Liudmila Smirnova

ELECTROMAGNETIC AND THERMAL DESIGN OF A MULTILEVEL CONVERTER WITH HIGH POWER DENSITY AND RELIABILITY

Thesis for the degree of Doctor of Science (Technology) to be presented with due permission for public examination and criticism in the Auditorium 1382 at Lappeenranta University of Technology, Lappeenranta, Finland on the 13th of August, 2015, at noon.

Acta Universitatis
Lappeenrantaensis 651
Electric energy demand has been growing constantly as the global population increases. To avoid electric energy shortage, renewable energy sources and energy conservation are emphasized all over the world. The role of power electronics in energy saving and development of renewable energy systems is significant. Power electronics is applied in wind, solar, fuel cell, and micro turbine energy systems for the energy conversion and control. The use of power electronics introduces an energy saving potential in such applications as motors, lighting, home appliances, and consumer electronics.

Despite the advantages of power converters, their penetration into the market requires that they have a set of characteristics such as high reliability and power density, cost effectiveness, and low weight, which are dictated by the emerging applications. In association with the increasing requirements, the design of the power converter is becoming more complicated, and thus, a multidisciplinary approach to the modelling of the converter is required.

In this doctoral dissertation, methods and models are developed for the design of a multilevel power converter and the analysis of the related electromagnetic, thermal, and reliability issues. The focus is on the design of the main circuit. The electromagnetic model of the laminated busbar system and the IGBT modules is established with the aim of minimizing the stray inductance of the commutation loops that degrade the converter power capability. The circular busbar system is proposed to achieve equal current sharing among parallel-connected devices and implemented in the non-destructive test set-up. In addition to the electromagnetic model, a thermal model of the laminated busbar system is developed based on a lumped parameter thermal model. The temperature and temperature-dependent power losses of the busbars are estimated by the proposed algorithm. The Joule losses produced by non-sinusoidal currents flowing through the busbars in the converter are estimated taking into account the skin and proximity effects, which have a strong influence on the AC resistance of the busbars.

The lifetime estimation algorithm was implemented to investigate the influence of the cooling solution on the reliability of the IGBT modules. As efficient cooling solutions have a low thermal inertia, they cause excessive temperature cycling of the IGBTs. Thus, a reliability analysis is required when selecting the cooling solutions for a particular application. The control of the cooling solution based on the use of a heat flux sensor is proposed to reduce the amplitude of the temperature cycles.
The developed methods and models are verified experimentally by a laboratory prototype.

Keywords: multilevel converters, neutral point clamped converter, electromagnetic modelling, busbars, inductance, thermal analysis, reliability, lifetime estimation
Acknowledgements

This work has been carried out at the department of Electrical Engineering, School of Energy Systems at Lappeenranta University of Technology, Finland, between 2011 and 2015. First of all, I would like to thank supervisor of this dissertation, Professor Juha Pyrhönen, for his guidance, interesting discussions, and support during these years. I would also like to acknowledge my supervisors, Professor Pertti Silventoinen and Professor Olli Pyrhönen, for their ideas and suggestions.

I would like to express my gratitude to the honoured preliminary examiners of this doctoral dissertation, Associate Professor Michal Frivaldský from University of Žilina and Dr. Veli-Matti Leppänen from ABB Oy, for their time and effort in evaluating my work. I appreciate the valuable comments and suggestions you have provided.

I express special thanks to Mr. Raimo Juntunen, Mr. Tatu Musikka, Mrs. Elvira Baygildina, Dr. Maria Polikarpova, Mr. Kirill Murashko for the valuable contributions to this work. I thank Professor Andrey Mityakov for possibility to use the GHFS in my research. I am grateful to colleagues from Department of Mechanical Engineering collaborating with me during these years: Dr. Mika Lohtander, Mr. Simo Valkeapiiä, Mr. Tapani Siivo, Mr. Antti Jortikka, Mr. Leevi Paajanen. For creating a positive work environment in our office and help in the laboratory, I thank Dr. Vesa Väisänen, Mr. Jani Hiltunen, and Mr. Joonas Talvitie. It has been a pleasure to work with you. Many thanks to laboratory and workshop personnel Mr. Martti Lindh, Mr. Jouni Ryhänen, Mr. Kyösti Tikkanen for their help in building and testing the prototype.

I also thank Dr. Hanna Niemelä for her efforts in providing me assistance and invaluable comments on my writing and grammar of this dissertation and my papers. Special thanks to Mrs. Piipa Virkki and Mrs. Tarja Sipiläinen for their help and cheerful attitude when organising working process, business trips, and defence of this dissertation. I would like to acknowledge Dr. Julia Vauterin-Pyrhönen and Dr. Pia Lindh for support in educational process. The financial support of Walter Ahlström foundation is highly appreciated.

I would like to thank all people working with me during exchange period at Aalborg University: Dr. Ke Ma, Mr. Rui Wu, Dr. Yan Liu, Dr. Chandrasekaran Subramanian, Dr. Huai Wang. In particular, I am grateful to Professor Frede Blaabjerg, Professor Francesco Iannuzzo and his wonderful family. You have made my stay fruitful and pleasant.

I am very grateful to all friends being there for me during these years. Thank you for sauna evenings with board games, discussions at coffee breaks, picnics, birthday parties, various sport activities, trips, etc! Thanks to you these years were full of moments worth remembering.

Я хочу выразить огромную благодарность моей большой семье. Несмотря на то, что вы далеко, я всегда чувствую вашу поддержку и любовь. Special thanks to my mother Antonina and sister Elena for believing in me and encouraging throughout my whole life.
Finally, I would like to thank my husband Alexander for love and understanding. Without your support this thesis would not have been possible.

Liudmila Smirnova
July 2015
Lappeenranta, Finland
## Contents

Abstract

Acknowledgements

Contents

Nomenclature

1 Introduction 13
   1.1 Multilevel converters ............................................................. 16
      1.1.1 Operation principles of NPC and ANPC converter ............ 17
      1.1.2 Modulation methods ....................................................... 19
      1.1.3 Power semiconductors ................................................... 20
   1.2 Motivation ......................................................................... 23
      1.2.1 Reliability ................................................................. 23
      1.2.2 Power density .......................................................... 25
      1.2.3 Cost effectiveness ...................................................... 28
   1.3 Objective of the work .......................................................... 29
   1.4 Outline of the work ............................................................ 31
   1.5 Scientific contributions and publications ............................ 32

2 Low-inductive design of the converter 35
   2.1 Stray inductance of the converter commutation loops ........... 35
   2.2 Theory of partial and loop inductances ................................. 36
   2.3 Design of the laminated busbar system ................................ 41
      2.3.1 Selection of materials .................................................. 42
      2.3.2 Laminated structure ...................................................... 44
      2.3.3 Location of the components ......................................... 46
   2.4 Inductance estimation of the laminated busbar system .......... 48
      2.4.1 Partial inductance estimation ....................................... 48
      2.4.2 Loop inductance estimation ......................................... 51
      2.4.3 Experimental verification .......................................... 52
      2.4.4 Detailed model of the IGBT module ............................. 53
   2.5 Summary ........................................................................... 57

3 Busbar system for the NDT set-up 59
   3.1 Description of the NDT set-up .......................................... 59
   3.2 Busbar system of NDT set-up I ......................................... 61
   3.3 Busbar system of NDT set-up II: current-sharing issues ....... 62
   3.4 Summary ........................................................................... 64

4 Thermal analysis of the laminated busbar system 65
   4.1 Power losses and temperature estimation ............................ 66
Nomenclature

Roman Letters

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
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<td>magnetic vector potential</td>
<td>Vs/m</td>
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<tr>
<td>B</td>
<td>magnetic flux density</td>
<td>T</td>
</tr>
<tr>
<td>b</td>
<td>distance</td>
<td>m</td>
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<tr>
<td>C</td>
<td>capacitance</td>
<td>F</td>
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<tr>
<td>C</td>
<td>thermal capacitance</td>
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<tr>
<td>C_{LC}</td>
<td>life cycle cost</td>
<td>€</td>
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<tr>
<td>CL</td>
<td>consumed lifetime</td>
<td>%</td>
</tr>
<tr>
<td>c</td>
<td>specific heat capacity</td>
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<tr>
<td>d</td>
<td>thickness</td>
<td>m</td>
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<td>ΔT_j</td>
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<tr>
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<td>temperature difference between case and heat sink</td>
<td>K</td>
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<tr>
<td>ΔT_{(h-a)}</td>
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<tr>
<td>ΔT_{(j-c)}</td>
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<tr>
<td>E</td>
<td>energy</td>
<td>J, kW·h</td>
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<tr>
<td>E_{max}</td>
<td>dielectric strength</td>
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<tr>
<td>E_{sw}</td>
<td>switching energy</td>
<td>J</td>
</tr>
<tr>
<td>EPB</td>
<td>energy payback time</td>
<td>years</td>
</tr>
<tr>
<td>e</td>
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<td>f</td>
<td>frequency</td>
<td>Hz</td>
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<tr>
<td>g</td>
<td>gravitational acceleration</td>
<td>m/s²</td>
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<td>h_c</td>
<td>convection heat transfer coefficient</td>
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<td>I</td>
<td>rms current</td>
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<td>I_{CM}</td>
<td>peak collector current of the IGBT</td>
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<td>maximum controllable turn-off current of the IGCT</td>
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<tr>
<td>i</td>
<td>current</td>
<td>A</td>
</tr>
<tr>
<td>k</td>
<td>thermal conductivity</td>
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<td>nut factor</td>
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<tr>
<td>L_p</td>
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<td>l</td>
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</tr>
<tr>
<td>LT</td>
<td>lifetime</td>
<td>years</td>
</tr>
<tr>
<td>M</td>
<td>mutual-partial inductance</td>
<td>H</td>
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<tr>
<td>N_{cyc}</td>
<td>number of thermal cycles</td>
<td></td>
</tr>
<tr>
<td>P</td>
<td>power, loss</td>
<td>W</td>
</tr>
<tr>
<td>P_{cond}</td>
<td>conduction loss</td>
<td>W</td>
</tr>
<tr>
<td>P_{out}</td>
<td>output power</td>
<td>W</td>
</tr>
<tr>
<td>P_{sw}</td>
<td>switching loss</td>
<td>W</td>
</tr>
<tr>
<td>q</td>
<td>heat flux</td>
<td>W/m²</td>
</tr>
<tr>
<td>R</td>
<td>electrical resistance</td>
<td>Ω</td>
</tr>
<tr>
<td>R</td>
<td>thermal resistance</td>
<td>K/W</td>
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## Nomenclature

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<th>Symbol</th>
<th>Description</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_b$</td>
<td>resistance of a busbar</td>
<td>$\Omega$</td>
</tr>
<tr>
<td>$R_{CC'EE'}$</td>
<td>lead resistance of an IGBT module</td>
<td>$\Omega$</td>
</tr>
<tr>
<td>$R_c$</td>
<td>constriction resistance</td>
<td>$\Omega$</td>
</tr>
<tr>
<td>$R_{\text{cont}}$</td>
<td>contact thermal resistance</td>
<td>K/W</td>
</tr>
<tr>
<td>$R_{(c-h)}$</td>
<td>thermal resistance between case and heat sink</td>
<td>K/W</td>
</tr>
<tr>
<td>$R_f$</td>
<td>film resistance</td>
<td>$\Omega$</td>
</tr>
<tr>
<td>$R_{(h-a)}$</td>
<td>thermal resistance between heat sink and ambient</td>
<td>K/W</td>
</tr>
<tr>
<td>$R_{(j-c)}$</td>
<td>thermal resistance between junction and case</td>
<td>K/W</td>
</tr>
<tr>
<td>$R_t$</td>
<td>total contact resistance</td>
<td>$\Omega$</td>
</tr>
<tr>
<td>$r$</td>
<td>radius</td>
<td>m</td>
</tr>
<tr>
<td>$S$</td>
<td>area</td>
<td>m$^2$</td>
</tr>
<tr>
<td>$S_c$</td>
<td>contact area</td>
<td>m$^2$</td>
</tr>
<tr>
<td>$S_o$</td>
<td>sensitivity of the GHFS</td>
<td>V/W</td>
</tr>
<tr>
<td>$T$</td>
<td>temperature</td>
<td>K</td>
</tr>
<tr>
<td>$T$</td>
<td>torque</td>
<td>N·m</td>
</tr>
<tr>
<td>$T_a$</td>
<td>ambient temperature</td>
<td>K</td>
</tr>
<tr>
<td>$T_b$</td>
<td>temperature of a busbar</td>
<td>K</td>
</tr>
<tr>
<td>$T_c$</td>
<td>case temperature</td>
<td>K</td>
</tr>
<tr>
<td>$T_h$</td>
<td>heat sink temperature</td>
<td>K</td>
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<tr>
<td>$T_j$</td>
<td>junction temperature</td>
<td>K</td>
</tr>
<tr>
<td>$T_m$</td>
<td>mean temperature</td>
<td>K</td>
</tr>
<tr>
<td>$T_s$</td>
<td>surface temperature</td>
<td>K</td>
</tr>
<tr>
<td>$t$</td>
<td>time</td>
<td>s</td>
</tr>
<tr>
<td>$V$</td>
<td>volume</td>
<td>m$^3$</td>
</tr>
<tr>
<td>$V_{\text{conv}}$</td>
<td>volume of converter</td>
<td>m$^3$</td>
</tr>
<tr>
<td>$U_c$</td>
<td>capacitor voltage stress</td>
<td>V</td>
</tr>
<tr>
<td>$U_{CE,0}$</td>
<td>collector-emitter threshold voltage</td>
<td>V</td>
</tr>
<tr>
<td>$U_{\text{CES}}$</td>
<td>collector-emitter voltage of the IGBT</td>
<td>V</td>
</tr>
<tr>
<td>$U_{\text{DC}}$</td>
<td>DC link voltage</td>
<td>V</td>
</tr>
<tr>
<td>$U_{\text{DRM}}$</td>
<td>repetitive peak off-state voltage of the IGCT</td>
<td>V</td>
</tr>
<tr>
<td>$U_{\text{spike}}$</td>
<td>voltage spike</td>
<td>V</td>
</tr>
<tr>
<td>$W_m$</td>
<td>energy stored in the magnetic field</td>
<td>J</td>
</tr>
<tr>
<td>$w$</td>
<td>width</td>
<td>m</td>
</tr>
<tr>
<td>$x$</td>
<td>x-coordinate</td>
<td>m</td>
</tr>
<tr>
<td>$y$</td>
<td>y-coordinate</td>
<td>m</td>
</tr>
<tr>
<td>$z$</td>
<td>z-coordinate</td>
<td>m</td>
</tr>
<tr>
<td>$Z$</td>
<td>thermal impedance</td>
<td>K/W</td>
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## Greek Letters

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<thead>
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<th>Symbol</th>
<th>Description</th>
<th>Unit</th>
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<tr>
<td>$\alpha$</td>
<td>thermal diffusivity</td>
<td>m$^2$/s</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>coefficient of resistivity variation with temperature</td>
<td></td>
</tr>
</tbody>
</table>
Nomenclature

\( \beta \)  
coefficient of thermal expansion  \( 1/K \)

\( \varepsilon \)  
emissivity

\( \mu \)  
magnetic permeability  \( \text{H/m} \)

\( \nu \)  
kinematic viscosity  \( \text{m}^2/\text{s} \)

\( \rho \)  
electrical resistivity  \( \Omega \cdot \text{m} \)

\( \rho \)  
density  \( \text{kg/m}^3 \)

\( \tau \)  
time constant  \( s \)

\( \nu \)  
velocity  \( \text{m/s} \)

\( \phi \)  
magnetic flux  \( \text{Vs} \)

Dimensionless numbers

Gr  Grashof number
Nu  Nusselt number
Pr  Prandtl number
Ra  Rayleigh number
Re  Reynolds number

Abbreviations

2D  two-dimensional
3D  three-dimensional
AC  alternating current
Al-Cap  aluminium electrolytic capacitor
ANPC  active neutral point clamped
CAD  computer-aided design
CFD  computational fluid dynamics
CFR  constant failure rate
CHB  cascaded H-bridge
CSP  concentrating solar power
CTE  coefficient of thermal expansion
DC  direct current
DF  dissipation factor
EMI  electromagnetic interference
EPBT  energy payback time
ESR  equivalent series resistance
FC  flying capacitor
FEM  finite element method
GHFS  gradient heat flux sensor
HID  high-intensity discharge
HVAC  heating, ventilation, and air conditioning
HVDC  high-voltage DC
IEEE  Institute of Electrical and Electronics Engineers
Nomenclature

IGBT insulated gate bipolar transistor
IGCT integrated gate-commutated thyristor
LCOE levelized cost of energy
LED light-emitting diode
LPTM lumped parameter thermal model
LSPWM level-shifted pulse-width modulation
MMC modular multilevel converter
MoM method of moments
NDT non-destructive testing
NPC neutral point clamped
PEEC partial element equivalent circuit
PoF physics-of-failure
PSPWM phase-shifted pulse-width modulation
PV photovoltaic
p.u. per unit
PWM pulse-width modulation
RMS root mean square
SHE selective harmonic elimination
SVM space vector modulation
THD total harmonic distortion
TI turbulence intensity
TIM thermal interface material
1 Introduction

Electric energy demand has been growing constantly and will continue to grow for decades to come as the global population increases and an increasing number of people are getting access to electric systems and are aspiring to higher living standards. The global energy demand may increase by 50% by the middle of the century. Almost 80% of this increase is expected to come from developing countries (International Energy Agency, 2013). More and more of the electric energy production must come from renewable sources such as solar radiation and wind. In both cases, power electronics has an enabling role, and therefore, efficient and reliable power electronics is increasingly needed in the future energy systems.

Figure 1.1 shows the world’s primary energy supply by fuels (International Energy Agency, 2013). At the moment, around 82% of the total energy is generated by fossil fuels (oil, coal/peat\(^1\), and natural gas) and 5% by nuclear fuels, which also are regarded as non-renewable energy sources. Considering the current rate of consumption and present availability, the depletion curves of fossil and nuclear fuels are presented in Figure 1.2 (Bose, 2013). There is an imminent need to shift toward renewable electric energy sources such as solar and wind energy. As a result, also electric energy storages are widely needed. It is somewhat frustrating but also challenging that the sun is shining more energy to the earth than we could ever consider needing. Nevertheless, mankind is still incapable of solving its energy problems based solely on the sun. Therefore, new technology is needed.

