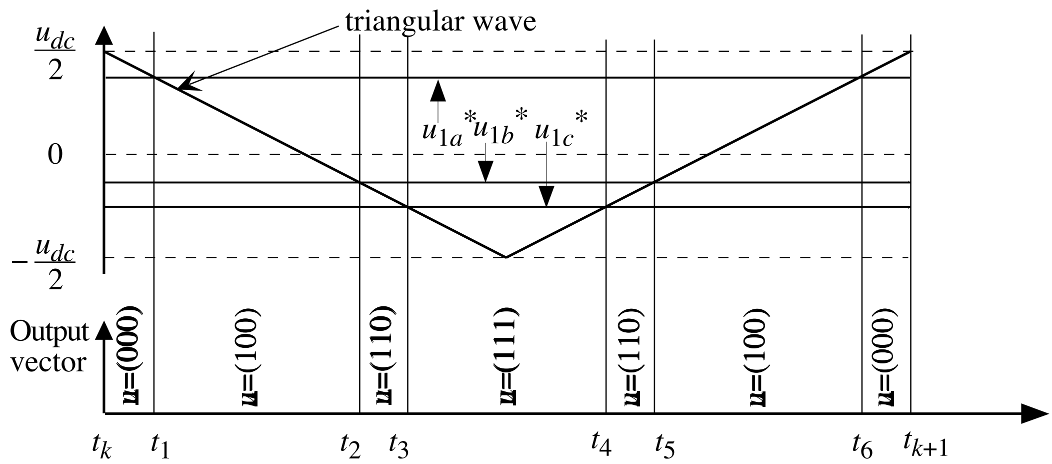

Fig. 2.3. Triangular carrier-wave based PWM and the output voltage vectors of the VSC during one sampling interval. The switching frequency is equal to the sampling frequency.

To obtain the voltage vector that is required by the vector controller, the reference vector in the dq-frame is transformed into a vector in the $\alpha\beta$-frame. Due to the uniform sampling, the reference voltage vector is constant during the sampling interval. The reference voltage vector in the $\alpha\beta$-frame is determined by

$$\underline{u}_1^{\alpha\beta *} = \underline{u}_1^{dq *} e^{j\left(\theta + \frac{3}{2}\omega_g T_s\right)} \tag{2.12}$$

where ω_g is the angular frequency of the grid and T_s is the sampling time. The transformation angle in the exponent, θ_1, is equal to the average angle during the succeeding sampling interval. In this transformation, it is also possible to compensate for small time delays of a few µs due to the gate circuits and the capacitive gates of the valves.

The reference vector required by the control system is obtained on average during the sampling interval, assuming that the d- and q-components of the voltage vectors of the VSC are constant throughout the sampling interval. Furthermore, the converter should switch according to the PWM pattern determined by the PWM circuit. However, the d- and q-components of the different voltage vectors are projections of the voltage vectors of the VSC onto the dq-frame and are, therefore, functions of the angle of the dq-frame. Consequently, due to the rotation of the d- and q-axes during the sampling intervals, time varying errors occur in the average voltage vector during the sampling intervals. The errors correspond to a nonlinearity and a coupling between the d- and q-directions in the control system.

The average d- and q-components obtained as a function of the angle of the reference voltage vector are displayed in Figs. 2.4 and 2.5 for a sampling frequency equal to the switching frequency, and for a sampling frequency at twice the switching frequency, respectively. A reference voltage vector of 0 V in the d-direction and of 400 V in the q-direction has been applied. The results have been obtained by integrating the projection of each voltage vector used during a sampling interval onto the d- and q-axes and dividing the result by the angle of one sampling interval.

According to Fig. 2.4, only minor errors occur when the sampling frequency is equal to the switching frequency. As shown in Fig. 2.5, significant errors that change signs between two sampling intervals occur if the sampling frequency is twice the switching frequency. This is unfortunate since it is important to use as high sampling frequency as possible at low switching frequencies, in order to reduce the time delay in the digital control system. However, as shown in Figs 2.4 and 2.5, the errors are functions of the angle of the reference voltage vector, and the sampling frequency, and it should, thus, be possible to compensate for the errors.

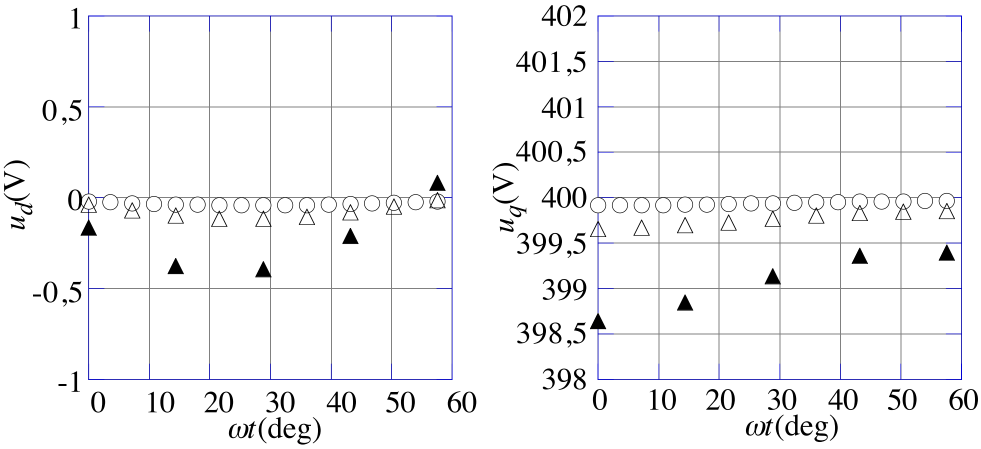

Fig. 2.4. The average d- *and* q-*components of the converter output voltage vector during the sampling intervals when the reference voltage vector rotates from 0 to 60 degrees in the* αβ-*frame. The sampling frequency is equal to the switching frequency. The sampling frequency is 1.25 kHz (solid triangles), 2.5 kHz (triangles), and 5 kHz (circles).*