![Figure 1.1. World primary energy supply by fuel in 2011 (International Energy Agency, 2013). Renewable sources include hydro, biofuels, and waste as well as solar, wind, and heat power.](image)

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\(^1\) Peat may be regarded as semi-fossil as it renews in thousands of years instead of millions of years.
In order to put off the approaching energy shortage, renewable energy sources and energy saving are now getting much emphasis all over the world. Renewable sources are not presented in Figure 1.2, because theoretically, the depletion curve will extend over a very long time outside the range in the figure. However, it is impossible to supply energy only by renewable sources because of the statistical nature of these sources, which requires backup power from fossil or nuclear power plants or a final solution to energy storing. In the case of electric energy storing, second-, hour-, week-, and month-level solutions are needed to change over to solar economy. Supercapacitors and lithium batteries can help in short-time storing of electricity, but so far, long-term electric energy storages seem to be only a dream. However, for instance the latest E to Gas project launched at Lappeenranta University of Technology (LUT) may provide a solution for the long-time storing of summertime solar power. In this project, extra electricity is converted into methane, which can then be used normally during times of no wind or sunshine.

Intensive research is conducted to find cost-effective energy storages, which could be used together with renewable sources to enhance the security of energy supply. Development of renewable sources will help to preserve non-renewable energy sources.

Turning away from fossil fuels will take decades. Meanwhile, energy efficiency provides an opportunity to immediately decrease energy consumption. This will bring energy savings and reduce the environmental impact of energy generation. Energy efficiency is therefore directly related to fuel poverty alleviation.

The role of power electronics in energy saving and the development of renewable energy systems is significant. By using power semiconductor devices operating in the switching mode, the efficiency of a power electronic apparatus may approach 98%–99% (Bose, 2013). In 2009, the world record 99.03% efficiency of a PV system inverter has been reported (Wilhelm et al., 2010).
Wind, solar, fuel cell, and micro turbine energy systems apply power electronics for the conversion and control of electrical energy. Electricity generated by these systems is converted into a form suitable for connection to the grid or an autonomous load by a power-electronics-based converter.

Figure 1.3. Global electricity demand by application (World Energy Council, 2013). The total electricity consumption was 22202 TWh in 2011.

Three sectors accounts for 70% of the total electricity consumption:

- motors – 40%
- lighting – 19%
- home appliances and consumer electronics – 24%

The use of power electronics in these three sectors introduces an energy saving potential. Motor-driven systems are the largest consumers of industrial electric energy, and they consume 40% of the world electricity (Figure 1.3). Motors can naturally be found also in electric vehicles (3% of consumption), and part of household electric energy (13%) is also consumed in motors. Their main electromechanical applications in industry (and households) are fans, pumps, and compressors, which are used for the control of fluid or gas flow. The majority of motors are run at a fixed speed, and the flow control is achieved by an inefficient throttle or damper opening, which wastes a lot of energy. Meanwhile, variable-speed drives are more efficient for the flow control. The main part of the variable-speed drive is the power electronic converter, which allows supplying variable frequency and amplitude voltage to the motor in order to vary its speed according to the load demands. Despite the strong evidence of attainable savings, only 10%–15% of all industrial motors presently use variable-speed drives (Waide and Brunner, 2011).

In lighting, which accounts for 19% of energy consumption, the trend is toward replacement of inefficient incandescent bulbs by more efficient fluorescent, high-intensity discharge (HID), low-pressure sodium, high-brightness LED lamps, and finally,
LED lighting. For these lighting technologies, power electronics is the enabling technology (Popović-Gerber et al., 2012).

The presence of power electronics and variable-speed drives in home appliances and consumer electronics, which consume 24% of the world’s electricity, is based on performance advantages, energy savings, and cost reduction of the power electronics. A higher comfort level for building inhabitants is achieved by the replacement of the turn on–turn off cycling control in refrigeration and heating, ventilation, and air conditioning (HVAC) equipment by more efficient variable-speed drives. Considerable energy savings are attained by high-efficient induction and microwave cooking equipment.

1.1 Multilevel converters

Multilevel power converters have been under research and development and also in high-voltage industrial use for more than three decades because of their lower switch power losses, harmonic distortion, $\frac{du}{dt}$ values, and common-mode voltage and current in comparison with traditional two-level counterparts. The history of the multilevel converters started in the early 1970s when the cascaded H-bridge (CHB) converter was first introduced (McMurray, 1971). In the same year, the concept of the flying capacitor (FC) topology for low power was introduced (Dickerson and Ottaway, 1971) and developed into the medium voltage FC converter in the early 1990s (Meynard and Foch, 1992). The diode-clamped converter, which later evolved into the three-level neutral point clamped (NPC) converter, was introduced in 1980 (Baker, 1980). The most recently emerged modular multilevel converter (MMC) was presented in 2002 (Marquardt, 2002).

Although there are various other multilevel converters, these four converters (Figure 1.4 (a) – (d)) are the most advanced and matured ones. While the FC converter has not found wide application in industry because of the high expenses of flying capacitors (Rodriguez et al., 2007) and the application of the MMC converter is limited to high-voltage DC (HVDC) transmission systems (Saeedifard and Iravani, 2010), the NPC and CHB converters are the most commonly used ones in industry nowadays.

The advantages of the CHB converter are the ability to reach high voltage and power levels using low-voltage IGBTs and the modularity and fault-tolerant operation. However, it requires an expensive phase-shifting transformer to supply each cell.

The three-level NPC converter is studied in this doctoral dissertation because it has a simpler structure resulting in a smaller size, and further, it is more applicable to back-to-back regenerative applications than the CHB converter, which requires three-phase two-level back-to-back converters for each cell to achieve the regenerative option. Although the NPC converter can be extended to a higher number of levels, these converters are seldom found in industry mainly because of the increase in the number of clamping diodes, which have to be connected in series to block the higher voltages (Franquelo et al., 2008) (Wu, 2005). In addition, the uneven distribution of losses in the outer and inner
devices and the complicated neutral point voltage balancing make these converters less attractive.

![Multilevel converters](image)

Figure 1.4. Multilevel converters (a) Cascaded H-bridge (CHB) converter. (b) Three-level flying capacitor (FC) converter. (c) Three-level neutral point clamped (NPC) converter. (d) Modular multilevel converter (MMC) with half-bridge and full-bridge sub-modules.

### 1.1.1 Operation principles of NPC and ANPC converter

Currently, NPC converters are found in various applications from low-power low-voltage to high-power and medium-voltage ones. The main idea in all neutral-point-clamped topologies is that the phase output is clamped to the DC link neutral point through semiconductor switches. The drawback of the NPC converter (Figure 1.5 (a)) is the uneven loss distribution between the inner and outer switching devices, which leads to...
In three-level active NPC (ANPC) converter introduced in (Bruckner and Bemet, 2001), the neutral clamping diodes are replaced by clamping switches as presented in Figure 1.5 (b) to provide a controllable path for the neutral point current and thereby control the distribution of losses among the devices. Initially, the clamping switches were proposed to guarantee an equal voltage sharing between the main switches without balancing resistors (Xiaoming et al., 1999). The clamping switches are also preferred from the standardization point of view and widely used in industry. If active clamping is not required, the clamping switches are turned off, and only clamping diodes are used. The clamping switches introduce new switch states (Table 1.2) and commutations compared with the NPC converter (Table 1.1).

### Table 1.1. Switch states of the three-level NPC converter.

<table>
<thead>
<tr>
<th></th>
<th>$S_{x1}$</th>
<th>$S_{x2}$</th>
<th>$S_{x3}$</th>
<th>$S_{x4}$</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>State P</strong></td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td><strong>State NP</strong></td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td><strong>State N</strong></td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>

### Table 1.2. Switch states of the three-level ANPC converter.

<table>
<thead>
<tr>
<th></th>
<th>$S_{x1}$</th>
<th>$S_{x2}$</th>
<th>$S_{x3}$</th>
<th>$S_{x4}$</th>
<th>$S_{x5}$</th>
<th>$S_{x6}$</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>State P</strong></td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td><strong>State NPU1</strong></td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td><strong>State NPU2</strong></td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td><strong>State NPL1</strong></td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td><strong>State NPL2</strong></td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td><strong>State N</strong></td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
</tbody>
</table>

Figure 1.5. One phase leg of an NPC converter with commutation loops. (a) Three-level NPC converter. (b) Three-level ANPC converter.
In many practical applications, straight commutation between P and N states is prohibited, or at least it is infrequent, and thus, commutations from P or N to NP, or vice versa, will be studied in this work. Moreover, it can be assumed that the upper and lower halves of the phase arm commutate symmetrically, and therefore, discussion of one of these will suffice.

When the switch state is changed from P to NP in the NPC converter with a positive output current direction, that is, from the DC link to the phase output, the current commutates from IGBT $S_{x1}$ to clamping diode $D_{x5}$, and commutation loop A in Figure 1.5 is established. Later on, this loop is called the “short commutation loop”. If the output current direction is reversed, the commutation takes place between freewheeling diode $D_{x1}$ and IGBT $S_{x3}$, and thus, commutation loop B is generated. This loop can be called the “long commutation loop” (Brückner, 2005).

Despite the considerable number of different commutations in the ANPC converter that can be used to distribute the switching losses among the switching devices, two basic commutation loops can be determined. When a commutation takes place between P and NPU1 or NPU2, commutation loop A is produced (Figure 1.5 (b)). If the commutation occurs between P and NPL1 or NPL2, commutation loop B is formed.

Parasitic inductance of the conductors and components along with a commutation loop generate a case-specific commutation inductance. Turning-off the power device and cutting-off an inductive current cause a voltage spike over this power device. The most common problem caused by these spikes is a switching component overvoltage breakdown. In Chapter 2, methods to minimize the stray inductance of the commutation loops are presented.

1.1.2 Modulation methods

Modulation methods are used with a primary target of generating a stepped waveform close to a reference signal with a variable frequency and amplitude. By a proper modulation method, almost sinusoidal output current of the converter can be achieved in the steady state. Other objectives such as neutral point voltage balancing, rejection of specific harmonics, common-mode voltage elimination, and loss minimization can also be achieved by advanced modulators.

Nowadays, there are three methods applied to three-level NPC converters: carrier-based PWM, space vector modulation (SVM), and selective harmonic elimination (SHE) (Rodriguez et al., 2010). The carrier-based PWM methods, divided into level-shifted (LSPWM) and phase-shifted (PSPWM) ones, have found successful industrial applications because of the simple way to relate the carrier signal with the gating signals of the NPCs. For the implementation of a carrier-based PWM, only the reference and carrier signals and a comparator are required to generate the gating signals. The SHE method having the advantage of a low number of commutations per cycle and thereby low switching losses is also widely adopted in industry. In this method, the switching
angles are computed offline and designed to eliminate particular harmonics. However, since the angles are computed based on the assumption of sinusoidal steady-state voltages, this method is limited to applications with a low dynamic performance (Kouro et al., 2010). The SVM method has been under research and development over the past decades because of its ability to use the redundant switch states of the multilevel converter to achieve the targets such as neutral point voltage balance, loss minimization, and common-mode voltage elimination, which are handled by an external controller when the carrier-based PWM or SHE methods are used. The practical implementation of the SVM method requires an algorithm with at least three stages: selection of the switch states or vectors for modulation, computation of the duty cycles of each vector, and choosing the sequence in which vectors are generated (Wu, 2005). Moreover, it has been demonstrated in (Leon et al., 2010) that the voltage waveforms generated by the most commonly used SVM methods can be obtained by a carrier-based PWM method in a much simpler way.

1.1.3 Power semiconductors

Selection of the power semiconductor in essence determines its design and performance along with the investment and operating costs of power converters. The switching devices for the converter are different for different voltage classes. According to the IEEE Standard 141-1993 (IEEE, 1994), the following system voltage classes are defined: low voltage—a class of nominal system voltages less than 1000 VAC; medium voltage—a class of nominal system voltages equal to or greater than 1000 VAC and less than 100 000 VAC; high voltage—a class of nominal system voltages equal from 100 000 VAC to 230 000 VAC. All voltages are root-mean-square phase-to-phase or phase-to-neutral voltages.

The dominant devices employed in the low-voltage and low-power converters are metal-oxide-semiconductor field-effect transistors (MOSFETs) capable of effectively operating at high frequencies and reducing the size of the converter passive components (Kroposki et al., 2010). Unfortunately, at higher voltages, MOSFETs suffer from large conduction losses, which limits their application. However, a new generation of MOSFETs based on wide bandgap semiconductor materials such as silicon carbide (SiC) and gallium nitride (GaN) are expected to solve this problem and extend the voltage and power ranges of the MOSFETs. Thus, SiC MOSFETs have recently been commercialized up to 1.2 kV 100 A, and the prototype of a 1.2 kV 800 A device has been presented in the literature (Millan et al., 2014).

In the medium-voltage class, usually below 6000 V, the IGCT, IGBT press-pack, and IGBT modules are currently feasible solutions for the voltage source converter. The nominal voltage and current ratings of the power devices available in the market are shown in Figure 1.6. For the comparison, the IGCTs are characterized by the repetitive peak off-state voltage \(U_{DRM}\) and the minimum controllable turn-off current \(I_{TGQM}\) and the IGBTs by the maximum collector-emitter voltage \(U_{CES}\) and the peak collector current \(I_{CM}\). The maximum blocking voltage of the IGBT and IGCT currently available is 6.5 kV; however, samples of 10, 18, and 40 kV IGBT and 10 kV IGCT devices have been tested
in the laboratories of several manufacturers (Kaufmann and Zwick, 2002), (Zorngiebel et al., 2009), (Ohkami et al., 2007), (Bernet et al., 2003).

The performance of the switches is compared by different aspects in a qualitative way in Table 1.3. The IGBTs are more suitable for medium-voltage class applications, where a high switching frequency and a small output filter are required as in grid-side converters. Again, the IGCTs are a preferred solution for applications with a low switching frequency and dominant conduction losses as in AC motor drives (Senturk, 2011).
IGBTs are available in two different packages: press-pack and module. The main advantages of the press-pack package are their improved reliability and high power density. The improved reliability is due to the absence of the bond wires and solder joints, and consequently, the failures associated with them. However, the mounting of a press-pack IGBTs is more complicated and expensive in comparison with an IGBT module. The high power density of the press-pack is achieved as a result of the low thermal

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>IGBT press-pack</th>
<th>IGBT module</th>
<th>IGCT</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Cost</strong></td>
<td>High</td>
<td>Moderate</td>
<td>High</td>
</tr>
<tr>
<td><strong>Failure mode</strong></td>
<td>Short-circuit</td>
<td>Open-circuit</td>
<td>Short-circuit</td>
</tr>
<tr>
<td><strong>Snubber circuit</strong></td>
<td>Recommended</td>
<td>Not required</td>
<td>Limits $di/dt$ and short circuit current</td>
</tr>
<tr>
<td><strong>Maintenance</strong></td>
<td>Complicated</td>
<td>Easy</td>
<td>Complicated</td>
</tr>
<tr>
<td><strong>Thermal resistance</strong></td>
<td>Small</td>
<td>Moderate</td>
<td>Small</td>
</tr>
<tr>
<td><strong>Switching frequency</strong></td>
<td>&lt; 1 kHz</td>
<td>&lt; 2 kHz</td>
<td>&lt;500 Hz</td>
</tr>
<tr>
<td><strong>Switching loss</strong></td>
<td>Higher than comparable IGBT module</td>
<td>Lower than comparable IGBT press-pack</td>
<td>Higher than IGBT press-pack and modules</td>
</tr>
<tr>
<td><strong>Conduction loss</strong></td>
<td>Higher than comparable IGBT module</td>
<td>Lower than comparable IGBT press-pack</td>
<td>Lower than IGBT press-pack and modules</td>
</tr>
<tr>
<td><strong>Gate driver</strong></td>
<td>Low power (typically 5 W per device)</td>
<td>Low power</td>
<td>Medium power (typically 100 W per device)</td>
</tr>
<tr>
<td><strong>Manufacturers</strong></td>
<td>ABB, Westcode</td>
<td>ABB, Infineon, Semikron, Dynex, Fuji, Mitsubishi, Hitachi</td>
<td>ABB</td>
</tr>
<tr>
<td><strong>Cooling</strong></td>
<td>Double-side expensive cooling (with deionized water), the use of thermally and electrically conductive grease on the contacting interfaces</td>
<td>Simple, mounting on isolated heat sink</td>
<td>Double-side expensive cooling with deionized water</td>
</tr>
<tr>
<td><strong>Mounting</strong></td>
<td>Complex and expensive mounting in stack, cleaning of the contacting interfaces prior to assembly</td>
<td>Simple, mounting on an isolated heat sink</td>
<td>Complex and expensive mounting in stack, cleaning of the contacting interfaces prior to assembly</td>
</tr>
<tr>
<td><strong>Reliability</strong></td>
<td>High, no bond wires and solder joints</td>
<td>Moderate, thermo-mechanical stress of bond wires and chip and substrate solder joints</td>
<td>High, no bond wires and solder joints</td>
</tr>
<tr>
<td><strong>Power density</strong></td>
<td>High</td>
<td>Moderate</td>
<td>High</td>
</tr>
</tbody>
</table>
resistance, which allows obtaining high current values without exceeding the maximum junction temperature. In (Ma and Blaabjerg, 2012) it is shown that the module solution has a better loss performance than the press-pack solution, and the difference in the thermal resistance between the two solutions comes from the thermal resistance from the case to the heat sink, which is, in practice, formed by the thermal resistance of the grease. However, the trend in the cooling systems of the IGBT modules is toward integrated cooling solutions where the grease between the IGBT module and the cooling system is eliminated (Bhunia et al., 2007), (Morozumi et al., 2013). In this case, the thermal resistances of the IGBT press-pack and module are almost equal. With this in mind, the IGBT modules are selected for the NPC converter not least because they do not require a complicated and expensive double-side cooling system and avoid mounting difficulties. The structure of the NPC converter power stage would be very complicated in the case of press-pack components.

1.2 Motivation

Despite the advantages of using power converters, the penetration of them into the market requires that they have a set of characteristics such as a low THD and level of electromagnetic interference (EMI), high quality of input and output current and voltage, efficiency, power density, reliability, and cost effectiveness. Some of these characteristics were translated into national and international standards while others are application specific. Thus, recently emerged applications (such as hybrid/electric vehicles, wind turbines, PV panels) are driving the development of power converters in the direction of higher reliability and power density, cost effectiveness, and reduction in weight (Popović-Gerber et al., 2012).

1.2.1 Reliability

According to (IEEE, 2010), reliability is defined as the ability of an item to perform a required function under stated conditions for a stated period of time. The demand for the converter with a higher reliability and a longer lifetime comes from the high penetration of devices operating in harsh and remote conditions such as offshore wind turbines, PV panels, and electric or hybrid vehicles, where power converters are more prone to failures. Table 1.4 presents the required lifetime of the power converter in different applications (Wang et al., 2014b). If a product fails within the warranty period, the replacement and repair costs are covered by the manufacturer of such a device. This will negatively affect the profits and also reflect on the manufacturer’s reputation. This is a reason for the high requirements applicable to the power converter reliability. Introducing reliability analyses at the design stage is an important step in taking corrective action, ultimately leading to a product that is more reliable.

Since the emergence of the reliability engineering discipline in the 1950s, two approaches of reliability prediction have been developing: 1) an empirical one, based on empirical data and various handbooks; 2) a physics-of-failure (PoF) one, focused on the modelling
of physical causes of the component failures. While the empirical approach considers the device as a box of components with constant failure rates given in handbooks, the PoF approach considers the device as a box of failure mechanisms. Until the 1980s, the constant failure rate (CFR) models were dominant for the prediction of the device lifetime. However, already in the 1990s when electronics devices became more complicated, the CFR models were declared inadequate for the reliability prediction (White and Bernstein, 2008). Thereafter, the PoF approach has started to play a more important role in the reliability engineering. In the PoF approach, the root cause of a dominant failure mechanism of a device is studied and corrected to achieve some determined lifetime.

Table 1.4. Typical lifetime target in different power electronic applications (Wang et al., 2014b).

<table>
<thead>
<tr>
<th>Applications</th>
<th>Typical design target of lifetime</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aircraft</td>
<td>24 years (100,000 hours of flight operation)</td>
</tr>
<tr>
<td>Automotive</td>
<td>15 years (10,000 operating hours, 300,000 km)</td>
</tr>
<tr>
<td>Industry motor drives</td>
<td>5–20 years (60,000 hours at full load)</td>
</tr>
<tr>
<td>Railway</td>
<td>20–30 years (10 hours of operation per day)</td>
</tr>
<tr>
<td>Wind turbines</td>
<td>20 years (24 hours of operation per day)</td>
</tr>
<tr>
<td>Photovoltaic plants</td>
<td>5–30 years (12 hours per day)</td>
</tr>
</tbody>
</table>

As shown in the survey presented in Figure 1.7 (Yang et al., 2010), in a power converter, the power semiconductor modules and the DC link capacitors are known to be the components most prone to failures. In this work, the focus is on the reliability analysis of the power module, which is considered to be the most critical component causing up to 34% of the power converter failures (semiconductors – 21%, solder joints – 13%). The results of an industry-based survey provided by (Yang et al., 2011) also indicated that the semiconductor modules are the most fragile components and have the highest failure/cost ratio (failure cost divided by original system cost). The DC link capacitor takes 30% of the power converter failures as shown in Figure 1.7, and voltage, current ripple, and ambient temperature are the dominant stressors for capacitors as indicated in (Wang et al., 2014b). Even though the reliability analysis of the DC link capacitor is not covered in this work, the low-inductive busbar system that connects the DC link capacitors with the power semiconductors is designed as presented in Chapter 2 of this doctoral dissertation in order to reduce the risk of the capacitor overvoltage.
The temperatures and temperature swings are known to be the dominant stressors of semiconductor devices leading to failures such as bond wire lift-off and fatigue of solder layers and ceramics (Ciappa, 2001), (Wang et al., 2014b). Because temperature has a prevalent influence on the reliability of the converter, electrothermal analyses are required to conduct a reliability study by taking a PoF approach. The thermal model of the devices and a properly defined mission profile are of primary importance for an accurate lifetime estimation of the power converter.

### 1.2.2 Power density

Power density is known to be a good figure of merit of a power converter and a measure of the progress in the power converter technology. Over the last four decades, an exponential growth of the converter power density has been observed as presented in Figure 1.8. The continuous need for a higher power density of converters and a higher level of integration may invoke new challenges in cooling, packaging, and passive component technologies.
1 Introduction

Figure 1.8. Change in the power density for inverters (Heldwein and Kolar, 2009).

Power density is defined as a ratio between the rated output power $P_{\text{out}}$ and the volume $V_{\text{conv}}$ of the converter. In order to maximize the power density of the converter, the power capability should be maximized while the converter volume is minimized. The power capability of the converter is limited by the power semiconductor current and voltage limits and thermally limited by the maximum allowed junction temperature (Senturk et al., 2012).

It is known that safety margins are used to take into account the effect of the parasitic components of the converter. Stray inductance introduces a voltage spike during the switch-off of the IGBT, and a voltage reserve prevents the switch damage. Stray inductance may result in a decrease in the realizable switching current. If the inductance is decreased, a higher current can be switched off without destroying the IGBT, and thus, the power capability can be increased using an IGBT with the same ratings. In this doctoral dissertation, methods to minimize the stray inductance of the converter commutation loops are considered. A model is developed to estimate the inductance value that allows virtual testing of the converter before it is built.

The thermal limits can be extended by adopting an efficient cooling system; however, the penalty of the cooling system is the additional cost, size, weight, and potential new failures in the cooling system and IGBT modules. In this work, the influence of the cooling system on the lifetime of the IGBT module is analysed. The improvement in converter efficiency is significant for power density because it requires a loss decrease. Again, lower losses lead to a smaller cooling system. Further, new power devices with a higher allowed junction temperature and lower losses will lead to a power density
increase. A lot of publications present converters with a high power density based on SiC devices (Rabkowski et al., 2012), (Puqi et al., 2013). The lower losses are achieved in new devices by increasing their speed. However, with the increased switching speed, the voltage spikes caused by stray inductances are higher. In modern power devices, the stray inductance is a limiting factor for their switching speeds. New low-inductive packaging technology is required to make the devices feasible for very high switching speeds. An alternative solution can be application of soft-switching topologies, which allows decreasing the device stresses and obtaining high switching frequency by minimizing or eliminating the device switching losses (Hua and Lee, 1995). However, this approach requires auxiliary circuits with passive and/or active components, which increase the complexity and cost of the power converter. The driving and control circuits also become more complicated. The above-mentioned drawbacks prevent their wide commercial application (Bellar et al., 1998).

The volume of the converter should be minimized in order to increase the power density. Two of the main factors that influence the volume of the converter are the cooling system and passive components, where the filter takes a significant part of the total volume. In order to minimize the size of the output filter, the switching frequency should be increased; however, this will introduce extra semiconductor switching losses, inductor core and winding losses, and dielectric losses (Kassakian and Jahns, 2013), which require a larger cooling system to dissipate these losses. Therefore, the optimum should be found in the design (Figure 1.9).

In order to minimize the volume of the converter, efficient packaging is needed. However, EMC issues arise in the converter with a high switching frequency of the semiconductors and densely packaged components, and should be considered already at the design stage to eliminate severe interference (Grobler and Gitau, 2013).
In practice, there is a trade-off between the power capability and the volume, and when one wants to maximize the power capability, the volume is also increased. Thermal management and electromagnetic effects must be considered simultaneously with the electrical design.

1.2.3 Cost effectiveness

A study of economic feasibility is needed in order to show the importance and value of power electronics. One method to assess the economic feasibility is to calculate the energy payback time \((EPB)\) of the system where the power converter is applied. The \(EPB\) is analogous to economic payback time \((PB)\), where the investments and economic value are defined in terms of energy. The energy payback time of a power electronic converter \((EPB_{PE})\) allows weighting the energy saving benefits per year \(E_{sav}\) achieved by the usage of power electronics against the embodied energy \(E_{emb}\) (energy required for manufacturing, installation, maintaining, and recycling of the power electronics) (Popović-Gerber et al., 2011)

\[
EPB_{PE} = \frac{E_{emb}}{E_{sav}}. \tag{1.1}
\]

This method emphasizes the importance of the converter efficiency for \(EPB\) minimization and is applied for the assessment of variable-speed drive and renewable energy systems. As the decrease in the embodied energy is usually obtained by sacrificing the efficiency, power density, or reliability of the converter, it is not an appropriate way to minimize the \(EPB\).

For the renewable energy systems such as wind turbines, PV panels, and fuel cells, which have a high investment cost, the levelized cost of energy \((LCOE)\) has also been introduced. It takes into account the cost of energy over the whole life cycle of the system (Hallam and Contreras, 2015)

\[
LCOE = \frac{C_{LC}}{E_{tot}}, \tag{1.2}
\]

where \(C_{LC}\) is the life cycle cost and \(E_{tot}\) is the energy delivered over the whole life cycle of the system.

The \(LCOE\) is used to evaluate the competitiveness of different electricity-generating technologies. Thus, the US Department of Energy in the course of the SunShot concentrating solar power (CSP) program has defined the target to reduce the \(LCOE\) of
CSP to $0.06/kWh or less as it is shown in Figure 1.10 (US Department of Energy Facilities, 2014). This goal requires reducing the \( LCOE \) of power electronics to $0.01/kWh. Recently, the \( LCOE \) began to be used as an objective function for the design optimization of a PV converter (Koutroulis and Blaabjerg, 2012), (Kerekes et al., 2013). The decrease in the \( LCOE \) can be achieved by minimizing the life cycle cost \( (C_{LC}) \) and maximizing the useful lifetime. In this respect, the reliability improvement is of primary importance because it allows reducing \( C_{LC} \) by decreasing the maintenance and repair costs and increasing the \( E_{tot} \) by increasing the useful lifetime.

The \( LCOE \) and \( EPB \) are effective methods to compare different converter solutions for a specific application (Popović-Gerber et al., 2011), (Jeng-Yue et al., 2010), which allow quantifying the improvements in the converter performance.

1.3 **Objective of the work**

The growing requirements for the power converters will be met by continuous progress in their design. Recently, the power converter is considered as the equipment between the electric power source and the load used for the conversion and the control of electromagnetic energy flow, not restricted to the concept in terms of electrical circuit diagram (abstract circuit topology). In association with the increasing requirements for the power converter, the design is becoming more complicated and a multidisciplinary approach to the modelling of the converter is required. As discussed in (van Wyk and Lee, 2013), the opportunities for the converter development tend to come from external technologies rather than internal (Figure 1.11). The internal technologies of semiconductors and converter circuits are approaching maturity (except for the wide band
gap devices), which is demonstrated by the fact that despite the variety of topologies emerged recently, the topologies being introduced decades ago are applied in industry. Meanwhile, the external technologies of packaging, cooling, manufacturing, and electromagnetic impact present remarkable opportunities for the development and contribute to the complex nature of the design and building of the power converter.

<table>
<thead>
<tr>
<th>Power electronics technology</th>
</tr>
</thead>
<tbody>
<tr>
<td>Internal technologies</td>
</tr>
<tr>
<td>Power switch technology</td>
</tr>
<tr>
<td>- Device technology</td>
</tr>
<tr>
<td>- Driving technology</td>
</tr>
<tr>
<td>- Snubbing technology</td>
</tr>
<tr>
<td>- Protection technology</td>
</tr>
<tr>
<td>Network technology</td>
</tr>
<tr>
<td>- Switching technology</td>
</tr>
<tr>
<td>- Topological arrangement</td>
</tr>
<tr>
<td>Passive component technology</td>
</tr>
<tr>
<td>- Magnetic components</td>
</tr>
<tr>
<td>- Capacitive components</td>
</tr>
<tr>
<td>- Conductive components</td>
</tr>
<tr>
<td>External technologies</td>
</tr>
<tr>
<td>Packaging technology</td>
</tr>
<tr>
<td>- Materials technology</td>
</tr>
<tr>
<td>- Interconnection technology</td>
</tr>
<tr>
<td>- Layout technology</td>
</tr>
<tr>
<td>- Mechanical construction technology</td>
</tr>
<tr>
<td>Electromagnetic environmental impact technology</td>
</tr>
<tr>
<td>- Harmonics</td>
</tr>
<tr>
<td>- Network distortion</td>
</tr>
<tr>
<td>- EMI</td>
</tr>
<tr>
<td>- EMC</td>
</tr>
<tr>
<td>Cooling technology</td>
</tr>
<tr>
<td>- Cooling fluids</td>
</tr>
<tr>
<td>- Circulation</td>
</tr>
<tr>
<td>- Heat extraction and conduction</td>
</tr>
<tr>
<td>- Heat exchanger construction</td>
</tr>
<tr>
<td>Physical environmental impact technology</td>
</tr>
<tr>
<td>- Acoustic interaction</td>
</tr>
<tr>
<td>- Physical material interaction</td>
</tr>
<tr>
<td>- Recycling</td>
</tr>
<tr>
<td>- Pollution</td>
</tr>
<tr>
<td>Manufacturing technology</td>
</tr>
<tr>
<td>Converter sensing and control technology</td>
</tr>
</tbody>
</table>

![Figure 1.11. Power electronics constituent technologies adapted from (van Wyk and Lee, 2013).](image)

The power converter design is a challenging task requiring analysis of different aspects and understanding the relations between them. Different modelling tools are employed such as 2D/3D numerical modelling tools (electromagnetic, thermal, and mechanical), Computer-aided Design (CAD) based programs, and circuit simulators to model the external technologies and analyse their influence on the converter performance. With the
modern availability of computational resources, modelling and visualization tools, the power electronic converter can be built on the computer, its operation can be studied, the design can be optimized, and only after that, a prototype is built.

The objective of the doctoral dissertation is to develop the methods and models used for the design of the main circuit of the converter that allow analysing the converter performance at an early design stage. Comprehensive electromagnetic and electrothermal models are drawn up to design a converter with the required characteristics and investigate the opportunities for the improvement of reliability, power density, weight, and cost.

An electromagnetic model of the converter main circuit is developed in the study to evaluate the stray inductance of the converter commutation loops. The model is used to minimize the stray inductance that affects the converter power capability and reliability.

The electrothermal models of the converter including the IGBT modules and the laminated busbar system applied to connect the converter components are established to analyse the thermal aspects of the converter design that have a strong influence on the power density, reliability, weight, and cost. The reliability analysis is provided to estimate the lifetime of the converter under study and define the necessity for improvements in the main circuit design.

1.4 Outline of the work

The doctoral dissertation comprises six chapters, which are organized as follows:

Chapter 1 presents the role of power electronics in energy saving and development of renewable energy sources. Operating principles, control, and modulation of a three-level NPC/ANPC converter are reviewed. A literature survey on the semiconductor devices used in power converters is provided. Motivation for the work is given, and challenges associated with the design of modern multilevel converters and opportunities for development are discussed.

Chapter 2 is dedicated to the low-inductive design of an ANPC converter main circuit. A low-inductive layout of the laminated busbar system for an ANPC converter is proposed. The design and modelling aspects are discussed in detail. The electromagnetic model of the laminated busbar system and the IGBT modules is presented.

Chapter 3 presents the design of the busbar system for a non-destructive test (NDT) set-up used to perform the short-circuit tests of IGBT modules. The circular symmetry is propped to obtain equal current sharing among the parallel components of the set-up.

In Chapter 4, a tool for the thermal analysis of the converter busbar system is developed. The thermal analysis of the designed busbar system is provided based on a lumped parameter thermal model (LPTM).
Chapter 5 presents the reliability analysis of the IGBT modules. The method to generate the accurate mission profile of a wind turbine converter is developed. The thermal model of the IGBT module and the cooling system is developed for the lifetime estimation. The influence of the cooling solution on the lifetime of the IGBT is investigated in this chapter. The method, based on the usage of the gradient heat flux sensor (GHFS) to control the thermal cycles of the IGBTs is introduced.

1.5 Scientific contributions and publications

The main scientific contributions of this work are:

- development of a 3D model of the converter main circuit with a detailed model of the IGBT module and an analysis of the influence of the mutual inductance between the IGBT modules and the IGBT module and the busbars on the inductance of the converter commutation loops with a numerical tool;
- investigation of an option to use the circular layout of the laminated busbar system to achieve equal current sharing among the parallel-connected components;
- development of an algorithm for the temperature and temperature-dependent loss estimation of the laminated busbar system;
- development of the 3D lumped parameter thermal model of the laminated busbar system;
- implementation of the IGBT lifetime estimation algorithm;
- investigation of the influence of the thermal inertia of the cooling solution on the lifetime of the IGBT modules;
- analysis of the opportunities of using a gradient heat flux sensor (GHFS) in the thermal control of an IGBT.

The results related to the topic of this doctoral dissertation have been presented in the following publications:


\(^2\) The maiden name of the author of this doctoral dissertation


Other papers published during the writing of this doctoral dissertation are listed below:


kA/1.1 kV non-destructive testing equipment." In *IECON 2014 - 40th Annual Conference of the IEEE Industrial Electronics Society*, Dallas, TX, USA.


Low-inductive design of the converter

With the emergence of fast semiconductor switches and rise in the current and voltage levels of the converters, the stray inductances are causing severe problems. According to Faraday’s induction law, a voltage spike $U_{\text{spike}}$ occurs during the switch-off of the current $i$ as a result of the presence of the stray inductance $L_{\text{stray}}$

$$U_{\text{spike}} = L_{\text{stray}} \frac{di}{dt}.$$ \hspace{1cm} (2.1)

The stray inductance decreases the converter power capability by limiting the switching current or switching speed of the semiconductor switches. A lower switching speed increases switching losses but keeps the voltage spikes at an acceptable level. In order to achieve the required current level, either the speed of the switch should be decreased (e.g. by increasing the turn-off gate resistor) or a switch with a higher rated voltage should be used. In both cases, the semiconductor losses increase and the efficiency of the converter decreases. In fact, nowadays the packaging stray inductance is already limiting the further speed improvement of semiconductor switches (van Wyk and Lee, 2013). New packaging of the switches themselves and new converter main circuit layouts are required in order to make new high-speed devices feasible. Clamping of voltage spikes by using snubber circuits is not desirable because it adds complexity, losses, and cost, and sacrifices the converter reliability and efficiency.

Another adverse effect of the stray inductance is the increase in the voltage rate of rise at the recovery of the freewheeling diode. This $\frac{du}{dt}$ increases the peak reverse recovery power of the diode and may stress the insulation of the motor connected to the converter. Furthermore, the inductive loops in conjunction with parasitic capacitive elements can cause oscillation and will act as EMI generators in normal and abnormal (short-circuit) operation conditions (Schnur et al., 1998), (Wu et al., 2015).