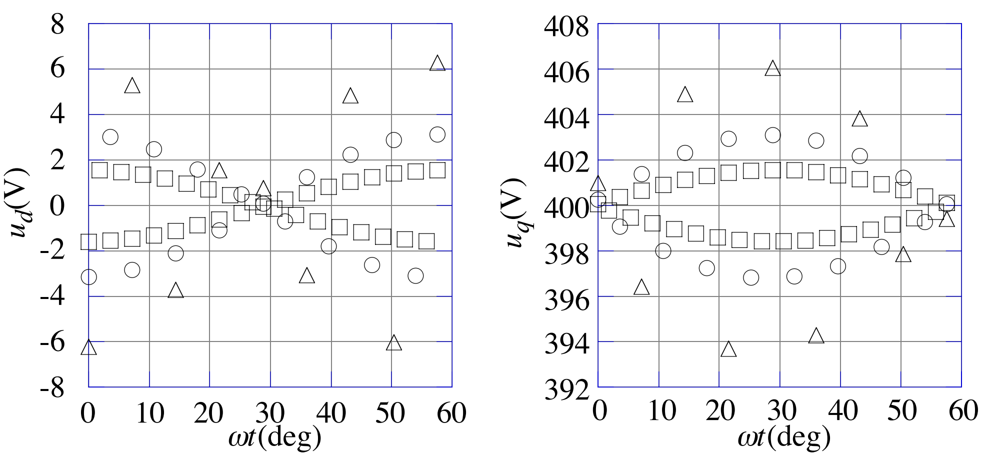

Fig. 2.5. The average d- *and* q-*components of the converter output voltage vector during the sampling intervals when the reference voltage vector rotates from 0 to 60 degrees in the* αβ-*frame. The sampling frequency is twice the switching frequency. The sampling frequency is 2.5 kHz (triangles), 5 kHz (circles) and 10 kHz (squares).*

The influence of the PWM method on the small-signal performance of PWM converters has been investigated by a few authors. An analog control system using continuous natural sampling PWM has been investigated in [17]. Natural sampling PWM denotes that the reference voltages of the PWM change during the sampling intervals; this is the case in an analog control system. In a digital control system, a uniform sampling PWM is normally used and the reference voltages are constant during the sampling intervals. In [17], the transfer functions from a reference voltage vector to an output voltage vector of an ideally switched PWM converter were investigated. Constant gain was obtained except for the frequencies where harmonics due to the PWM occur.

Analytical expressions of the spectral output for sinusoidal inputs with carrier-wave based PWM can be obtained by using the technique described in [19]; both natural sampling and uniform sampling have been considered. According to this reference, natural sampling is linear and the phase-shift is zero. With uniform sampling, the amplitude of the fundamental output is a function of the frequency ratio and, furthermore, a phase-shift is introduced. Publication [19] presents theoretical results, and consequently describes the operation of an ideally switching converter with ideal valves that switch according to the PWM pattern obtained from the modulator. Thus, only the influence of the PWM method is studied.

In [32], the linearity of a uniform space-vector PWM method was studied in the *dq*-frame. Measured and theoretical transfer functions from reference voltages in the *d*- and *q*-directions to average voltages in the *d*- and *q*-directions were displayed. These transfer functions are nonlinear and in addition time variant with the angle of the reference voltage vector. The nonlinearity varies with the ratio between the PWM frequency and the frequency of the reference voltage, and also with the amplitude of the components of the reference voltage vector. In addition to the nonlinear gain, a cross coupling between the *d*- and *q*-directions was displayed. According to the publication, uniform PWM methods are nonlinear. The measured results are, however, influenced not only by the uniform PWM method, but also by the *dq*-transformations and the errors caused by the non-ideal valves of the converter.

In the thesis, the small-signal transfer functions from current references in the *d*- and the *q*-directions to currents obtained in the *d*- and *q*-directions are presented in **Publications C**, **D** and **G**. An important objective of these publications has been to study the influence of nonlinearities and to verify whether any nonlinearity or cross coupling is introduced in comparison with the ideal performance of the vector control system. In **Publication C** and **D**, the L-type line-filter is used. In **Publication G**, transfer functions are also presented for the LCL-filter. In all publications, high performance predictive vector controllers are developed and utilized.

In **Publication C**, small-signal transfer functions for different current controllers of the P- and PI-type are presented. The transfer functions are compared with Bode-diagrams obtained from linear analytical models. No significant coupling is introduced by the PWM method and the coordinate transformations between the $\alpha\beta$- and the *dq*-frame. In the publications in the thesis, the switching frequency has been from 6 to 7.2 kHz, and the sampling frequency has been equal to the switching frequency. As displayed in Fig. 2.4, the influence of the coordinate transformations has been of minor significance.

As described in **Publicaton C**, the transfer functions are affected by losses in the line filter and by the errors introduced by the non-ideal converter.