Considering the stated arguments, efforts are put to minimize the stray inductance of the converter commutation loops at an early design stage. Because of a larger number of devices participating in the commutation loops of a multilevel converter in comparison with a two-level counterpart, its stray inductance is generally higher and more difficult to minimize. In this chapter, the design of the three-level ANPC converter main circuit is described. The relevant aspects such as the arrangement of the circuit components and the design of the low-inductive laminated busbar system to connect the components are discussed.

2.1 Stray inductance of the converter commutation loops

Each component of the converter has a parasitic inductance that contributes to the inductance of the commutation loops described in Section 1.1.1 and shown in Figure 2.1.
The parameters of the ANPC converter and its components are listed in Table A.1 and Table A.2 of Appendix A. Thus, the commutation loop A contains two IGBT modules and eight capacitors, and the commutation loop B contains four IGBT modules and eight capacitors. In order to reduce the inductance of the commutation loops, components with low stray inductances have to be selected. Among the manufacturers, the trend is toward minimization of the stray inductances of the capacitors and switching components (Frisch and Ernö, 2010), (Li et al., 2011). However, to benefit from the low inductance of the switches and capacitors, a low inductance of the connectors is also required as a high number of them are needed to connect the converter components. This is the reason for using the laminated busbar system, which allows achieving a low inductance when properly designed (Caponet et al., 2002).

Figure 2.1. One phase leg of a three-level ANPC converter, commutation loops A and B. P – positive busbar, N – negative busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link, A3 – additional busbar of the lower phase arm, and A4 – additional busbar of the lower part of the DC link.

2.2 Theory of partial and loop inductances

The concepts of loop and partial inductances are reviewed in this section to facilitate further discussion about the inductance of the converter commutation loops. A comprehensive description of the theory of partial inductance is presented in (Ruehli, 1972). The difference between the concepts of loop and partial inductances is explained in detail in (Paul, 2010).
Inductance of a single-turn current loop presented in its general form is defined as a ratio of the total magnetic flux $\Phi$ penetrating the surface enclosed by the current loop and the current $I$ that produce it as

$$L = \frac{\Phi}{I}.$$  

(2.2)

This inductance is known as a self-inductance of the loop and is referred to as loop inductance in this doctoral dissertation. The complete current loop should be defined in order to obtain the total magnetic flux with a surface integral as

$$\Phi = \int_S B \cdot dS ,$$  

(2.3)

where $B$ is the magnetic flux density and $S$ is the surface enclosed by the current loop.

The magnetic flux can be obtained alternatively by using Stoke’s theorem and the magnetic vector potential $A$, which is defined by

$$B = \nabla \times A$$  

(2.4)

as

$$\Phi = \int_S B \cdot dS = \int_S (\nabla \times A) \cdot dS = \oint_l A \cdot dl ,$$  

(2.5)

where $l$ is the contour that encloses the surface $S$.

Thus, an alternative method of calculating the loop inductance is

$$L = \frac{\oint_l A \cdot dl}{I} .$$  

(2.6)

Further, Equation 2.6 can be decomposed into the line integral along the $n$ segments of the current loop
where the contour \( l \) is divided into \( n \) segments and \( A_i \) is the total magnetic vector potential along the segment \( l_i \) caused by the current of this segment and the other segments of this contour or of some other contours.

The inductance contribution of each segment to the loop inductance is called the net inductance of that segment and defined as

\[
L_i = \frac{\int_{l_i} A_i \cdot dl}{I}, \quad (2.8)
\]

and the total loop inductance is the sum of the net inductances of the loop segments

\[
L = \sum_{i=1}^{n} L_i. \quad (2.9)
\]

The net inductance of the segment can be written as a sum of partial inductances

\[
L_i = \sum_{j=1}^{n} L_{pij}. \quad (2.10)
\]

The partial inductance \( L_{pij} \) is defined as a ratio of the magnetic flux penetrating the rectangular surface between the segment \( i \) and infinity to the current \( I_j \) that produces that flux and is presented as

\[
L_{pij} = \frac{\int_{l_i} A_{ij} \cdot dl}{I_j}, \quad (2.11)
\]

where \( A_{ij} \) is the magnetic vector potential caused by the current of segment \( j \). When \( i = j \), the \( L_{pij} \) is called the self-partial inductance for the segment \( i \), because the magnetic flux
penetrating the surface between the segment $i$ and infinity is produced by the current of that segment (Figure 2.2 (a)). When $i \neq j$, $L_{pi}$ is called the mutual-partial inductance between the segment $i$ and $j$ and is also denoted as $M_{pij}$; it can be positive or negative depending on the relative orientation of the current in the segment $i$ to the magnetic vector potential $A_{ij}$ (Figure 2.2 (b)). In practice, if the currents in the segments $i$ and $j$ flow in the same direction, the mutual-partial inductance between these segments is positive, and if the currents flow in the opposite direction, the mutual-partial inductance is negative. The negative mutual-partial inductance effectively decreases the net inductance of the segment (Equation 2.10), and consequently, the total loop inductance (Equation 2.9). The value of the mutual-partial inductance between the loop segments depends on the distance between these segments as shown by substituting the explicit equation for $A_{ij}$ in Equation (2.11) (Paul, 2010)

$$M_{pij} = \frac{\mu_0}{4\pi} \int_{l_i} \int_{l_j} \frac{1}{b_{ij}} dl_i \cdot dl_j,$$

(2.12)

where $b_{ij}$ is the distance between the differential segment $dl_i$ along the contour $l_i$ and $dl_j$ along the contour $l_j$ (Figure 2.2 (b)).

The equation to calculate the mutual-partial inductance between two parallel long rectangular busbars having equal dimensions and shown in Figure 2.3 is given in (Holloway et al., 2013) as

$$M_p = \frac{\mu_0}{2\pi} l \left( \ln \frac{2l}{b} - 1 \right),$$

(2.13)

Figure 2.2. Physical meaning of (a) self-partial inductance and (b) mutual-partial inductance.
where the length of the busbars \( l \) is much larger than their thickness \( d \) and the distance between them \( b \). Thus, the mutual-partial inductance between the segments is higher when the distance between them is smaller.

![Two parallel rectangular busbars](image)

**Figure 2.3.** Two parallel rectangular busbars. \( l \) – the length of the busbar, \( w \) – the width of the busbar, \( d \) – the thickness of the busbar, and \( b \) – the distance between the busbars.

The equivalent lumped circuit of the current loop can be constructed using partial inductances by dot convention and solved with a circuit simulator. For example, in Figure 2.4 (b), the lumped circuit of the current loop shown in Figure 2.4 (a) is developed. The current loop is divided into four segments, and each segment is characterized by partial inductances.

![Rectangular loop](image)

**Figure 2.4.** (a) Rectangular loop. The loop current \( I \) produces a magnetic flux density \( B \) inside the loop. (b) Rectangular loop in terms of partial inductances for the conductor length \( l \). Similar observation is needed for all lengths \( l \). \( L_p \) is a self-partial inductance and \( M_p \) is a mutual-partial inductance.
The voltage drop across any segment can be determined in terms of the net inductance of that segment as

\[ U_i = L_i \frac{dI}{dt} \quad (2.14) \]

In this doctoral dissertation, the concept of partial inductance is applied when dealing with the stray inductance of the converter commutation loops because of issues related to the use of the concept of loop inductance. The first issue of using the loop inductance is that in order to calculate the inductance of the current loop, the magnetic flux that passes through the enclosed surface must be computed. Therefore, the complete current loop should be identified. For intentional inductors, such as solenoids or toroids, the complete current loop is obvious and defined by a designer of a device. Therefore, these devices can be characterized by lumped-circuit elements calculated by Equation (2.2) and (2.3). However, for unintentional inductors such as the inductance of the converter commutation loops, the definition of the complete current loop is not so straightforward because the internal geometry of some components (IGBT modules, capacitors) is not known. However, the components of the commutation loops with a known geometry such as the laminated busbar structure can be characterized by the partial inductances.

The second issue is that the inductance of the current loop cannot be placed in any unique position in that loop. For example, there is no unique position for the calculated inductance of commutation loop A in the circuit presented in Figure 2.1, and therefore, the effect of this inductance cannot be modelled with a circuit simulator. However, the lumped-circuit model of the commutation loop can be developed using partial inductances and solved in any circuit simulator.

### 2.3 Design of the laminated busbar system

The laminated busbar system consists of several conducting busbars with thin insulation layers between them as presented in Figure 2.5. An advantage from a thermal point of view is the larger contact area with the ambient of that system, which allows better heat dissipation.

![General 3D view of the busbar system.](image)
In this section, the design of the laminated busbar system is considered, and a layout for the ANPC converter is proposed. The parameters of the ANPC converter and its components are listed in Table A.1 and Table A.2 of Appendix A. The selection of materials, laminated structure, and the component arrangement are considered for achieving a low stray inductance.

2.3.1 Selection of materials

The selection of the materials for the conducting busbars is considered taking into account the electrical, thermal, mechanical, and cost requirements. Table 2.1 shows the properties of some commonly used materials for the busbars with non-magnetic behaviour.

<table>
<thead>
<tr>
<th>Material</th>
<th>Electric conductivity, S/m</th>
<th>Density, kg/m³</th>
<th>Thermal conductivity, W/(m·K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gold</td>
<td>4.1·10⁷</td>
<td>19300</td>
<td>317</td>
</tr>
<tr>
<td>Aluminium</td>
<td>3.5·10⁷</td>
<td>2702</td>
<td>237</td>
</tr>
<tr>
<td>Silver</td>
<td>6.2·10⁷</td>
<td>10500</td>
<td>429</td>
</tr>
<tr>
<td>Copper</td>
<td>5.8·10⁷</td>
<td>8933</td>
<td>401</td>
</tr>
</tbody>
</table>

Despite the good properties of gold and silver, these materials are not widely used for busbars because of their high prices. Gold is, however, often used as thin coating to prevent surface oxidation. Aluminium and copper are two suitable options with a reasonable cost. Copper has better electrical and thermal conductivities, mechanical properties, and corrosion resistance than aluminium. However, aluminium is cheaper and significantly lighter than copper. The specific conductivity, that is, the conductivity divided by density, is 12960 Sm²/kg for aluminium and 6621 Sm²/kg for copper, making aluminium very attractive when lightness is desired.

However, for the busbars of the multilevel converter with a high number of layers, copper is chosen because of its superior thermal and electrical conductivities that allow obtaining a more compact solution with better heat transfer.

In order to achieve better resistance to fatigue, creep, and wear, copper-based alloys are widely used. Copper that contains less than 1% impurities is used for electrical applications. Figure 2.6 shows the effect of adding materials such as tin, silver, zinc, iron, or phosphorus on the electrical conductivity of copper (Askeland and Haddleton, 1996).
The properties of insulation materials that can be used to isolate the conducting plates are presented in Table 2.2 (Du Pont Teijin Films, 2014), (Du Pont Films, 2014a), (Du Pont Films, 2014b). The thickness of the inner insulation layer in the busbar is calculated by

\[
d \geq \frac{U_{\text{max}}}{E_{\text{max}}},
\]

(2.15)

where \(U_{\text{max}}\) is the possible maximum voltage between the conducting busbars and \(E_{\text{max}}\) is the dielectric strength of the insulation material. A material with a high dielectric strength is preferred to obtain thin insulation that leads to a lower stray inductance of the busbar system.

However, in practice, \(d\) has to be larger than the limiting value that Equation (2.15) gives as local stresses occur, especially, in points of conductor discontinuity. It is also clear that from a mechanical point of view it is recommendable to select slightly thicker materials than Equation 2.15 gives for the minimum thickness.

As an example we can calculate the minimum thickness for Mylar in the case of 4.5 kV DC voltage. We get \(d > 4.5 \text{ kV/} 315 \text{ kV/mm} = 0.014 \text{ mm}\). Such a thin foil can easily
suffer from mechanical defects, and it is advisable to select a thicker material, for instance 0.0762 mm, in practice.

Table 2.2. Properties of insulating materials (Du Pont Teijin Films, 2014), (Du Pont Films, 2014a), (Du Pont Films, 2014b), (Pyrhönen et al., 2008).

<table>
<thead>
<tr>
<th>Material</th>
<th>Relative dielectric constant $\varepsilon$</th>
<th>Thickness, mm</th>
<th>Dielectric Strength $E_{\text{max}}$, kV/mm</th>
<th>Maximum Continuous Use temperature, °C</th>
<th>Thermal conductivity, W/(m·K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mylar$^{(1)}$, Polyester film</td>
<td>60 Hz</td>
<td>3.3</td>
<td>0.0254</td>
<td>315</td>
<td>150</td>
</tr>
<tr>
<td></td>
<td>1 GHz</td>
<td>2.8</td>
<td>0.0508</td>
<td>217</td>
<td>0.16</td>
</tr>
<tr>
<td>Kapton$^{(1)}$, Type HN and VN (Polyimide film)</td>
<td>1 kHz</td>
<td>3.5</td>
<td>0.0254</td>
<td>303</td>
<td>400</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>0.0508</td>
<td>240</td>
<td>0.12</td>
</tr>
<tr>
<td>Teflon$^{(1)}$, PEF (Fluorocarbon film)</td>
<td>100 Hz – 1 MHz</td>
<td>2.0</td>
<td>0.0254</td>
<td>260</td>
<td>205</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>0.508</td>
<td>70</td>
<td>0.195</td>
</tr>
<tr>
<td>Nomex$^{(1)}$, type 411 paper</td>
<td>60 Hz</td>
<td>1.2</td>
<td>0.13</td>
<td>18</td>
<td>220</td>
</tr>
<tr>
<td></td>
<td>1 GHz</td>
<td>1.3</td>
<td>0.18</td>
<td>18</td>
<td>0.11</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>0.38</td>
<td>16</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>0.58</td>
<td>16</td>
<td></td>
</tr>
</tbody>
</table>

1 Mylar®, Kapton®, and Teflon® are registered trademarks of Dupont

2.3.2 Laminated structure

Figure 2.7 presents a cross-sectional view of a laminated busbar system used to connect the components in the ANPC converter. An exploded view of the busbar system is shown in Figure 2.8. The busbar system contains seven layers. The conducting layers are made of copper (thickness is 2 mm), and the insulating layers are made of Teflon (thickness is 1 mm). The order of the busbars is chosen such that the inductances of the commutation loops are minimized.

When designing the laminated busbar system for the converter, the main objective is to minimize the net inductance of each busbar, which leads to the low inductance of the commutation loops. As the net inductance of the busbar consists of the self-partial inductance and mutual-partial inductances between this busbar and other busbars of the commutation loop (Equation 2.10), it can be minimized by minimizing the self-partial inductance and positive mutual-partial inductances while maximizing the negative mutual-partial inductances. In order to minimize the self-partial inductance of the busbar, its width should be increased but the length decreased (Skibinski and Divan, 1993). However, there are other factors constraining the selection of the busbar dimensions such as the physical layout, connection problems, mechanical rigidity, and busbar temperature rise.
Figure 2.7. Cross-sectional view of the laminated busbar system for the ANPC converter with commutation loop A. P – positive busbar, N – negative busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link, A3 – additional busbar of the lower phase arm, A4 – additional busbar of the lower part of the DC link, C1 – C16 – the DC link capacitors, T\textsubscript{x1}, T\textsubscript{x2}, T\textsubscript{x5} – the IGBT modules.

Figure 2.8. Exploded view of the laminated busbar system for the ANPC converter. P – positive busbar, N – negative busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link, A3 – additional busbar of the lower phase arm, A4 – additional busbar of the lower part of the DC link.
The mutual-partial inductance between two busbars can be changed by adjusting the distance between these busbars (Equation 2.13) and thereby the position of the busbars in the laminated system. Thus, busbar P is placed above busbars NT and A2 because the current in commutation loops A and B flows in busbar P in one direction and in busbars A2 and NT in the opposite direction; consequently, the mutual-partial inductances between busbars P and A2 and between busbars P and NT are negative. This placement minimizes the distance between these busbars to the thickness of the insulation and maximizes the negative mutual-partial inductance between busbars P and A2 and between busbars P and NT. For the same reason, busbar N is located above busbars A4 and NT. There is no need to consider the mutual-partial inductance between the busbars that are not included in the same commutation loops. Consequently, busbars P and N are located in the same plane similarly as the additional busbars A2 and A4.

In the multilayer busbar system of the multilevel converter, the distance between the busbars in which the current flows in the opposite direction is not only the thin insulation layer but also the conducting plate. To maximize the negative mutual-partial inductance between such layers, the thickness of the conducting plates should be decreased, which leads to economic benefits and saving in material. However, it is known that the temperature rise of the busbars limits the thickness decrease; therefore, an accurate thermal model of the busbars will allow estimating the temperature rise of the busbars in the design and then determining the allowed minimum thickness of the conducting plates. The thermal analysis of the busbars is presented in Chapter 4.

2.3.3 Location of the components

The proposed placement of the converter main circuit components has been presented in (Popova et al., 2014) and chosen to minimize the length of the busbars and to ensure equal inductances of the commutation loops in the upper and lower phase arms and between the phase legs. The influence of the component placement on the stray inductance of the commutation loops is studied by the author of this doctoral dissertation, and the results are presented in (Popova et al., 2012).

As shown in Figure 2.9, the capacitors of the upper and lower parts of the DC link are placed symmetrically about the switching components located in the centre. This configuration allows obtaining equal lengths of the commutation loops of the upper and lower arms of the converter.
The arrangement of the IGBT modules is selected to achieve equal lengths of the commutation loops of three phase legs and to decrease the lengths of the busbars included in the commutation loop B that has a higher inductance than commutation loop A because of the higher number of the IGBT modules. For this purpose, IGBT modules $T_{x1}$ and $T_{x2}$ are placed close to the upper part of the DC link (C1–C8) and $T_{x3}$ and $T_{x4}$ are located close to the lower part of the DC link (C9–C16). This arrangement allows decreasing the length of commutation loop B by increasing the distance between IGBT module $T_{x5}$ and the upper part of the DC link and, consequently, by increasing the length of commutation loop A. The distance between IGBT module $T_{x5}$ and the lower part of the DC link is also increased. The 3D view of the converter main circuit with the designed busbar system is shown in Figure 2.10.
2.4 **Inductance estimation of the laminated busbar system**

The stray inductance of the commutation loops has to be estimated at an early design stage to verify that the selected switching components are not subjected to a severe overvoltage during the converter operation and to take corrective actions if problems are detected. The components of the commutation loops can be modelled using partial inductances that allow constructing a lumped-circuit model of the commutation loops. Thus, the behaviour of the converter can be analysed in a circuit simulator taking stray inductances into account.

2.4.1 **Partial inductance estimation**

Accurate estimation of the partial inductances of the current loop components is critical for correct calculation of the loop inductance and constructing the equivalent lumped-circuit model. Selection of the appropriate method for partial inductance calculation of the laminated busbar system is discussed further. A review of existing analytical and numerical methods is given.

The primary assumption made for the analytical inductance estimation is that the current is uniformly distributed over the conductor cross section. There are several analytical methods for the partial inductance estimation of a loop segment. The methods can be divided into direct and indirect ones. Direct methods require computation of the magnetic flux penetrating the surface between the segment and infinity. Magnetic flux is calculated using the magnetic flux density (Equation 2.3), which can be obtained by applying the
Biot-Savart law, Ampère’s law for problems with symmetry, or using magnetic vector potential.

An indirect method of inductance estimation is to compute the energy $W_m$ stored in the magnetic field

$$L = \frac{2W_m}{I^2}. \quad (2.16)$$

The partial inductance of a wire having a circular cross section is fairly simple to compute by direct methods. The reason is that for the computation of the magnetic field produced by the current of the wire, the wire can be replaced by a filament containing the total current. However, for a conductor with a rectangular cross section, this assumption is not valid. The partial inductances of rectangular conductors are computed by the indirect method in terms of stored magnetic energy. Analytical formulas for the calculation of the partial inductances of rectangular conductors are presented for instance in (Grover, 1946), (Hoer and Love, 1965), and derivations are given in (Paul, 2010).

Even though the laminated busbar system considered in this dissertation consists of busbars having a rectangular cross section, the application of the analytical methods is hindered because of the necessity to estimate the inductance at the highest critical frequency $f_{cr}$ associated with the IGBT fall time when the voltage spike occurs (Skibinski and Divan, 1993). At this frequency, the skin effect may be significant. The proximity effect is also pronounced for the busbars located very close to each other. These two effects influence the current distribution in the busbars, and thus, a primary assumption of uniform current distribution is not valid and analytical methods are not applicable. With the enormous computational capabilities of modern computers, numerical modelling tools are successfully used to estimate the partial inductances considering the skin and proximity effects.

All the numerical tools for electromagnetic analysis are fundamentally based on solving Maxwell’s equations either in a differential or integral form.

The most popular numerical technique for solving differential equations is called the Finite Element Method (FEM), which is based on direct discretization of Maxwell’s equations. FEM-based numerical tools are widely applied for inductance estimation as shown in (Zare and Ledwich, 2002), (Lai et al., 2006), (Popova et al., 2012).

Numerical tools using integral equation methods have shown good performance in calculating the inductance of complex geometries (Clavel et al., 2009), (Ardon et al., 2010), (Schanan et al., 1994), (Tran et al., 2010), (Popova et al., 2013). The Method of Moments (MoM) and the Partial Element Equivalent Circuit (PEEC) method are two commonly used numerical techniques for solving the integral equations. A FEM-PEEC-
coupled method is also presented in the literature (Tran et al., 2010). All these methods allow modelling of the skin and proximity effects.

In this work, the ANSYS Q3D software is used for the partial inductance estimation of the busbars. This 3D numerical modelling tool is used for extraction of the resistance, partial inductance, capacitance, and conductance. The Q3D extractor performs the electromagnetic field simulation by using a combination of the FEM and MoM methods. The results provided by this tool take proximity and skin effects into account.

The geometry of the busbar system has been drawn in a CAD program and then exported to Q3D, where the material properties and boundary conditions are provided. In order to describe the procedure of the inductance estimation, a simple busbar system shown in Figure 2.11 and consisting of three copper busbars B1, B2, and B3 connecting IGBT modules $T_1$ and $T_2$ in series is considered. The measuring points, that is, the source and sink are assigned in Q3D as shown in Figure 2.11 (c), and thus, the current path is from point S to point F. The results obtained at 10 kHz frequency are presented in Table 2.3, where the diagonal elements are partial self-inductances, and the non-diagonal elements are mutual-partial inductances.
Table 2.3. Self-partial and mutual-partial inductances of busbars B1, B2, and B3. The diagonal elements ($L_{pi,i}$) are the self-partial inductances of the components, and the off-diagonal elements ($M_{pi,j}$) are the mutual-partial inductances.

<table>
<thead>
<tr>
<th>Inductance, nH</th>
<th>B1</th>
<th>B2</th>
<th>B3</th>
</tr>
</thead>
<tbody>
<tr>
<td>B1</td>
<td>3.6</td>
<td>0.1</td>
<td>-0.02</td>
</tr>
<tr>
<td>B2</td>
<td>0.1</td>
<td>14.1</td>
<td>0.1</td>
</tr>
<tr>
<td>B3</td>
<td>-0.02</td>
<td>0.1</td>
<td>3.6</td>
</tr>
</tbody>
</table>

2.4.2 Loop inductance estimation

The inductances of commutation loops A and B of a three-level ANPC converter are calculated as a sum of the self-partial inductances of the DC link capacitors ($L_{p,DC\_link}$) and the IGBT modules ($L_{p,IGBT}$) taken from the datasheets provided by manufacturers, and the equivalent loop inductance of the laminated busbar system ($L_{busbars}$) as follows:

\begin{align*}
L_{\text{loopA}} &= L_{p,DC\_link} + 2L_{p,IGBT} + L_{\text{busbars}}, \\
L_{\text{loopB}} &= L_{p,DC\_link} + 4L_{p,IGBT} + L_{\text{busbars}}.
\end{align*}

The equivalent loop inductance of the laminated busbar system is computed by Equation (2.9) using partial inductances estimated numerically in Q3D according to the procedure described in Section 2.4.1 and presented in Table 2.4. This approach assumes that the mutual-partial inductances between the components are small and can be neglected. The loop inductances computed by Equation (2.17) and (2.18) are also shown in Table 2.4.

Table 2.4. Estimated stray inductances of the commutation loops at 1.2 MHz.

<table>
<thead>
<tr>
<th>Notation</th>
<th>Estimated inductance, nH</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Phase A</strong></td>
<td></td>
</tr>
<tr>
<td>$L_{\text{busbarsA}}$</td>
<td>21</td>
</tr>
<tr>
<td>$L_{\text{loopA}}$</td>
<td>63</td>
</tr>
<tr>
<td>$L_{\text{busbarsB}}$</td>
<td>29</td>
</tr>
<tr>
<td>$L_{\text{loopB}}$</td>
<td>103</td>
</tr>
<tr>
<td><strong>Phase B</strong></td>
<td></td>
</tr>
<tr>
<td>$L_{\text{busbarsA}}$</td>
<td>18</td>
</tr>
<tr>
<td>$L_{\text{loopA}}$</td>
<td>60</td>
</tr>
<tr>
<td>$L_{\text{busbarsB}}$</td>
<td>26</td>
</tr>
<tr>
<td>$L_{\text{loopB}}$</td>
<td>100</td>
</tr>
<tr>
<td><strong>Phase C</strong></td>
<td></td>
</tr>
<tr>
<td>$L_{\text{busbarsA}}$</td>
<td>20</td>
</tr>
<tr>
<td>$L_{\text{loopA}}$</td>
<td>62</td>
</tr>
<tr>
<td>$L_{\text{busbarsB}}$</td>
<td>28</td>
</tr>
<tr>
<td>$L_{\text{loopB}}$</td>
<td>102</td>
</tr>
</tbody>
</table>
As it is shown in Table 2.4, the proposed laminated busbar system and the layout of the main circuit allow obtaining a low stray inductance and a small variation in the stray inductance values between the phases. 100 nH allows 400A/100ns $d_i/dt$ values if 400 V peaks are allowed during switch-off. This seems reasonable in the case of a 690 V converter with a 1000 V DC link voltage and 1700 V IGBTs.

2.4.3 Experimental verification

The results of the stray inductance estimation shown in Table 2.4 were verified by measuring the voltage spike $U_{\text{spike}}$ and the current slope $d_i/dt$ during the turn-on and turn-off of the IGBT modules. The current has been measured by a Rogowski coil. In Figure 2.12 and Figure 2.13, the measured transient waveforms of IGBT modules $T_{x1}$ and $T_{x2}$ are smoothed by using a 10-point moving average.

![Figure 2.12](image1.png)

Figure 2.12. Measured switching voltage and current waveforms of IGBT module $T_{x1}$ during (a) turn-off ($d_i/dt$ is 715-10$^6$ A/s, $U_{\text{spike}}$ is 46 V) and (b) turn-on ($d_i/dt$ is 689-10$^6$ A/s, $U_{\text{spike}}$ is 43 V).

![Figure 2.13](image2.png)

Figure 2.13. Measured switching voltage and current waveforms of IGBT module $T_{x2}$ during (a) turn-off ($d_i/dt$ is 1490-10$^6$ A/s, $U_{\text{spike}}$ is 164 V) and (b) turn-on ($d_i/dt$ is 764-10$^6$ A/s, $U_{\text{spike}}$ is 81 V).
Based on the retrieved waveforms, the stray inductances of commutation loops A and B are calculated as 64 nH and 110 nH by

\[ L_{\text{loop}} = \frac{U_{\text{spike}}}{di/dt}, \]  

(2.19)

which are in good agreement with the estimated results.

As presented in (Popova et al., 2014), the equivalent inductance of the busbar system was also measured before the inverter was completely assembled. Table 2.5 presents a comparison between the measured and estimated inductance values of the laminated busbar system of loops A and B for each phase, where a good correlation between the estimated and measured results is observed.

Table 2.5. Comparison of the estimated and measured equivalent loop inductance of the laminated busbar system at 10 kHz.

<table>
<thead>
<tr>
<th></th>
<th>Estimated</th>
<th>Measured</th>
</tr>
</thead>
<tbody>
<tr>
<td>( L_{\text{busbarsA}} )</td>
<td>26</td>
<td>23</td>
</tr>
<tr>
<td>( L_{\text{busbarsB}} )</td>
<td>35</td>
<td>35</td>
</tr>
<tr>
<td>( L_{\text{busbarsA}} )</td>
<td>21</td>
<td>18</td>
</tr>
<tr>
<td>( L_{\text{busbarsB}} )</td>
<td>31</td>
<td>32</td>
</tr>
<tr>
<td>( L_{\text{busbarsA}} )</td>
<td>24</td>
<td>20</td>
</tr>
<tr>
<td>( L_{\text{busbarsB}} )</td>
<td>34</td>
<td>34</td>
</tr>
</tbody>
</table>

2.4.4 Detailed model of the IGBT module

The analysis of the stray inductance of the commutation loops showed that the contribution of the IGBT modules to the total loop inductances is the most significant, particularly for commutation loop B, which contains four IGBT modules (Figure 2.14).

The net inductance of the IGBT module basically contains only its self-partial inductance, and the mutual-partial inductances between the module and the other components of the commutation loop are assumed negligible. However, as it was shown in (Musznicki et al., 2003), for example, the mutual-partial inductance between the IGBT module and the conducting busbar can have a considerable impact on the loop inductance.

In order to minimize the net inductance of the IGBT module, the mutual-partial inductances between the module and the other components should be analysed. In this case, modelling of the IGBT modules (which requires knowledge of the module’s internal
structure, which is not typically provided by the manufacturers), is needed to estimate the mutual-partial inductances.

Figure 2.14. Inductance of commutation loops A and B. $L_{p,\text{DC\_link}}$ is the self-partial inductance of the DC link capacitors, $L_{p,\text{IGBTs}}$ is the sum of the self-partial inductances of the IGBT modules included in the commutation loop, and $L_{\text{busbars}}$ is the equivalent loop inductance of the laminated busbar system.

Figure 2.15. Lumped-circuit model of commutation loop A of a three-level ANPC converter in terms of the partial inductances of the components. P – positive busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link.
In this work, the mutual-partial inductances between the IGBT modules (e.g., $M_{p,Tx1,Tx5}$ in Figure 2.15) and between the IGBT module and the busbars (e.g., $M_{p,Tx1,P}$ in Figure 2.15) are estimated to analyse the influence of these mutual-partial inductances on the total loop inductance. In Figure 2.15, all mutual-partial inductances are not shown for the clarity of the figure.

In the converter under study, a 1700 V, 400 A single switch IGBT module is used as a switching device (Figure 2.16). The module contains four IGBT chips and eight freewheeling diode chips. One module was opened and its geometry drawn in a CAD-based tool and then used for the extraction of the partial inductances.

The surface current distribution at 1.2 MHz (the frequency associated with the fall time of the device (Skibinski and Divan, 1993)) is shown in Figure 2.17. A pronounced proximity effect, which forces the current to flow at the adjustment edges of the emitter and the collector terminals, is observed in Figure 2.17. The estimated self-partial inductance of the module at 1.2 MHz operating as an IGBT is 13.9 nH and 13.4 nH when operating as a diode. The DC inductances are 18.7 nH and 18 nH, respectively. The small difference between the inductance values of the module operating as an IGBT and as a diode is explained by the fact that the dominant stray inductance inside the module comes from the terminal leads (Xing et al., 1998). The manufacturer of the device provides a value of 16 nH; the small difference with the estimated value can be explained by the inaccuracy of the drawn geometry.
The estimated partial inductances of the components of commutation loop B are presented in Table 2.6. The diagonal elements of the inductance matrix are the self-partial inductances, and the off-diagonal elements are the mutual-partial inductances. The partial inductance matrix of the commutation loop A is presented in Table B.1 of Appendix B.

Table 2.6. Self-partial and mutual-partial inductances of the components constituting commutation loop B of phase a (at 1.2 MHz). The partial inductances of six busbars (P, A1, Ph, A3, NT, A2) and four IGBT modules (Tx1, Tx2, Tx3, and Tx6) are presented. The diagonal elements (L_{pi}) are the self-partial inductances of the components, and the off-diagonal elements (M_{pi}) are the mutual-partial inductances.

<table>
<thead>
<tr>
<th>Inductance, nH</th>
<th>P</th>
<th>A1</th>
<th>Ph</th>
<th>A3</th>
<th>NT</th>
<th>A2</th>
<th>Tx1</th>
<th>Tx2</th>
<th>Tx3</th>
<th>Tx6</th>
</tr>
</thead>
<tbody>
<tr>
<td>P</td>
<td>81.2</td>
<td>0.9</td>
<td>-10.3</td>
<td>-11.3</td>
<td>-39.0</td>
<td>-15.9</td>
<td>3.4</td>
<td>2.6</td>
<td>-3.1</td>
<td>4.4</td>
</tr>
<tr>
<td>A1</td>
<td>0.9</td>
<td>19.1</td>
<td>1.2</td>
<td>-6.4</td>
<td>-2.3</td>
<td>-1.7</td>
<td>-1.9</td>
<td>-1.0</td>
<td>0.0</td>
<td>-0.9</td>
</tr>
<tr>
<td>Ph</td>
<td>-10.3</td>
<td>1.2</td>
<td>21.5</td>
<td>-2.2</td>
<td>3.1</td>
<td>-0.1</td>
<td>-2.9</td>
<td>-3.7</td>
<td>2.7</td>
<td>-2.2</td>
</tr>
<tr>
<td>A3</td>
<td>-11.3</td>
<td>-6.4</td>
<td>-2.2</td>
<td>17.1</td>
<td>5.2</td>
<td>1.8</td>
<td>-0.5</td>
<td>0.1</td>
<td>-0.5</td>
<td>-0.8</td>
</tr>
<tr>
<td>NT</td>
<td>-39.0</td>
<td>-2.3</td>
<td>3.1</td>
<td>30.7</td>
<td>3.2</td>
<td>-1.1</td>
<td>-0.9</td>
<td>1.5</td>
<td>-2.4</td>
<td></td>
</tr>
<tr>
<td>A2</td>
<td>-15.9</td>
<td>-1.7</td>
<td>-0.1</td>
<td>1.8</td>
<td>3.2</td>
<td>7.2</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
</tr>
<tr>
<td>Tx1</td>
<td>3.4</td>
<td>-1.9</td>
<td>-2.9</td>
<td>-0.5</td>
<td>-1.1</td>
<td>0.0</td>
<td>13.9</td>
<td>1.2</td>
<td>-0.6</td>
<td>0.5</td>
</tr>
<tr>
<td>Tx2</td>
<td>2.6</td>
<td>-1.0</td>
<td>-3.7</td>
<td>0.1</td>
<td>-0.9</td>
<td>0.0</td>
<td>1.2</td>
<td>13.9</td>
<td>-0.5</td>
<td>0.6</td>
</tr>
<tr>
<td>Tx3</td>
<td>-3.1</td>
<td>0.0</td>
<td>2.7</td>
<td>-0.5</td>
<td>1.5</td>
<td>0.0</td>
<td>-0.6</td>
<td>-0.5</td>
<td>13.9</td>
<td>-1.1</td>
</tr>
<tr>
<td>Tx6</td>
<td>4.4</td>
<td>-0.9</td>
<td>-2.2</td>
<td>-0.8</td>
<td>-2.4</td>
<td>0.0</td>
<td>0.5</td>
<td>0.6</td>
<td>-1.1</td>
<td>13.4</td>
</tr>
</tbody>
</table>

As shown in Table 2.6, the mutual-partial inductance between IGBT modules Tx1 and Tx2 is positive because the currents flow in the same direction, and the mutual-partial inductance between Tx3 and Tx6 is negative because the currents flow in the opposite direction. The negative mutual-partial inductance helps to decrease the equivalent loop inductance. Although the self-partial inductance of the IGBT module cannot be changed, the sign and value of the mutual-partial inductances between the module and the other components of the loop can be altered. Thus, in the converter design by analysing the
mutual-partial inductances between the components, the location of the components can be changed to minimize the loop inductance.

However, as presented in Table 2.6, the mutual-partial inductances between the busbars are quite pronounced when they are placed close to each other and contribute significantly to the net inductances of the components. The mutual-partial inductances between the IGBT modules considered in this work are rather small and have a limited influence on the net inductances of the modules. The increase in the mutual-partial inductance between the IGBT modules requires a dense arrangement that is hard to obtain with such a module.

The increase in the mutual-partial inductance between the switching components is possible with a high level of integration when all the semiconductor switches of the converter are in a single case. That leads to considerable reduction in the commutation loop inductance and also in the converter size. However, this poses new challenges for the cooling system and may cause electromagnetic compatibility (EMC) problems.