In **Publications D**, the influence of the losses is described further by comparing operations with two different inductors. Transfer functions are presented for line filter inductors with oriented and non-oriented cores of 0.3 and 0.5 mm laminations, respectively. The publication also describes the influence of the operating point based on measured small-signal frequency responses both in inverter and rectifier operations. In **Publication G**, small-signal frequency responses are used to verify the dynamic performance of a control principle for the LCL-filter.

2.3. Nonlinearities due to the Non-ideal Converter

In the converter, nonlinearities occur due to two different phenomena: blanking time and forward voltage across valves.

The nonlinearity due to the blanking time occurs as a result of the non-instantaneous commutation of a valve. Each commutation between two valves in a phase leg is divided into two steps. First, the valve that is conducting is turned off. Since a semiconductor valve needs time to recover before it can block voltages, a blanking time is introduced between the turn-off of one valve and the turn-on of the other valve in the phase leg. If the transistor is conducting in the valve that is turned off, the diode in the valve that is to be turned on starts to conduct as soon as the transistor is turned off. Thus, the switching time instant is not influenced by the blanking time. At a commutation from a diode to a transistor, the commutation is delayed by the blanking time since the diode in the conducting valve is not affected by the gate of the valve. The delay of the commutation introduces an error

$$\Delta u_{blank} = u_{dc} \frac{T_{blank}}{T_s} \qquad (2.13)$$

between the average phase voltage during a sampling interval and the phase voltage required by the PWM modulator. T_{blank} is the blanking time and T_s is the sampling time.

Thus, a commutation from the lower valve to the upper valve is delayed by the blanking time if the phase current is positive, according to the reference direction in Fig. 2.1. A commutation from the upper valve to the lower valve is delayed by the blanking time if the current is negative. Consequently, the average voltage during a sampling interval is reduced by Δu_{blank} if the current is positive at the commutation from the lower valve to the upper valve and is increased by Δu_{blank}, if the current is negative at the commutation from the upper valve to the lower valve.

If the sampling frequency is equal to the switching frequency, two commutations occur in each phase during a sampling interval. In this case, the average phase voltage is increased by Δu_{blank} if the current is negative at both commutations, and reduced by Δu_{blank} if the current is positive at both commutations. If the current is positive at the first commutation and negative at the second commutation, errors of different signs occur and the total error due to the blanking time is zero. Similarly, if the current is negative at the first commutation and positive at the second commutation, no errors are introduced by the blanking time.

Consequently, no error occurs if the phase-current has different signs at the two commutations during a sampling interval. For a converter with IGBT valves, blanking times of a few μs are used; this is sufficient to introduce significant errors at switching frequencies above a few kHz.

In addition to the errors introduced by the blanking time, the forward voltage across the valves gives rise to an error between the reference phase voltages and the average phase voltages. The forward voltage across the transistor, u_{onT}, and the diode, u_{onD}, of the valves can be determined from the equations

$$u_{onD} = U_{D0} + R_{onD} i \tag{2.14}$$

$$u_{onT} = U_{D0} + R_{onT} i \tag{2.15}$$

where U_{D0} and U_{T0} are the no-load voltages for the diode and the transistor, respectively, and R_{onD} and R_{onT} are the resistances for the diode and the transistor, respectively.

The forward voltage across the valves varies with the rated current of the valves. In the publications in Part II of the thesis, IGBT valves with a rated current of 400 A have been used. According to the data sheets, the typical voltage across the transistor is 3 V at the rated current 400 A. The the forward voltage across the valves have been identified from measurements of the voltage across the valves for phase currents between 0 and 50 A. The resulting parameters are listed in Table I.

TABLE I

MEASURED PARAMETERS OF THE IGBT VALVES USED IN THE THESIS

U_{D0}= 1.05 V	U_{T0}= 1.08 V	R_{onD}= 10 mΩ	R_{onT}= 12 mΩ

As reported in [33], the harmonic distortion of the phase currents for electrical machines supplied by PWM inverters can be reduced by compensating for the nonlinearities caused by the blanking time and non-ideal valves. In ASDs with IGBT valves, the current ripple caused by the PWM is usually low due to high switching frequencies and due to the leakage inductance of the machine. At low ripple currents, it is sufficient to use the sign of the phase current at the samples in the compensation for errors. When converters are connected to the grid, however, the current ripple caused by the PWM may be high, especially in systems with a high transient capability and when the LCL-filter is used. In systems with a high current ripple, the sign of the phase currents at the commutations may be the opposite of the sign at the sample. In such a system, a method for predicting the sign of the phase currents at the commutations during each sampling interval should be used to compensate for the error caused by the blanking time accurately.

In the thesis, methods for compensating errors caused by the non-ideal valves and by the blanking time for grid-connected PWM converters are dealt with. The goal has been to investigate the influence of the valves and study if small-signal performance can be improved by including feedforward compensation for the errors.

In **Publication C**, it is verified that the dynamic performance can be increased and that the coupling between the *d*- and *q*-directions can be reduced by compensating for the errors caused by the blanking time and non-ideal valves. In the publication, the compensation function is not used at low currents to avoid erroneous compensation due to the current ripple caused by the PWM.

Publication D shows the influence of the compensation function at different operating points.

In **Publication E**, a compensation method based on prediction is introduced to make it possible to compensate for errors due to the blanking time even at low currents, and in systems with a high current ripple. The phase currents and the currents in the *dq*-frame are shown in the time domain with and without the compensation method based on prediction. The operation of the compensation principle is verified at different operating points, and also at low currents, where the phase currents change signs between the commutations during a large part of the line period.