2.5 Summary

This chapter dealt with the design of a low-inductive laminated busbar system and selection of the main circuit physical layout. The laminated busbar system for a three-level ANPC converter was proposed where the symmetrical placement of the main circuit components was adopted to achieve equal commutation loop inductances in three phases. A low inductance of the commutation loops was achieved thanks to the negative mutual-partial inductances between the loop components such as the IGBT modules and the busbars. The modelling of the busbar system was discussed in detail, and the model for the stray inductance estimation including the model of IGBT modules was developed.
3  Busbar system for the NDT set-up

The principles of the low-inductive design described in Chapter 2 are adopted for the design of the busbar system for a non-destructive test (NDT) set-up. An NDT set-up is used to analyse the behaviour of the semiconductor devices during a short-circuit. The presence of the protection circuit allows performing a lot of tests without destruction of the device (Busatto et al., 2009).

Several issues arise during the design of the busbar system for the NDT set-ups having high voltage and current ratings. First, the stray inductance of the circuit should be minimized because high current slopes occurring at the turn-on and turn-off of the short-circuit current during the test cause dangerous under- and overvoltages (Busatto et al., 2008), (Abbate et al., 2010). Second, an access for the infrared camera used to analyse the temperature distribution inside an open module is required. Third, an equal current sharing among the parallel-connected devices of the circuit should be ensured to avoid the damage of the device experiencing a higher current than the other devices during fast transients. As discussed in (Wu et al., 2015), the stray inductance of the circuit plays an important role in the divergent oscillations observed during the short-circuit test.

In the work, the busbar systems for two NDT set-ups are designed. The first NDT set-up is intended for a short-circuit current up to 6 kA (NDT set-up I), and the second set-up for the current up to 10 kA (NDT set-up II). The current-sharing issue is addressed in the design of NDT set-up II as a high number of parallel components is used to achieve the 10 kA current.

3.1  Description of the NDT set-up

The circuit diagram of the NDT with commutation loop 1 and loop 2 is shown in Figure 3.1. It includes the device under test (DUT), the series protection, the parallel protection, the load inductance \( L_{\text{load}} \), the DC link capacitance \( C_{\text{DC}} \), the high-voltage power supply \( U_{\text{DC}} \), Schottky diodes, and the negative-voltage capacitance \( C_{\text{NEG}} \) with the corresponding negative voltage supply \( U_{\text{NEG}} \). In order to achieve a high short-circuit current during the test, a lot of devices are connected in parallel as shown in Figure 3.1. The component specifications of NDT set-up I and II are listed in Table 3.1 and Table 3.2. The operation principles of the set-up are described in detail in (Smirnova et al., 2014), (Wu et al., 2014), (Wu et al., 2015). The set-ups are used for testing of a 1.7 kV, 1 kA IGBT module. The inductances of loop 1 and loop 2 are minimized in the design.
Figure 3.1. Electrical schematic of the NDT set-up. B1–B7 indicates the busbars.

**Table 3.1. Component specifications of NDT set-up I.**

<table>
<thead>
<tr>
<th>Component</th>
<th>Quantity</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC link capacitor $C_{\text{DC}}$ EPCOS B25620B1118K103</td>
<td>5</td>
<td>Capacitance, mF</td>
<td>1.1</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Rated voltage, kV</td>
<td>1.1</td>
</tr>
<tr>
<td>Negative-voltage capacitor $C_{\text{NEG}}$ EPCOS B25620B1118K103</td>
<td>3</td>
<td>Capacitance, mF</td>
<td>1.1</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Rated voltage, kV</td>
<td>1.1</td>
</tr>
<tr>
<td>Series protection switch Dynex DIM1500ESM33-TS000</td>
<td>2</td>
<td>Maximum collector-emitter voltage, kV</td>
<td>3.3</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Peak collector current, kA</td>
<td>3</td>
</tr>
<tr>
<td>Schottky diode STPS200170TV1</td>
<td>5</td>
<td>Repetitive peak reverse voltage, V</td>
<td>170</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Surge non-repetitive forward current, kA</td>
<td>1.2</td>
</tr>
<tr>
<td>Parallel protection switch Mitsubishi CM1200HC-66H</td>
<td>2</td>
<td>Maximum collector-emitter voltage, kV</td>
<td>3.3</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Peak collector current, kA</td>
<td>2.4</td>
</tr>
</tbody>
</table>

**Table 3.2. Component specifications of NDT set-up II.**

<table>
<thead>
<tr>
<th>Component</th>
<th>Quantity</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC link capacitor $C_{\text{DC}}$ Electronicon E50.S34-504NT0</td>
<td>10</td>
<td>Capacitance, mF</td>
<td>0.5</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Rated voltage, kV</td>
<td>2.4</td>
</tr>
<tr>
<td>Negative-voltage capacitor $C_{\text{NEG}}$ EPCOS B25620B1118K103</td>
<td>8</td>
<td>Capacitance, mF</td>
<td>1.1</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Rated voltage, kV</td>
<td>1.1</td>
</tr>
<tr>
<td>Series protection switch Dynex DIM1500ESM33-TS000</td>
<td>4</td>
<td>Maximum collector-emitter voltage, kV</td>
<td>3.3</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Peak collector current, kA</td>
<td>3</td>
</tr>
<tr>
<td>Schottky diode STPS200170TV1</td>
<td>10</td>
<td>Repetitive peak reverse voltage, V</td>
<td>170</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Surge non-repetitive forward current, kA</td>
<td>1.2</td>
</tr>
<tr>
<td>Parallel protection switch Mitsubishi CM1200HC-66H</td>
<td>4</td>
<td>Maximum collector-emitter voltage, kV</td>
<td>3.3</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Peak collector current, kA</td>
<td>2.4</td>
</tr>
</tbody>
</table>
3.2 Busbar system of NDT set-up I

The busbar system designed for NDT set-up I is shown in Figure 3.2. The busbars are made of electrical copper. The busbars in which the current flows in the opposite direction are placed close to each other to obtain a low inductance; the distance is only the thickness of the insulation, which is 0.2 mm (Mylar). As shown in Figure 3.2, busbars B5–B7 are positioned in a 90° angle with respect to the busbars B1–B4 in order to allow measuring of the temperature of an open DUT during the test with a thermal camera looking at it from above.

The inductances of loop 1 and loop 2 are estimated to be 45.6 nH and 59.7 nH, respectively (Smirnova et al., 2014). According to the experimental results, the inductance of loop 1 is 37 nH. The assembled set-up is show in Figure 3.3.
3.3 Busbar system of NDT set-up II: current-sharing issues

In this doctoral dissertation, a circular symmetry is proposed to obtain equal current sharing among the parallel devices of NDT set-up II. The designed circular layout is presented in Figure 3.4. As shown in Figure 3.4 (b), the arrangement of the components is similar to the one used for NDT set-up I (Figure 3.2 (b)).
The current distribution among the parallel devices in the layout shown in Figure 3.4 is analysed by modelling the busbar system with the power devices being replaced by copper stripes. The results shown in Figure 3.5 demonstrate that the current distribution among the parallel devices is uniform in the circular layout (current imbalance is 7%). The estimated inductance of loop 1 and loop 2 are 30.5 nH and 36 nH, respectively. A comparison of the circular layout with the traditional one is presented in (Smirnova et al., 2014). The traditional layout has the same placement of components as NDT set-up I (Figure 3.2). The current distribution in the traditional layout is shown in Figure 3.6, where the current concentration in the middle devices can be observed. The current imbalance may lead to a damage of these devices during short-circuit tests.

![Figure 3.5](image1.png)

**Figure 3.5.** Surface current density distribution of the circular busbar system of the NDT set-up II. The parallel components are replaced by copper stripes.

![Figure 3.6](image2.png)

**Figure 3.6.** Surface current density distribution of the traditional busbar system of NDT set-up II. The parallel components are replaced by copper stripes.
3.4 Summary

This section focused on the key issue of equal current sharing among parallel-connected devices, which arises in the design of a power electronics device with a parallel connection of switching devices for increasing power levels. Unequal current sharing generates a dangerous imbalance in the steady-state and transient operation of the parallel switching devices, which leads to the damage of the most stressed device. This issue affects the design of the laminated busbar system and the physical layout of the converter main circuit, which are modified to achieve not only a low stray inductance but also equal current sharing.
4 Thermal analysis of the laminated busbar system

This chapter deals with the thermal analysis of the laminated busbar system presented in Chapter 2. Thermal analysis is an important part of the busbar system design because temperature is the main limiting factor when determining how much current can flow in the busbars. Too high temperatures cause problems for the insulation system, degrade the conductors, and conduct extra heat to the semiconductor switches. However, too low temperatures mean that the busbar system is oversized, and an unnecessarily large amount of material is used. Thus, the busbar system should be designed in such a way that not only a low stray inductance is achieved but also the temperature limits defined by the manufacturers of the semiconductor devices applied are not exceeded during the converter operation. The allowed temperatures of the other components of the converter are usually higher than the temperature of the IGBT module terminals. For example, the manufacturer of the IGBT module applied in the converter under study allows a maximum temperature of the power terminals \( T_{\text{term}} = 125 \, ^{\circ}\text{C} \) (Infineon, 2013). Exceeding the thermal limits of the power module leads to undesired mechanical stresses, changes in geometry, and an increase in the temperature inside the module, and should be avoided to prevent negative effects on the reliability.

In order to estimate the temperature of the busbars, a proper thermal model is needed that can be used in the design stage to optimize the dimensions of the laminated busbars for the stray inductance and the cost and material minimization. To the author’s knowledge, the thermal analysis of the busbar system of the power converters is not sufficiently covered in the literature. However, the thermal analysis of the busbars used in power distribution systems, where numerical and analytical methods are applied, is presented. Because of the axial symmetry of such busbars and the fact that a sinusoidal or DC current flows through them, temperatures can be estimated analytically by directly solving the heat transfer equation or applying a 2D numerical tool. Analytical steady-state and transient analyses are described for instance in (Coneybeer et al., 1994), (Hus, 1990), (Plesca, 2012), and a 2D analysis is presented in (Hwang et al., 1998), (Popa et al., 2014). A 3D numerical analysis is usually applied to predict the temperature of critical regions such as the connections of the busbars (Wu et al., 2014) or structures having a complicated geometry such as the busbars of the low-voltage switchgear (Bedkowski et al., 2014). Despite the high accuracy of the numerical methods such as the FEM and the Computational Fluid Dynamics (CFD), they are rather expensive in terms of model set-up and computational time. This limits their application for optimization, where the computational speed is critical. In this sense, analytical methods are preferable because of their high computational speed. Also an iterative process taking into account the temperature-dependent parameters in the temperature estimation is easy to implement. However, an analytical approach requires accurate definition of the circuit that models the main heat transfer paths and a sufficient number of elements to ensure the required accuracy.

In this work, an analytical lumped parameter thermal model (LPTM) of the laminated busbar system is developed to solve the thermal problem by applying thermal networks,
Thermal analysis of the laminated busbar system

which have a straightforward analogy to electrical circuits. This method is widely used in
the design of electrical machines (Alexandrova et al., 2014), (Popova et al., 2011),
(Rostami et al., 2013). It allows fast and accurate prediction of the temperature of the
main machine parts. In this study, steady-state and transient analyses are performed. A
transient analysis is usually performed to determine the overloading capability of the
system, or it is used in the structural analysis for thermal stress evaluation.

4.1 Power losses and temperature estimation

For accurate thermal analysis, the power loss estimation is of primary importance because
the temperature changes in the busbar system are caused by losses, and also losses change
as a function of temperature. Hence, accurate loss estimation is impossible without
knowledge of the temperature.

In this study, the heat losses in the power converter that are dissipated through the busbar
system are considered and estimated. The primary source of power losses is the Joule
losses because of the resistance of the busbars. As the current flowing in the busbars of a
converter is alternating and non-sinusoidal, the current harmonics cause an increase in
the Joule losses. In the converter studied, the THD in the current flowing through a busbar
is up to 88%. Thus, the Joule loss calculation should take into account the losses caused
by the current harmonics. A fast Fourier decomposition of the non-sinusoidal current
flowing through the busbar in the converter with the defined parameters is performed to
obtain the spectrum of the current harmonics

\[
I_h = \frac{1}{\sqrt{2}} \cdot \text{abs} \left( \frac{\text{FFT}(i(t))}{n} \right),
\]  

(4.1)

where \( I_h \) is the RMS value of the \( h \)-th harmonic current, \( i(t) \) is the current flowing in the
busbars, and \( n \) is the number of samples.

The Joule losses of the busbars are calculated by

\[
P_b = \sum_{h=0}^{H} R_{b,h}(T_b(i)) \cdot I_h^2,
\]  

(4.2)

where \( P_b \) is the Joule loss in the busbar, \( R_{b,h} \) is the resistance of the busbar at a temperature
\( T_b(i) \) and a frequency corresponding to the \( h \)-th harmonic.

The resistance of the busbars at different temperatures and frequencies is estimated by a
numerical tool to obtain a 3D lookup table, where the skin and proximity effects are given.
The electrical resistivity has a significant temperature variation, and the linear variation of the resistivity is assumed

\[ \rho = \rho_0 [1 + \alpha (T_h - T_a)], \]  

(4.3)

where \( T_a \) is the ambient temperature, \( \alpha \) is the coefficient of electrical resistivity variation with temperature, \( \rho \) is the electrical resistivity, and \( \rho_0 \) is the electrical resistivity at ambient temperature. In the Joule loss calculation, the AC resistance of the busbar at the required temperature and frequency is retrieved from the lookup table.

There are two additional sources of power losses in the busbar system. The first additional source of power losses is the Joule losses caused by the contact resistance at each connection of the busbar to the IGBT or capacitor terminals. These losses constitute a considerable part of total losses in the laminated busbar and should be carefully estimated for an accurate thermal analysis. In Figure 4.1, the connection used in the busbar system under study is presented. The copper and aluminium pipes of different lengths are applied to connect the terminals of the converter components to the copper busbars.

![Figure 4.1. Connection between the busbar and the IGBT module terminal. (a) 3D view. (b) Cross-sectional view.](image)

Because of the roughness of the metal surface at the interface, the contact appears at discrete spots when mechanical force is applied. The current flows only through these spots, and thus, a constriction resistance arises. The constriction resistance is determined by

\[ R_c = \frac{\rho_1 + \rho_2}{4r}, \]  

(4.4)
where $\rho_1$ and $\rho_2$ are the resistivities of the contacting metals and $r$ is the radius of the real contact area $S_c$. The real contact area $S_c$ is generally much smaller than the apparent contact area, and it is found by

$$S_c = \frac{F}{H},$$  \hspace{1cm} (4.5)

where $F$ is the force applied and $H$ is the hardness of the softer of the two contacting metals (Slade, 2014). The applied force $F$ can be calculated for the bolt connection by

$$F = \frac{T}{k_n \cdot 2 \cdot r_b},$$  \hspace{1cm} (4.6)

where $T$ is the applied torque, $k_n$ is the nut factor, and $r_b$ is the radius of the bolt. The parameters used for the constriction resistance calculation are presented in Table C.1 of Appendix C.

The calculation of the constriction resistance is made with the assumption that the metal surfaces are perfectly flat and clean. However, an oxide layer grows quickly on the base metals such as copper and aluminium exposed to air, and the film resistance $R_f$ should be added to the constriction resistance $R_c$ to obtain the total contact resistance $R_t$ (Braunovic, 2002)

$$R_t = R_c + R_f.$$  \hspace{1cm} (4.7)

Calculation of the film resistance is difficult because of the uncertainties of the film characteristics. For example, the thickness of the film affects the method by which the current is conducted through the film. The contact resistances of the connections applied in the busbar system under study have been measured by a Potentiostat/Galvanostat/Zero-Resistance Ammeter (Gamry Reference 3000) capable of measuring a resistance of several micro ohms with a 90% accuracy (Gamry Instruments, 2012).

The Joule loss $P_c$ resulting from the contact resistance $R_t$ at the temperature $T_{b(i)}$ is calculated by

$$P_c = R_t \sum_{k=0}^{n} I_k^2,$$  \hspace{1cm} (4.8)
and added as an additional heat source to the thermal model.

Another source of additional losses in the busbar system is the Joule losses caused by the resistance of the main terminals of the converter components \( P_{\text{term}} \). The heat generated in the terminals is also dissipated through the busbars. The resistance of the main terminals of the IGBT module is given in the datasheet of the device as the lead resistance of the module \( R_{CC'+EE'} \). However, some manufacturers include the resistance of the bond wires in the lead resistance of the module. If the geometry of the component is known, the resistance can be estimated. The resistance of the capacitor terminals is seldom given in the datasheets. In this study, the resistances of the component terminals have been estimated by a numerical tool and measured. The Joule losses are calculated by Equation (4.2) and also added as heat sources to the thermal model.

### 4.2 Lumped parameter thermal model

For the temperature estimation, the LPTM of the busbars is developed. The analysis by the LPTM is based on the calculation of the thermal resistances and capacitances of the busbar parts. It is also possible to consider the heating of the cooling fluid by adding a cooling matrix. The advantage of the LPTM is in its simple mathematical form and high computational speed. Further, the model can be easily implemented in a circuit simulator because of its analogy to an electrical circuit.

The following assumptions are made in the analysis by the LPTM: the heat generated in the element is dissipated by convection and radiation; the heat is uniformly distributed in the element volume; there are independent heat transfers in the \( x \), \( y \), \( z \) directions so that all axes can be analysed individually.

#### 4.2.1 Heat transfer mechanisms

In order to consider the heat transfer by conduction in the conducting busbar with internal heat generation, the thermal network in the \( x \)-direction for a cubical element introduced in (Wrobel and Mellor, 2010) and presented in Figure 4.2 is applied. This thermal network can be used to represent each axis.
Thermal analysis of the laminated busbar system

Figure 4.2. (a) Cubical element. (b) Equivalent thermal network of the cubical element. \( T_1 \) and \( T_2 \) are surface temperatures, and \( T_m \) is the mean temperature assuming only heat transfer in the x-axis direction (Wrobel and Mellor, 2010).

The thermal resistances \( R_{x1} \) and \( R_{x2} \) of the network are calculated from the solution of the heat equation for a cubical element with zero internal heat generation by

\[
R_{x1} = R_{x2} = \frac{l_x}{2k_xS_x},
\]  

where \( l_x \) is the length of the cubical component in the x-direction, \( k_x \) is the thermal conductivity in the x-direction, \( S_x \) is the cross-sectional area normal to \( q_x \), which is the heat transfer rate in the x-direction. When there is no internal heat generation in an element, the heat transfer can be considered by only two resistances \( R_{x1} \) and \( R_{x2} \), as for example in the insulating layers of the busbar system. In order to correctly determine the mean temperature of the element with internal heat generation such as a copper busbar, a third resistor \( R_{x3} \) has been introduced in (Wrobel and Mellor, 2010). The third resistor value is calculated based on the mean temperature obtained from the solution of the heat equation for the cubical element with internal heat generation and given as follows

\[
R_{x3} = -\frac{l_x}{6k_xS_x}.
\]  

The resistor \( R_{x3} \) with a negative sign allows obtaining a correct mean temperature while a standard network consisting of only the resistors \( R_{x1} \) and \( R_{x2} \) would overestimate the mean temperature of the element with internal heat generation (Wrobel and Mellor, 2010).
A complete 3D network of the cubical element is presented in Figure 4.3, where the thermal capacitance and the heat source are connected to the point representing the mean temperature of the element $T_m$.

Figure 4.3. 3D network of the cubical element (Wrobel and Mellor, 2010). The grey node shows the place where the heat source is concentrated.

The thermal capacitance of the element is calculated by

$$C = c \rho V,$$

(4.11)

where $c$ is the specific heat capacity, $\rho$ is the density of the element material, and $V$ is the volume of the element.

The convective heat transfer is modelled by a single thermal resistance $R_{\text{conv}}$

$$R_{\text{conv}} = \frac{1}{h_c S},$$

(4.12)

where $h_c$ is the convective coefficient and $S$ is the surface area of the convective heat transfer between two regions. The convective coefficient $h_c$ is calculated based on the ambient air properties and the busbar system dimensions as
Thermal analysis of the laminated busbar system

\[ h_{\ell} = \frac{k_{\text{air}}Nu}{l}, \quad (4.13) \]

where \( k_{\text{air}} \) is the thermal conductivity of air, \( Nu \) is the Nusselt number, and \( l \) is the length of the busbar system.

The Nusselt number for a vertical plate is calculated by an equation presented in (Incropera, 2006)

\[ Nu = \left\{ 0.825 + \frac{0.387Ra^{1/6}}{1 + (0.492 / Pr)^{9/16}} \right\}^2, \quad (4.14) \]

where \( Ra \) is the Rayleigh number and \( Pr \) is the Prandtl number. For free or natural convection, the Rayleigh number is

\[ Ra_N = Gr \cdot Pr = \frac{g\beta(T_s - T_a)l^3}{\nu\alpha_{\text{air}}}, \quad (4.15) \]

where \( g \) is the gravitation constant, \( \beta \) is the coefficient of thermal expansion of air, \( T_s \) and \( T_a \) are the surface and ambient temperatures, respectively, \( \nu \) is the air kinematic viscosity, \( \alpha_{\text{air}} \) is the thermal diffusivity of air, and \( Gr \) is the Grashof number. The Prandtl number is calculated by

\[ Pr = \frac{\nu}{\alpha_{\text{air}}}. \quad (4.16) \]

For forced convection, the Rayleigh number is calculated as follows

\[ Ra_F = Re^2Pr, \quad (4.17) \]

where the Reynolds number \( Re \), which defines whether the air flow regime, laminar or turbulent at a particular air velocity \( u_{\text{air}} \), is obtained by
\[ Re = \frac{\nu \mu}{\nu}. \] (4.18)

If \( Gr/Re^2 \ll 1 \), the effect of free convection can be neglected, and if \( Gr/Re^2 \gg 1 \), the effect of forced convection can be neglected. A combined free and forced convection should be considered if \( Gr/Re^2 \approx 1 \). The Nusselt number for combined convection \( Nu_L \) is then calculated as follows (Incropera, 2006)

\[ Nu_L = Nu_f + Nu_N, \] (4.19)

where \( Nu_f \) is the Nusselt number calculated for a pure free convection and \( Nu_N \) for a pure forced convection case.

The heat transfer by radiation is also considered by a single resistance as in the case of convection, but the convective coefficient is replaced in Equation (4.12) by the equivalent radiation coefficient \( h_r \)

\[ h_r = \sigma a \frac{(T_i^4 - T_s^4)}{(T_i - T_s)}, \] (4.20)

where \( \sigma \) is the Stefan-Boltzmann constant and \( \varepsilon \) is the emissivity of the surface.

The heat transfer between two adjacent objects is modelled by a contact thermal resistance \( R_{cont} \). As a result of the roughness of adjacent surfaces, a small air gap with a length \( l_6 \) occurs between the two objects, which has a final thermal resistance (Pyrhönen et al., 2008)

\[ R_{cont} = \frac{l_6}{k_{air} S}. \] (4.21)

The length of the air gap \( l_6 \) can be estimated by summing up the roughnesses of the contacting surfaces.

As it is shown in Figure 4.6, the convective and radiation coefficients are calculated at every iteration based on the temperatures estimated by the LPTM. The thermophysical properties of the air at different temperatures are shown in Table C.2 of Appendix C.
4.2.2 LPTM of the ANPC converter busbar system

In this section, the 3D LPTM for the busbar system of the ANPC converter described in Chapter 2 (Figure 2.1) is developed. The cross section of the busbar system consisting of seven layers is illustrated in Figure 4.4. The conducting layers are made of copper with a thermal conductivity 394 W/(mK), and the insulating layers are made of Teflon with a thermal conductivity 0.2 W/(mK). The operating temperature range of Teflon is from –190 to +250 °C.

Figure 4.4. Cross-sectional view of the laminated busbar system. P – positive busbar, N – negative busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link, A3 – additional busbar of the lower phase arm, A4 – additional busbar of the lower part of the DC link. The conducting layers are 2, 4, and 6, and the insulating layers are 1, 3, 5, and 7.

Figure 4.5. (a) One quarter of the laminated busbar system. (b) Exploded view of the busbar system, the division of the layers into the cubical elements is shown. P – positive busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link, A3 – additional busbar of the lower phase arm. The conducting layers are made of 2 mm thick copper, and the insulating layers are made of 1 mm thick Teflon.
Because of symmetry, only one quarter of the busbar system is modelled (Figure 4.5). As shown in Figure 4.5, the solid components of the busbar system are modelled by 137 cubical elements to consider conductive heat transfer: the cubical elements representing the copper busbars involve heat generation while the elements representing the insulating layers do not. The thermal capacitance of the element and the heat generated in the element are concentrated in one point, which represents a mean temperature $T_m$. As the assumption of uniformly distributed heat losses is made in the analysis by the LPTM, the Joule losses of the busbar $P_b$ are distributed among the cubical elements of this busbar in proportion to the volume of each element. The Joule loss caused by the resistance of the component main terminal $P_{term}$ and the Joule loss resulting from the contact resistance $P_c$ of the connection between the main terminal and the busbar both are concentrated in the cubical element of the copper busbar where this connection is located. The cubical elements of the inner layers of the busbar system are connected to each other using contact thermal resistances. The surrounding air is modelled as a single element with a constant temperature ($T_a$), which is connected to the cubical elements of the outer layers of the busbar system by thermal resistances representing heat transfer by convection and radiation. The heat generated in the laminated busbar system is removed by free and forced convection. From the front surface of the busbar system (layers 6 and 7), the heat is removed by free convection. The back surface of the busbars system (layer 4) is indirectly blown over by fans, which are installed for cooling of the IGBT modules. Since the air velocity is low (about 0.2 m/s), the heat is transferred by combined free and forced convection, for this regime $Gr/Re^2 \approx 1$.

### 4.3 Thermal analysis

Figure 4.6 presents the algorithm applied to the calculation of the power loss and temperature of the busbar system. The temperature dependence of the power losses and the thermal parameters is considered by implementing an iterative procedure until the specified value of error between the current and previous values of the estimated temperature is achieved. The temperatures obtained from the LPTM are iteratively updated for re-calculation of the temperature-dependent Joule losses and thermal parameters.

The thermal calculations can be performed for steady states only, but several different steady-state analyses are needed depending on the operating point of the converter. The converter modulation has a significant impact on the current distributions in different switches and layers of the converter busbar system. Therefore, in principle, different voltage and power factor values would need steady-state thermal simulations of their own.
The temperature rise, which is the difference between the element and ambient temperatures, is used to represent the element. The temperature rise of each element of the thermal model is computed by the matrix equation

\[ \Delta T = G^{-1} \cdot P, \] (4.22)
where $\Delta \mathbf{T}$ is the temperature rise vector, $\mathbf{P}$ is the vector including the Joule losses of each element, and $\mathbf{G}$ is $n \times n$ thermal conductance matrix, where $n$ is the number of elements in the model. The conductance matrix is defined by

$$
\mathbf{G} = \begin{bmatrix}
\sum_{i=1}^{n} \frac{1}{R_{1,i}} & -\frac{1}{R_{1,2}} & \cdots & -\frac{1}{R_{1,n}} \\
-\frac{1}{R_{2,1}} & \sum_{i=1}^{n} \frac{1}{R_{2,i}} & \cdots & -\frac{1}{R_{2,n}} \\
\vdots & \vdots & \ddots & \vdots \\
-\frac{1}{R_{n,1}} & -\frac{1}{R_{n,2}} & \cdots & \frac{1}{R_{n,n}} \\
\end{bmatrix}, \quad (4.23)
$$

where the $n^{th}$ diagonal element is the sum of the conductances connected to the element $n$, and $G(i,j)$ is the thermal conductance-connecting elements $i$ and $j$ with a minus sign (Nerg et al., 2008). When the order of the model becomes high, it is easier to implement it in a circuit simulator.

In the transient analysis, the vector of the element temperatures $\mathbf{T}$ at each time step $\Delta t$ can be calculated in a discretized form as (Hey et al., 2011)

$$
\mathbf{T}^{t+\Delta t} = \mathbf{T}^{t} + \Delta t \mathbf{C}' (\mathbf{G} \cdot \mathbf{T}^{t} + \mathbf{P}), \quad (4.24)
$$

where the temperatures of the next iteration $\mathbf{T}^{t+\Delta t}$ depend only on the temperature of the current iteration $\mathbf{T}^{t}$, and $\mathbf{C}$ is the diagonal matrix of the thermal capacitances of the cubical elements. Equation 4.24 is the difference equation discretized from the energy balance equation

$$
\mathbf{C} \mathbf{T} = \mathbf{G} \mathbf{T} + \mathbf{P}, \quad (4.25)
$$

which determines the temperature change of each element based on the balance of the generated losses, the stored energy, and heat flowing in and out of the element. The loss vector $\mathbf{P}$ is calculated in the steady-state analysis by the algorithm presented in Figure 4.6.

4.3.1 Simulation results

The losses and temperatures of the busbars are estimated using the algorithm presented in Figure 4.6 for the ANPC converter operated at the nominal power with the parameters
Thermal analysis of the laminated busbar system

listed in Table 4.1. The converter is modelled in PLECS/Simulink, the modulation method used is the level-shifted pulse width modulation (LSPWM), and the ambient temperature is assumed to be 25 °C. The power losses obtained are shown in Table 4.2. The total losses include the Joule losses resulting from the resistance of the busbars, the terminals, and the contact resistance. The Joule losses caused by the contact resistance and the resistance of the main terminals constitute a considerable part of the total losses, because the resistance of the busbars that have large cross sections for the current flow is low compared with the resistance of the connections and the main terminals.

Table 4.1. Parameters of the ANPC converter used in the simulation.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC link voltage, V</td>
<td>1100</td>
</tr>
<tr>
<td>modulation index</td>
<td>0.85</td>
</tr>
<tr>
<td>power factor</td>
<td>1</td>
</tr>
<tr>
<td>phase current rms, A</td>
<td>150</td>
</tr>
<tr>
<td>switching frequency, Hz</td>
<td>4000</td>
</tr>
<tr>
<td>fundamental frequency, Hz</td>
<td>50</td>
</tr>
</tbody>
</table>

The results obtained with the LPTM have been compared with the results obtained by 3D numerical simulations in a FEM-based software (Comsol Multiphysics) in order to verify that the LPTM correctly models the heat transfer in the busbar system. The losses presented in Table 4.2 have been used as heat sources for the thermostatic model. Because of symmetry, one quarter of the laminated busbar system shown in Figure 4.5 is modelled, and symmetric boundary conditions are imposed. The convective and radiation coefficients are defined as boundary conditions on the outer surfaces. The thermal coefficients, the conductivities of materials, and the dimensions are the same as in the LPTM. The resultant temperature distribution of the laminated busbar system is shown in Figure 4.7. The mean temperatures evaluated by the LPTM and the FEM are compared in Table 4.2, where a good correlation between the results is observed. The obtained temperatures of the busbars are much lower than allowed (125 °C), and thus, the busbar system studied is oversized. Consequently, the busbar system can be used for a 60% higher current than it is designed for.

Table 4.2. Comparison of the results obtained by the LPTM and FEM models. P – positive busbar, NT – neutral busbar, Ph – phase out busbar, A1 – additional busbar of the upper phase arm, A2 – additional busbar of the upper part of the DC link. $P_b$ is the Joule losses in the busbar, $P_c$ is Joule losses caused by the contact resistance, and $P_{term}$ is the Joule losses caused by the component terminals.

<table>
<thead>
<tr>
<th></th>
<th>Power losses, W</th>
<th>Estimated mean temperature, °C</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$P_b$</td>
<td>$P_c$</td>
</tr>
<tr>
<td>P busbar</td>
<td>2.3</td>
<td>4.0</td>
</tr>
<tr>
<td>NT busbar</td>
<td>1.5</td>
<td>2.7</td>
</tr>
<tr>
<td>A1 busbar</td>
<td>0.6</td>
<td>1.1</td>
</tr>
<tr>
<td>Ph busbar</td>
<td>1.3</td>
<td>0.4</td>
</tr>
<tr>
<td>A2 busbar</td>
<td>0.08</td>
<td>1.7</td>
</tr>
</tbody>
</table>
Figure 4.7. (a) FEM thermostatic model of one quarter of the laminated busbar system. (b) Exploded view of the busbar system, the copper busbars are shown. A1 – additional busbar of the upper phase arm, A3 – additional busbar of the lower phase arm, NT – neutral busbar, A2 – additional busbar of the upper part of the DC link, Ph – phase out busbar, P – positive busbar. The volume temperature is depicted.

Figure 4.8. Estimated temperatures of the laminated busbar system. P – positive busbar, NT – neutral busbar, A1 – additional busbar of the upper phase arm, Ph – phase out busbar, A2 – additional busbar of the upper part of the DC link.
Thermal analysis of the laminated busbar system

The estimated mean temperatures of the busbars as a function of time are shown in Figure 4.8. As shown, the time needed to reach a steady state is about 50 minutes.

4.4 Experimental verification

In order to validate the proposed algorithm and the obtained simulation results, the temperatures of the busbar system have been measured. The measurement set-up is shown in Figure 4.9. In the experiment, the ANPC converter initially designed to operate as a 150 kVA grid-side converter was operated as a motor-side converter driving a 55 kW induction machine with a 50% load. The parameters of the converter in the experiment are listed in Table 4.3. The measured output phase current and voltage of the converter in the experiment are shown in Figure 4.10.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC link voltage, V</td>
<td>600</td>
</tr>
<tr>
<td>modulation index</td>
<td>0.6</td>
</tr>
<tr>
<td>power factor</td>
<td>0.86</td>
</tr>
<tr>
<td>phase current rms, A</td>
<td>65</td>
</tr>
<tr>
<td>switching frequency, Hz</td>
<td>2000</td>
</tr>
<tr>
<td>fundamental frequency, Hz</td>
<td>25</td>
</tr>
</tbody>
</table>

Figure 4.9. Measurement setup. 1 – a 55 kW induction machine, 2 – ANPC converter, 3 – control electronics, 4 – oscilloscope showing converter output current and voltage, 5 – data processing computer.
The temperatures have been measured when the converter has operated long enough, and the temperatures are nearly stable (approximately 70 minutes). For the temperature measurements, Pt-100 thermal sensors (measurement error is ±1 °C) are used that have been placed in the same positions as the reading point in the thermal models. The ambient temperature is 24 °C. Table 4.4 shows a comparison between the measured temperatures and the temperatures estimated by the algorithm presented in Figure 4.6.

<table>
<thead>
<tr>
<th>Busbar Type</th>
<th>Estimated temperature, °C</th>
<th>Measured temperature, °C</th>
</tr>
</thead>
<tbody>
<tr>
<td>P busbar</td>
<td>30.6</td>
<td>32.0</td>
</tr>
<tr>
<td>NT busbar</td>
<td>31.3</td>
<td>32.7</td>
</tr>
<tr>
<td>A1 busbar</td>
<td>32.7</td>
<td>34.3</td>
</tr>
<tr>
<td>Ph busbar</td>
<td>32.3</td>
<td>33.2</td>
</tr>
<tr>
<td>A2 busbar</td>
<td>30.1</td>
<td>31.5</td>
</tr>
</tbody>
</table>

There is a good correlation between the estimated and experimental temperature rises taking into account the assumptions made in the model.

### 4.5 Discussion

The LP electrothermal model of the proposed laminated busbar system was validated both by a FEM analysis and measurements and can be used to analyse the loss and temperature distribution in the laminated busbar system in the design and optimization of a power converter.
In order to improve the thermal performance of the busbar system, the heat generated or/and the thermal resistances should be decreased. In the studied laminated busbar system, the electrical connections are found to be the main heat sources. Thus, the temperature decrease can be achieved by improving these connections and minimizing the Joule losses caused by the contact resistance. The method known to reduce the contact resistance is to expand the contact area by applying a greater contact force, and cleaning or/and finishing of the contact surfaces (Braunovic, 2002). The conducting pastes and liquid metals, which have high electrical and thermal conductivity can be used to reduce the contact resistance (Zhang, and Liu, 2013), (Leong and Chung, 2004). The contact resistance could also be minimized for instance by using gold plating of the contact parts to prevent corrosion caused by environmental conditions. If the contact resistances are decreased to 10 μΩ in the busbar system under study, the temperatures of the busbars are reduced considerably in comparison with the temperatures presented in Table 4.2, where the nominal resistance values are used. The comparison is shown in Figure 4.11.

![Figure 4.11](image-url)

Figure 4.11. Comparison of the results obtained by the LPTM with the nominal value of contact resistances and reduced to 10 μΩ. P, NT, A1, Ph, and A2 indicate the busbars.