2.4. Influence of Line Filter

In applications of voltage source converters, two types of line filters are mainly used: the L-filter and the LCL-filter. The L-filter is a first-order filter that is obtained by using a series inductor in each phase. With the L-filter, the switching frequency of the converter has to be high to obtain sufficient attenuation of the harmonics caused by the PWM converter at a reasonable size of the line filter. An LCL-filter is obtained by connecting capacitors in delta- or wye-connection on the line side of the L-filter. Furthermore, inductors are connected on the line side of the capacitors to stop current harmonics from parallel loads from overloading the capacitors of the line filter. The inductors on the line side of the capacitors are also used to tune the resonance frequency of the line filter. Since capacitors are used in the LCL-filter, reactive power is produced. This makes the filter most suitable for applications where the converter system is used to generate reactive power.

The LCL-filter has two main advantages compared to the L-filter. When the LCL-filter is used, the attenuation of the harmonics that are caused by the PWM converter increases at a rate of 60 dB per decade above the resonance frequency in comparison with the increase of 20 dB per decade for the L-filter. This is most advantageous since it makes it possible to obtain sinusoidal line currents even at low and moderate PWM frequencies.

The possibility of using a low inductance in the line filter is the second main advantage of the LCL-filter; if high switching frequencies are used, the total inductance of the line filter can be reduced as compared with the L-filter. In such a system, the transient performance of the converter system can be very high since the dc voltage required for performing current steps is proportional to the inductance of the line filter. When the L-filter is used, the transient response is limited by the inductance required to obtain sufficient attenuation of the harmonics caused by the PWM. The transient response at overmodulation can be optimized to obtain the fastest possible response or to avoid coupling between the *d*- and *q*-directions [34][34]. Moreover, integrator wind-up should be avoided when overmodulation occurs.

Selecting the parameters of an LCL-filter is a complicated task. The resonance frequency should be set to obtain sufficient attenuation of the harmonics caused by the PWM converter. In applications above approximately 500 kVA, the switching frequency should be below approximately 1-2 kHz with the semiconductor valves available today. Therefore, to obtain sinusoidal line currents, the resonance frequency of the filter should be set to approximately 0.5 to 1 kHz. At more moderate power levels, higher switching frequencies, such as 5 to 10 kHz can be used. In this case, a resonance frequency of a few kHz is appropriate.

The losses in the line-filter inductors, especially in the inductor connected to the converter, are affected by the switching frequency of the PWM converter. If inductors with a low inductance are connected on the converter side of the capacitors, a significant current ripple occurs, and to avoid saturation of the magnetic core, the physical size of the inductors may need to be almost as large as the size of the inductors with a higher inductance.

In addition to the static characteristics, the dynamic performance of the control system is affected by the resonance frequency since it is difficult to obtain a bandwidth above the resonance frequency. Furthermore, harmonic distortion of the grid voltages or current harmonics injected by parallel loads may initiate oscillations between the capacitors of the line filter and the inductance of the line. To avoid oscillations, the resonance frequency of the line filter should be set to a frequency with a low harmonic distortion.

When the L-filter is used, predictive dead-beat vector controllers or hysteresis controllers are usually used. To obtain dead-beat performance, compensation for the time delay of one sampling interval should be incorporated into the controllers. If no compensation is used, an oscillatory behavior is obtained at high gains such as dead-beat gain. In general, prediction methods can be used for the compensation [36][37]. Another solution is to utilize two intervals of PWM during each sampling interval. This facilitates supplying the desired voltage vector within one sampling interval [38]. However, when such a method is used, setting the sampling frequency to twice the switching frequency is not possible.

Control principles for the LCL-filter have been described in a few publications. In [2], different PWM methods and controllers for a VSC connected to the grid by LCL-filters are described. It is concluded that a modulation method with a fixed switching frequency should be used to obtain a stable system, since harmonics at frequencies close to the resonance frequency of the line filter give rise to a nonstable system otherwise.

In [10], the static characteristics of a control principle utilizing space vector modulation, as well as a fuzzy logic controller are compared. It is verified that the fuzzy controller increases the performance compared to the SVM controller. At low loads, a significant current harmonic distortion occur at frequencies of a few hundred Hz when a controller utilizing SVM is applied . These harmonics at low frequencies are significantly reduced by the use of the fuzzy controller. Another advantage of the fuzzy controller is that the calculation time in the digital control system is reduced by 30 % compared to the SVM controller. Moreover, a soft switching VSC is successfully implemented and connected to the grid by an LCL-filter in [10].

A high-performance predictive vector control principle has been presented in [28]. In [28], the controller is used as a tracking controller in the active filtering of current harmonics. Furthermore, a control principle without current sensors has been described in [39].

In this thesis, dynamic performances of converters connected to the grid by the L-filter as well as the LCL-filter are dealt with. An important goal has been to investigate the potential of obtaining a high dynamic performance also for the LCL-filter.

In **Publication C**, the dynamic performance obtained when using an L-filter and a predictive dead-beat current control system is investigated. In the publication, measure small-signal frequency responses for different current controllers of the P- and PI-type are presented.

In **Publication F**, a control principle for the LCL-filter designed to reduce the number of measurements is presented. In this control principle, the inner current loop is used to attenuate oscillations in the capacitor voltage. When this control principle is used, only the currents in the inner current loop and the line voltages are measured. Thus, the number of measurements is the same as for the L-filter. Still, rather a high dynamic performance is obtained, as shown in simulations. Moreover, a method for compensating for the time delay of one sampling interval is introduced. The method uses the prediction of the *d*- and *q*-currents to compensate for the time delay and is based on a dead-beat current response. The method for compensating for the time delay has been used in the measurements in each publication in Part II of the thesis.

In **Publication G**, the capacitor voltages are measured and used to improve the performance as compared with the principle in **Publication F**. Measured small-signal frequency responses are presented for the LCL-filter, as well as for the L-filter with a dead-beat control system. A similar dynamic performance is obtained with the LCL-filter and the L-filter. In the publication, the dynamic performance of the current control systems are also verified by the active filtering of the current harmonics generated by a parallel thyristor rectifier. Furthermore, different principles of active filtering are compared. As shown by measurements, the performance of an active filter is reduced due to phase shifts between the reference current components in the *d*- and *q*-directions and the currents in the *d*-and *q*-directions in the vector control system. The result is also affected by coupling between the *d*- and *q*-directions. As verified by the measurements, by compensating for the phase shifts in the current control system of the VSC, it is possible to implement efficient active filters even when high phase shifts occur in the vector current control system of the VSC.

Chapter 3

Analytical Modeling

In the design of control systems for PWM converters, analytical models are important tools for predicting dynamic performance and stability limits of different control laws and system parameters. This chapter deals with techniques to model vector controlled grid connected VSCs.

3.1. State Space Modeling of VSCs Connected to the Grid

The system to be modeled is described in Fig. 3.1. In the thesis, the inductance L_{grid} is assumed to be zero and the q-axis of the dq-frame is synchronized to the vector set up by the voltages u_{2a}, u_{2b}, and u_{2c}. If a significant inductance is present, the angle of the dq-frame changes rapidly at steps in the line currents, and the inductance L_{grid} should be included in the analytical models.

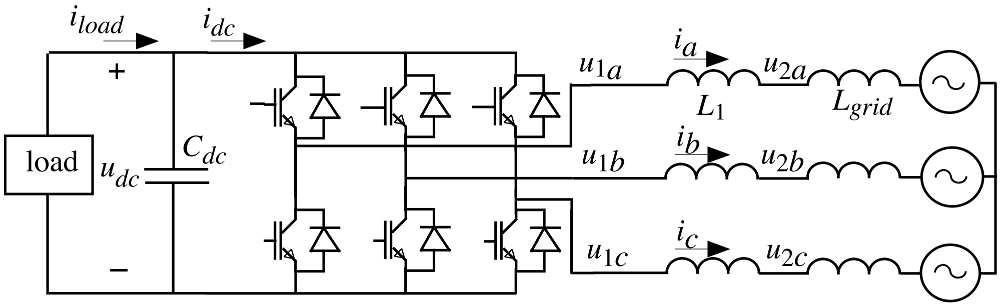

Fig. 3.1. Main circuit of the VSC connected to a three-phase grid.

The ac side of the converter system is modeled by differential equations for each phase. Assuming that the inductors are not saturated and that iron losses and copper losses due to the skin effect can be neglected, the equations for the L-filter are

$$L_1 \frac{di_a}{dt} + R_1 i_a = u_{1a} - u_{2a} \tag{3.1}$$

$$L_1 \frac{di_b}{dt} + R_1 i_b = u_{1b} - u_{2b} \tag{3.2}$$

$$L_1 \frac{di_c}{dt} + R_1 i_c = u_{1c} - u_{2c} \tag{3.3}$$

By using vector notation, these equations can be written in the $\alpha\beta$-frame

$$L_1 \frac{d\underline{i}^{\alpha\beta}}{dt} + R_1 \underline{i}^{\alpha\beta} = \underline{u}_1^{\alpha\beta} - \underline{u}_2^{\alpha\beta} \tag{3.4}$$

and in the rotating dq-frame as

$$L_1 \frac{d\underline{i}^{dq}}{dt} + \left(R_1 + j\omega_g L_1\right)\underline{i}^{dq} = \underline{u}_1^{dq} - \underline{u}_2^{dq} \tag{3.5}$$

The decoupled equation can be written in the state space form as

$$\frac{d\mathbf{x}_L}{dt} = \mathbf{A}_L\mathbf{x}_L + \mathbf{B}_L\mathbf{u}_L \tag{3.6}$$

where the state vector and the input vector are defined by

$$\mathbf{x}_L = \begin{bmatrix} i_d & i_q \end{bmatrix}^T \tag{3.7}$$

and

$$\mathbf{u}_L = \begin{bmatrix} u_{1d} & u_{1q} & u_{2d} & u_{2q} \end{bmatrix}^T \tag{3.8}$$

respectively. The system matrix and the input matrix are given by

$$\mathbf{A}_L = \begin{bmatrix} -\frac{R_1}{L_1} & \omega_g \\ -\omega_g & -\frac{R_1}{L_1} \end{bmatrix} \tag{3.9}$$

and

$$\mathbf{B}_L = \begin{bmatrix} \frac{1}{L_1} & 0 & -\frac{1}{L_1} & 0 \\ 0 & \frac{1}{L_1} & 0 & -\frac{1}{L_1} \end{bmatrix} \tag{3.10}$$

The state space equation for the ac side of the system is linear for the L-filter. The same applies for the LCL-filter as shown in **Publication G.**

The dc side of the system is modeled by the equation

$$C_{dc}\frac{du_{dc}}{dt} = i_{load} - i_{dc} \tag{3.11}$$

The current in the dc link can be found from the power of the ac side since the power on the dc side must be equal to the power on the ac side of the converter. Here, the losses in the valves are neglected. The instantaneous power on the ac side and the dc side of the converter can be obtained from

$$p_{ac} = \mathrm{Re}\left\{\underline{u}_1^{\alpha\beta}\,\underline{i}_1^{\alpha\beta*}\right\} \tag{3.12}$$

and

$$p_{dc} = u_{dc}i_{dc} \tag{3.13}$$

respectively. As shown in Chapter 2, the voltage vector of the converter can be described by a switching function. By inserting (2.10) into (3.12), and setting (3.12) equal to (3.13), the dc current is obtained as

$$i_{dc} = \mathrm{Re}\left\{\underline{sw}^{\alpha\beta}\underline{i}_1^{\alpha\beta*}\right\} \tag{3.14}$$

which is equal to

$$i_{dc} = sw_\alpha i_\alpha + sw_\beta i_\beta \tag{3.15}$$

where sw_α and sw_β are the α- *and β*-components of the switching vector $\underline{sw}^{\alpha\beta}$. The dc current can also be written in *dq*-coordinates as

$$i_{dc} = sw_d i_d + sw_q i_q \tag{3.16}$$

As a result, a complete state space equation in the *dq*-coordinates for the VSC connected to the grid by the L-filter, can be formed by adding the first order equation

$$C_{dc}\frac{du_{dc}}{dt} = i_{load} - sw_d i_d - sw_q i_q \tag{3.17}$$

to the state space equation (3.6). Thus, the system order is increased by one when the dc side is included in the model. The system is both nonlinear and time variant since switching functions that are functions of the states occur in the system matrix. This is a main drawback since it is much more complex to evaluate the stability and dynamic performance of nonlinear systems [40]. However, a linearized small-signal model can be used to obtain the transfer-function matrix of the linearized system [17]. In the linearized model, the PWM converter is assumed to be linear and the switching behavior of the converter is omitted. As a result, the output vector of the converter is assumed to have a constant amplitude and rotate at a constant angular frequency in steady state. In this case, the components of the switching vector, sw_d and sw_q, are constants and the system can, thus, be linearized in the operating point sw_{d0} and sw_{q0} of the switching vector.

In [17], transfer functions for VSCs are derived for linearized models of the VSC connected to the grid. Transfer function matrices are derived both in the *αβ*-frame and in the *dq*-frame. The poles of the transfer functions of the linearized model are significantly affected by the operating point. Therefore, the Bode diagrams or the eigenvalues of the linearized system model should be studied at no load and at full load rectifier as well as inverter operations.

The modeling of the closed loop system is a complicated task due to the multiple-input multiple-output (MIMO) structure of the system. In [17], the control system for the dc voltage of a VSC is designed by the use of the transfer function matrices of the linearized system. In this publication, an analog control system is modeled. Therefore, the time delay occurring in a predictive digital control system is not present. Still, the transfer functions become complicated. The gains in the control system were tuned to obtain sufficient amplitude and gain margins in the operating point with the lowest gain and amplitude margins. The stability of nonlinear control systems should, however, be studied by the use of methods for nonlinear systems in order to guarantee stability not only for small-signal pertubations in the vicinity of the operating point, usually denoted by global stability [40]. Such a technique is utilized in [12].

In the thesis, the dynamic performance of different current control systems are studied by using state space equations. An important objective is to study if the dynamic performance of the current control system can be predicted by using Bode-diagrams of linear analytical models. In each publication in Part II of the thesis, the dc voltage is assumed to be constant or slowly varying, and the equation for the dc voltage is not included in the state space models of the converter system. In the measurements, variations in the dc voltage are compensated in the reference voltages of the PWM. As a result, small variations in the dc voltage do not affect the performance of the current control system.

In **Publication B**, the modeling of the closed-loop converter system by means of a discrete state space equation is described. In the publication, the time-delay of one sampling interval is conveniently taken into account by modeling the converter voltage vector and the time delay of one sampling interval by using state variables. The reference voltage vector from the control system is assumed to be obtained on average during each sampling interval, with a time delay of one sampling interval. Consequently, the errors introduced by the blanking time and the non-ideal valves, as well as the errors introduced by the coordinate transformations are neglected. The state vector of the line filters is expanded with two states, one state for each component of the voltage vector. As a result, the transfer functions for the closed-loop system conveniently can be obtained by the z-transformation of the discrete state space equation, or directly by using the Matlab TM function DBODE, intended for state space models.

In **Publications C**, **D** and **G,** measured small-signal frequency responses are compared with Bode diagrams obtained from an analytical model created by the technique described in **Publication B**. The small-signal frequency responses are affected by the blanking time and the non-ideal valves, and also by the losses in the line filter, still, the measured frequency responses at different operating points are very similar to the Bode diagrams obtained from the linear analytical models. The linearity of the system, however, relies on the linearity of the PWM and that the errors due to coordinate transformations are small. Therefore, the results may be different at lower sampling frequencies.

Chapter 4

Summary of Appended Publications

The publications appended in Part II of the thesis can be divided into three sections. Section one deals with the line interference of PWM converters. In section two, analytical modeling and the influence of nonlinearities on the dynamic performance is on focus. Section three deals with control techniques. In the following, the objectives and the results of the appended publications are summarized.

4.1. Line Interference of PWM Voltage Source Converters

The objective of the publication "Connecting Fast Switching Voltage Source Converters to the Grid — Harmonic Distortion and its Reduction" (**Publication A**) are to describe the consequences of high switching speeds, and to compare the attenuation of the harmonics caused by the PWM converter when the L-filter or the LCL-filter is used. Fast-switching voltage source converters may give rise to significant over voltages at the load connected to the converter. This is verified by measurements of transients at an asynchronous machine connected to the converter by cables of different lengths. Such transients can be avoided by a reduction of the switching speed. As described in the publication, a third-order line filter can be used to obtain a high power quality even at low to moderate switching frequencies.

4.2. Analytical Modeling

In the publication "Analysis and Simulation of Digitally Controlled Grid-connected PWM Converters using the Space-vector Average Approximation" (**Publication B**), a modeling technique based on discrete state-space modeling is presented. The technique is most useful in modeling of the closed loop converter system with a time-delay of one sampling interval due to the sampled control system. By modeling the converter system in the rotating *dq*-frame, a linear time-invariant analytical model is obtained. In the model, the converter is modeled by using the space-vector average approximation. The publication shows the influence of the space-vector approximation by using simulations.

The objectives of the publication "Influence of Non-linearities on the Frequency Response of a Grid-connected Vector-controlled VSC" (**Publication C**) are to study the influence of nonlinearities on the performance of the control system and the accuracy of linear analytical models. It is most difficult to verify the dynamic performance of PWM converter systems in the time domain. Therefore, a small-signal analysis is used for this purpose. Small-signal transfer functions of gain and coupling for the vector current control system are displayed. In the evaluation of the dynamic performance of the current control system, frequencies up to 1 kHz are considered. The accuracy of the analytical models are described by comparing measured small-signal frequency responses with Bode diagrams from analytical models. One conclusion from the transfer functions is that space-vector averaging can be used in modeling the investigated closed-loop converter system.

In the literature, losses in the line filter and their effect on dynamic performance are usually not described. In **Publication C**, it is also shown that dynamic performance is affected by the losses in the line-filter inductors. Furthermore, it is verified that an improved dynamic performance and reduced coupling between the control of the *d*- and *q*-components of the current vector can be obtained by compensating for nonlinearities caused by the blanking time and non-ideal semiconductor valves.

The influence of the losses in the line-filter is further described in the publication "Vector Current Controlled Grid Connected Voltage Source Converter — Influence of Nonlinearities on the Performance" (**Publication D**). Measured small-signal frequency responses are presented for two different line-filter inductors. As shown, a more ideal response is obtained by using inductors with low iron losses. To verify that the LTI analytical model is valid for different operating points, small-signal frequency responses are presented at four different operating points representing rectifier, as well as inverter, operation. At each operating point, experimental results are presented with and without the compensation for the nonlinearities caused by the blanking time and non-ideal semiconductor valves. As reported, the compensation function improves the performance at each operating point. The small-signal responses are similar at the different operating points. An important result of this is that the performance of current control systems can be predicted by the analytical model both in inverter as well as in rectifier operation.

4.3. Control Techniques

As shown in **Publications C** and **D**, the dynamic performance of PWM converters can be improved by compensating for the nonlinearity caused by the blanking time and non-ideal valves. Such a compensation is based on the sign of the phase currents. When the phase currents are close to zero, the sign of the current at the commutations may be different from the sign at the samples, and an accurate compensation principle has to apply a prediction method to compensate for the error caused by the blanking time.

In the publication "A Method to Compensate for Errors Caused by the Blanking Time in PWM Systems with High Ripple Currents" (**Publication E**), a method for predicting the signs of the phase currents at the commutations is introduced. Experimental results are presented with and without the compensation method. The focus is on the line-current distortion and static errors introduced in the vector control system. As shown in the publication, the static errors caused by the blanking time and non-ideal valves can be avoided by using the compensation method. The principle makes it possible to apply compensation even at low phase currents, and when small line filters or low switching frequencies resulting in high current ripples are used.

In power electronic systems, the number of measurements contributes to the cost of the system. With the principle proposed in the publication "Feed Forward — Time Efficient Control of a Voltage Source Converter Connected to the Grid by Lowpass Filters" (**Publication F**), only one third of the states are measured. Still, rather a high bandwidth can be obtained. The principle is an alternative to the reduction of measurements by using estimation. When the principle is used, the number of measurements in the control system become the same as for the L-filter. The publication also presents a method for compensating for the delay time of one sampling interval introduced by the sampled control system.

In the publication "Control of a Voltage-source Converter Connected to the Grid through an LCL-filter — Application to Active Filtering" (**Publication G**), a control principle for the LCL-filter is introduced. The dynamic performance of a converter connected to the grid through two different line filters is compared. As shown in measured frequency responses, the dynamic performance obtained when the LCL-filter is used is similar to that of the L-filter. The current harmonic distortion in the line currents, due to the PWM, is significantly lower for the LCL-filter.

Publication G also describes the influence of the active filtering principle on the control system. When a direct method is used, the active filter tries to compensate for all harmonics based on the instantaneous active and reactive currents. When such a method is used, high derivatives occur in the control system for the active filter. As shown in measurements, these derivatives saturate the controller in the *q*-direction.

By using a Fourier method, the saturation can be avoided solely by only including the low-frequency components in the references for the active filter. In addition, a flexible compensator is obtained since individual harmonics can be compensated separately. Furthermore, when the Fourier method is used, the compensation for phase shifts in the current control system is done in the frequency domain. Another advantage obtained by the use of the Fouriermethod is that dead-beat control does not have to be applied, which is the case if the phase shifts in the current control system are compensated in the time domain. The compensation can also be based on measured frequency responses, which makes it feasible to apply individual compensation in the *d*- and *q*-directions.

By compensating for phase shifts, it is possible to operate even at low switching and sampling frequencies, which is a major advantage at high power levels and for the LCL-filter.

Chapter 5

Conclusions

As shown in this thesis, PWM converters can be used to obtain a high power quality and dynamic performance of ac to dc converters. Compared with line-commutated converters, the harmonic distortion of the line currents is very low and it is fair to say that sinusoidal line currents can be obtained. Another important advantage of PWM converters as compared with line-commutated converters is that the active and reactive powers can be independently controlled, thus a unity power factor can be obtained. Moreover, by using high performance vector control systems, PWM converters can be used in new applications, such as active filtering, where the controllability of the PWM converter is utilized to compensate for harmonic currents that are injected into the grid by nonlinear loads.

The power quality of a PWM converter depends on the switching frequency and on the line filter that is connected between the converter and the grid. As shown in the thesis, it is advantageous to use LCL-filters since sinusoidal line currents can be obtained even at low and moderate switching frequencies, such as 5 to 7 kHz. It is also verified that a high dynamic performance can be obtained with the LCL-filter. A control principle for the LCL-filter resulting in a high dynamic performance is developed. Experiments have shown that the small-signal dynamic performance is rather close to the small-signal dynamic performance when using a dead-beat controller for the L-filter. One drawback of the LCL-filter as compared with the L-filter is that the control system becomes more expensive if each current and voltage are used in the control. To avoid an increased cost, a method resulting in the same number of measurements as with the L-filter has been developed.

The dynamic performance of a vector control system is affected by nonlinearities that may occur due to the PWM method, due to coordinate transformations in the control system, and due to the non-ideal converter. These nonlinearities depend on the switching and sampling frequency. In the thesis, switching frequencies from 5 to 7 kHz are used. At such switching and sampling frequencies, nonlinearities and cross couplings caused by the uniform PWM method and by coordinate transformations, do not have a significant effect of the small-signal frequency responses for frequencies below 1.2 kHz.

The nonlinearities due to the blanking time between the switch-off of one valve in a phase leg, and the turn-on of the other valve in a phase leg, introduce an error that affects the small-signal dynamic performance. In the thesis, a method for compensating for the blanking time and non-ideal vales is developed. As shown in experimental results, a more ideal small-signal frequency response is obtained by using the compensation method.

Moreover, in the measured frequency responses, small errors are introduced by losses in the line filter. Experimental results show that the nonlinearity due to the losses in the line filter is reduced when inductors with lower iron losses are used.

A method to model switching PWM converters by using discrete state space models is described and utilized in the design of control principles for the L-filter as well as the LCL-filter. As shown in Bode diagrams and measured small-signal frequency responses, the small-signal dynamic performance can be predicted rather accurately from the Bode diagrams obtained from analytical models.

In a vector control system used for the active filtering of harmonic currents, the performance of the active filter is affected by the phase shifts in the current control system, and by the cross coupling between the control of the active and the reactive currents. If a direct method is used, it is possible to compensate for the phase shifts if a dead-beat performance is obtained in the current control system of the PWM converter. By using a Fourier method, it is possible to compensate for the phase shifts even when a dead-beat performance is not obtained. This is most advantageous since it makes active filtering possible even when the LCL-filter is used. Furthermore, the compensation can be based on measured small-signal frequency responses. By compensating for phase shifts, it is possible to obtain high-performance active filtering even at moderate switching frequencies.

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Part II
Appended Publications

Publication A

M. Lindgren, J. Svensson, "Connecting Fast Switching Voltage Source Converters to the Grid — Harmonic Distortion and its Reduction," *IEEE Stockholm Power Tech Conference*, Stockholm, Sweden, June 18-22, 1995, Proceedings of Power Electronics, p. 191-195.

Edited version.

Publication B

M. Lindgren, "Analysis and Simulation of Digitally Controlled Grid-connected PWM-converters using the Space-vector Average Approximation, " *IEEE 5th Workshop on Computers in Power Electronics*, Portland, USA, August 11-14, 1996, p. 157-162.

Edited version.

Publication C

J. Svensson, M. Lindgren, "Influence of Non-linearities on the Frequency Response of a Grid-Connected Vector-Controlled VSC," to appear in *IEEE Transactions on Industrial Electronics*.

Publication D

J. Svensson, M. Lindgren, "Vector Current Controlled Grid Connected Voltage Source Converter — Influence of Non-linearities on the Performance," *IEEE Power Electronics Specialists Conference (PESC 98)*, Fukuoka, Japan, May 17-22, 1998, Vol. 1, p. 531-537.

Edited version.

Publication E

M. Lindgren, J. Svensson, "A Method to Compensate for Non-ideal Commutation and Valves in PWM-systems with High Ripple Currents," *1998 IEEE Nordic Workshop on Power and Industrial Electronics* (*NORPIE/98*), Espoo, Finland, August 26-27, 1998, p. 147-152.

Edited version.

Publication F

M. Lindgren, "Feed forward — Time Efficient Control of a Voltage Source Converter Connected to the Grid by Lowpass filters," *IEEE Power Electronics Specialists Conference (PESC 95)*, Atlanta, USA, June 18-22, 1995, Vol. 2, p. 1028-1032.

Edited version.

Some of the equations have been corrected as compared with the original paper.

Publication G

M. Lindgren, J. Svensson, "Control of a Voltage-source Converter Connected to the Grid through a LCL-filter — Application to Active Filtering," *IEEE Power Electronics Specialists Conference (PESC 98)*, Fukuoka, Japan, May 17-22, 1998, Vol. 1, p. 229-235.

Edited version.