With small contact resistances (10 μΩ), the busbar system can conduct current by 73% more than it is designed for without exceeding 125 °C. Thus, the nominal power of the converter can be increased to 260 kW without applying additional cooling solutions.

Another way to obtain a better heat dissipation is to increase the convective coefficient by means of a fan. By ventilating the busbar system by a fan which produces a 3 m/s air velocity, the convective coefficient value increases to 20 W/(m²K) for the vertically placed busbar system. Figure 4.12 shows the change in the busbar temperatures compared
with the temperatures presented in Table 4.2, where the convective coefficient is about 5 W/(m² K). As shown, a significant temperature reduction is possible by applying forced air cooling. The heat dissipation can be further improved by increasing the emissivity of copper, which is low for copper without a thick oxide layer. Painting of the copper considerably increases the emissivity and thereby the heat dissipation.

![Figure 4.12. Influence of the convective coefficient value on the temperature of the busbars. P, NT, A1, Ph, and A2 indicate the busbars.](image)

Extreme situations such as an ambient temperature increase, overmodulation, and peak currents should be considered in the design of the converter. In some applications, the peak current may occur for a limited time only. In that case, by performing a transient analysis, one can verify if the busbar system can be designed for a lower rated current. It should also be mentioned that the contact resistance experiencing thermal cycles during the converter operation is not constant but increases with time (Sun et al., 1999), (Koktsinskaya et al., 2014). This fact should also be considered in the design.

### 4.6 Summary

This chapter focused on the development of a tool for an accurate thermal analysis of the laminated busbar system used in power electronics to be used in the design. A comprehensive electrothermal model has been developed for the estimation of power losses and temperatures in the busbar system on the basis of an analytical thermal model. All relevant aspects of the model development have been discussed in detail. The estimation of heat sources in the busbar system was considered. An LPTM that takes into
account all major heat transfer mechanisms has been developed for the busbar system under study. The FEM simulations verified the suitability of the analytical model. The measurements performed demonstrated the validity of the proposed electrothermal model.
5 Thermal modelling and reliability analysis of IGBT modules

The maximum junction temperature of the semiconductor device specified in the datasheet limits the power capability of the converter, and thus, the power density. Nowadays, silicon-based components can normally be used with maximum junction temperatures up to 200°C (Millan et al., 2014). Exceeding the temperature limits can lead to a sudden device failure or at least to a significantly shorter lifetime. In order to increase the power capability, an efficient cooling system is required that can dissipate a large amount of losses without overheating.

The cooling solution for the power converter is designed to achieve the maximum power density. Usually, the requirements for the cooling solution are set based on the static thermal model of the converter. When the converter is used in applications with varying loads such as wind turbines, PV panels, hybrid, or electrical vehicles, a reliability analysis is required because the IGBT devices are subjected to a high number of different thermal cycles (Sintamarean et al., 2015), (Ma et al., 2015). A reliability assessment is needed at the design stage to select a proper cooling solution that not only provides the maximum power density but also meets the reliability requirements. Thus, thermal modelling has become an essential part of the converter design. A dynamic thermal model is particularly important for the lifetime estimation, because dynamic temperature changes (temperature swing) have a significant contribution of their own to the lifetime reduction such as the absolute temperature level (Zhou et al., 2014).

Thermal cycles are the primary cause of the fatigue of the IGBT modules (Busca et al., 2011). Temperature variations cause expansion and contraction of materials with different coefficients of thermal expansion (CTE) inside the module, eventually leading to failures such as bond wire lift-off and the fatigue of solder layers and ceramics (Ciappa, 2001). Thermal cycles are caused by factors such as environmental conditions, control, and power variations. Thus, the expected lifetime of the semiconductor device strongly depends on the application. In wind power applications, the IGBTs are subjected to a variety of load profiles, which cause cyclic thermomechanical stress in all components of the module, finally leading to device failures. Meanwhile, IGBT module failures are critical to the wind power system, and the repair cost is high especially for wind turbines installed off-shore. In order to estimate the expected lifetime of an IGBT, the PoF method is applied. It allows determining how much temperature loading the IGBT can tolerate.

In this doctoral dissertation, a grid-side wind turbine converter is taken as a case study. In this chapter, the static and dynamic models of the converter cooling system are developed. The static model is used to specify the requirements for the cooling system while the dynamic model is used to obtain the temperature load profile of the IGBT, which will be used further for the lifetime estimation of the IGBT in the grid-side power converter of the wind turbine. Different cooling solutions used nowadays with power converters are reviewed and compared. The following study aims at indicating and
Thermal modelling and reliability analysis of IGBT modules

improving the reliability of the power converter. The influence of the thermal capacitance of the cooling system on the temperature load profile of an IGBT module is investigated. Special attention is paid to the influence of the cooling system time constant on the reliability of the IGBT module. Three cooling solutions are compared in terms of reliability. By performing a reliability analysis at the design stage, the choice of the cooling solution for a particular application can be made taking into account the mission profile and the required reliability. A control strategy for the cooling system is proposed to improve the reliability of the semiconductor devices based on heat flux sensor measurements.

5.1 Power losses

In the converter operation, the IGBTs and the diodes generate losses that heat the devices. Proper cooling measures must be taken to dissipate the generated losses and keep the junction temperature $T_j$ of the device at an acceptable level. The losses are estimated in the converter design to specify the requirements for the cooling system.

The total losses of the IGBT and the diode are comprised of the conduction, switching, and driving losses.

$$P_t = P_{\text{cond}} + P_{\text{sw}} + P_d.$$  \hspace{1cm} (5.1)

The losses $P_d$ produced in the gate driver are usually small and can be neglected. The average conduction losses are obtained by

$$P_{\text{cond}} = \frac{1}{T} \left( \int (U_{CE,0}(T_j) + R(T_j, U_{DC})i(t)) \cdot i(t) \, dt \right),$$  \hspace{1cm} (5.2)

where $U_{CE,0}$ is the collector-emitter threshold voltage, which varies with the junction temperature, $R$ is the on-state resistance, which depends on the junction temperature and the DC link voltage $U_{DC}$, and $i$ is the collector current.

Switching losses occur during the turn-on and turn-off of the device and depend on the switching frequency $f_{sw}$

$$P_{\text{sw}} = E_{sw}(T_j, U_{DC}, I) \cdot f_{sw},$$  \hspace{1cm} (5.3)
where $E_{sw}$ is the energy of one switching, which is the sum of the turn-on energy $E_{on}$ and the turn-off energy $E_{off}$. The switching energy curves for different values of junction temperature, collector current, and DC link voltage are provided by the manufacturers of the switching devices. The proportion of the conduction and switching losses in the device depends on the operating conditions of the converter such as modulation index, power factor, and switching frequency.

In this work, the loss model presented in (Blaabjerg et al., 1995), (PLECS, 2014) is applied and simulations are carried out in the PLECS Toolbox of Matlab/Simulink. The estimated losses are temperature dependent and based on the datasheet parameters of the devices. The thermal model of the semiconductor device and the cooling system is also implemented in the PLECS to provide temperature feedback for the loss model.

In order to specify the requirements for the cooling system, the losses of the semiconductor devices are estimated at the maximum possible apparent power of the converter. The reactive power delivered by the wind turbine has to be regulated in a certain range defined by the regulations concerning the performance of wind power converters under normal and fault conditions. These regulations are developed in many countries, and the widest ranges are found in the German grid code, where three variants of the allowed boundaries of reactive power versus active power are defined as presented in Figure 5.1 (Altun et al., 2010). Usually, the variant is chosen in agreement with the grid operator. As it can be seen in Variant 1, the underexcited reactive power should be at least 23% of the rated active power $P_{\text{rated}}$, and the overexcited reactive power should be at least 48% of $P_{\text{rated}}$.

![Figure 5.1. $P, Q$ range of the wind power converter defined by the German grid code (Altun et al., 2010).](image)

At an early design stage, the losses are evaluated at the highest allowed junction temperature to determine the required capability of the cooling system. Then, the cooling
Thermal modelling and reliability analysis of IGBT modules

solution able to dissipate the amount of losses estimated and provide the junction temperature lower than allowed is selected.

Accurate loss estimation is possible only with the thermal model providing the estimated junction temperature. Further, static and dynamic thermal models of the power devices and the cooling solution are derived.

5.2 Requirements for the cooling system

The requirements for the cooling system are specified in terms of thermal resistance. The calculation of the required thermal resistance of the cooling solution is based on the static thermal network shown in Figure 5.2, which models the thermal behaviour of the IGBT module (containing one IGBT and a freewheeling diode) placed on the heat sink (Figure 5.3). The model is simplified and the thermal resistances between the elements inside the module and between the modules on the heat sink are not considered. Inclusion of these elements makes the model complicated and difficult to parameterize. When several modules are placed on a single heat sink, their models are added to the point representing the heat sink temperature $T_h$ shown in Figure 5.2. The total losses dissipated through the heat sink $P_{t,\text{loss}}$ are the sum of losses of every device on that heat sink.

The thermal resistance of the cooling system $R_{(h-a)}$ is selected to guarantee that the maximum junction temperature $T_{j,\text{max}}$ of every semiconductor on the heat sink calculated as follows is not exceeded when the average power losses $P_{\text{loss}}$ are generated in the device

$$T_{j,\text{max}} = T_{j,m} + \Delta T_j, \quad (5.4)$$

where $T_{j,m}$ is the mean value of the junction temperature obtained from Figure 5.2 as

$$T_{j,m} = P_{\text{loss}} \cdot (R_{(j-c)} + R_{(c-h)}) + P_{t,\text{loss}} \cdot R_{(h-a)} + T_a \quad (5.5)$$

and $\Delta T_j$ is the amplitude of the junction temperature variation during the fundamental period of the converter output current, which can be analytically estimated by

$$\Delta T_j = P_{\text{loss}} \cdot Z_{(j-c)} \left( \frac{3}{8f_o} \right) + 2P_{\text{loss}} \cdot Z_{(j-c)} \left( \frac{1}{4f_o} \right), \quad (5.6)$$
where \( Z_{(j-c)} \) is the time-based thermal impedance of the device provided by the manufacturer and \( f_o \) is the fundamental frequency (Ma et al., 2015).

![Figure 5.2. Static thermal model of the IGBT and the diode inside the IGBT module.](image)

![Figure 5.3. Cross section of the IGBT module placed on a heat sink. TIM is the thermal interface material such as silicon grease.](image)

By minimizing the thermal resistances included in the model, a higher amount of losses can be dissipated without exceeding \( T_{j,\text{max}} \), thereby increasing the power density of the converter by increasing the current or switching frequency.

Thus, the selection of the IGBT modules with small thermal resistances \( R_{(j-c)} \) is preferred. The resistance \( R_{(j-c)} \) of the module depends on the thickness and conductivity of the layers
inside the module (from silicon chips to the copper baseplate in Figure 5.3). As discussed in (Ji et al., 2015), the layers that contribute most to the thermal resistance $R_{(j-c)}$ are the solder and the ceramic isolation. In recent years, the focus was on the improvement of their thermal performance. Thus, new materials with high thermal conductivity such as aluminium nitride (AlN) for the ceramic isolation and 95Pb-5Sn or 96Sn-4Ag for the solder have been introduced (Wang et al., 2013).

The thermal interface material (TIM) may have a high thermal resistance $R_{(c-h)}$ and should be minimized or even eliminated. The TIM is usually thermal grease or paste applied between the baseplate of the module and the surface of the heat sink to improve the thermal contact conductance by replacing the air, which has a thermal conductivity of 0.025 W/(m·K) (at 25 °C) while a grease or paste typically has a thermal conductivity of approximately 1 W/(m·K). The manufacturers of IGBT modules suggest different techniques to improve the distribution of the TIM and decrease the thermal resistance (Schulz, 2011), (ABB, 2014). However, as it is discussed in (Skuriat et al., 2013), even a properly applied TIM is subjected to ageing and degradation, which increase its thermal resistance with time. Over the last decade, there was also a trend of eliminating the TIM by applying integrated cooling solutions such as integrated liquid-cooled cold plates or micro-heat exchangers. This allows a considerable improvement in the thermal performance.

The thermal resistance of the cooling solution $R_{(h-a)}$ depends on many factors such as the type of the cooling medium (air or liquid), the type of convection (natural or forced), the geometry of the heat sink, the thermal properties of the materials, and the speed of the fan or pump. These factors affect not only the thermal resistance but also the cost, volume, weight, and reliability of the cooling solution. Hence, there are many limitations that have to be considered in the selection of the cooling solution for a specific application. In Section 5.4, different cooling solutions are reviewed.

### 5.3 Dynamic thermal model

Accurate dynamic thermal modelling is essential for the lifetime estimation of the IGBT modules. It allows obtaining the temperature load profile of the semiconductor device based on the mission profile of the converter.

There are two types of thermal models, Cauer and Foster, commonly used for the dynamic thermal modelling of the IGBT modules. Both models use the thermal resistance $R$ and the capacitance $C$ that are analogous to the electrical resistance and capacitance. Thus, the models can be solved in any circuit simulator.

The Cauer model shown in Figure 5.4 allows obtaining a realistic physical representation of the transient thermal behaviour (Yun et al., 2001). With the knowledge of the IGBT module internal geometry (Figure 5.3) and the properties of the material used, the $R$ and $C$ values of each layer of the model can be calculated as follows.
\[ R = \frac{l}{k \cdot S}, \quad (5.7) \]

\[ C = c_p \cdot \rho \cdot l \cdot S, \quad (5.8) \]

where \( R \) is the thermal resistance of the layer, \( l \) is the length of the layer, \( k \) is the thermal conductivity of the material, \( S \) is the area of the layer, \( C \) is the thermal capacitance, \( c_p \) is the heat capacity of the material, and \( \rho \) is the density of the material.

In order to achieve good accuracy, a sufficient number of layers should be used so that the heat is evenly spread in each layer. The Cauer model allows extending the model of the IGBT modules and adding the layers representing the TIM and the heat sink. Thus, a complete model is obtained, where the junction and case temperatures can be estimated.

\[ \text{Figure 5.4. Cauer thermal model with } n \text{ layers.} \]

However, the internal geometry of the IGBT modules is usually not known, and the designers of the power converter have to rely on the information provided in the datasheet of the IGBT module. The manufacturers usually provide the \( R, C \) values of the Foster model presented in Figure 5.5 (ABB, 2012). The \( R, C \) values have no physical meaning and are obtained by mathematical fitting of the measured thermal impedance \( Z(j \cdot \omega) \) of the module to the function

\[ Z(j \cdot \omega) (\omega) = \sum_{i=1}^{n} R_i (1 - e^{-j/\omega}), \quad (5.9) \]

where the thermal time constant is

\[ \tau_i = R_i \cdot C_i, \quad (5.10) \]
Figure 5.5. Foster model with \( n \) layers.

The Foster model of the IGBT module is valid only when the case temperature is constant. The series connection of the Foster models of the IGBT module, the TIM, and the heat sink leads to erroneous results as has been shown in many papers (ABB, 2012), (Ma et al., 2013). In order to obtain a complete Foster model from the junction to the ambient, the thermal impedance of a specific application should be measured, and then, new \( R, C \) values are obtained by mathematical fitting (ABB, 2012).

5.4 Cooling solutions

Nowadays, many cooling solutions with different benefits and drawbacks are offered for power electronics devices. The cooling solutions are divided based on the cooling medium used in the air and liquid solutions. The advantages of air cooling solutions are their low cost, high reliability, ease of implementation, and maintainability. However, minimization of the thermal resistance of the cooling solution requires the use of bulky heat sink and a considerable increase in the fan speed, which leads to mechanical stresses, and a rise in the converter size and the acoustic noise level. The cooling effectiveness of the air cooling is limited to 60 W/cm\(^2\) as shown in Figure 5.6 (Meysenc et al., 2005).

Liquid cooling is used for high-power applications, where air cooling is not efficient enough to dissipate the generated losses. Liquid cooling is an alternative for the air cooling solution in applications where size reduction is required. Liquid cooling is also applied in low-power applications having liquid circulation. The liquid cooling solutions, again, are divided into single-phase and two-phase ones. There are passive two-phase systems such as heat pipes and thermosyphons (Moreno et al., 2014), (Sacco et al., 2014) and active ones such as looped heat pipes, forced convection in microchannels, and capillary pumped loops (Lachassagne et al., 2013). In order to considerably increase the cooling capability, a high level of integration of the cooling solution with power electronics devices is a new trend in single-phase and two-phase solutions (Zhang et al., 2009), (Vladimirova et al., 2013), (Ivanova et al., 2006), (Barnes and Tuma, 2010). Despite the higher cooling effectiveness of the liquid cooling solutions, their implementation is more laborious and requires auxiliary equipment such as a pump, filters, a water tank, and a heat exchanger.

With a more efficient cooling solution, the power density of the converter can be increased considerably. However, because of the small thermal time constant of an
efficient cooling solution (Figure 5.6), a major change in the junction temperature during a load variation is observed, which leads to a shorter lifetime. Thus, a reliability analysis should be performed in order to select a cooling solution for a particular application. A comprehensive study of the lifetime of the IGBT with different cooling solutions is presented in Section 5.6.

![Figure 5.6. Comparison of cooling solutions in terms of cooling effectiveness and thermal time constant.](image)

### 5.5 Lifetime estimation of the IGBT based on the thermal load profile

For the lifetime estimation in this work, the lifetime model provided by the manufacturer of the IGBT device is applied (Infineon, 2010). The model is given as a series of curves, which allow mapping of the thermal cycle information (Figure 5.7) such as the maximum junction temperature $T_{j,\text{max}}$, the junction temperature change $\Delta T_j$, the cycle period $t_{\text{cyc}}$, and the number of cycles $N_{\text{cyc}}$ to the IGBT failure. In Figure 5.8, the curves are provided for the thermal cycles with $t_{\text{cyc}} = 3$ s. It is shown in Figure 5.8 that the influence of the junction temperature variations $\Delta T_j$ on the number of cycles to the failure $N_{\text{cyc}}$ is much higher than the influence of the maximum junction temperature. The dependence of the $N_{\text{cyc}}$ on the $t_{\text{on}}$ (half of $t_{\text{cyc}}$) variation is shown in Figure 5.9, and also an analytical description is provided by the manufacturer (Infineon, 2010)
Thermal modelling and reliability analysis of IGBT modules

\[ \frac{N_{\text{cyc}}(t_{\text{on}})}{N_{\text{cyc}}(1.5\text{s})} = (\frac{t_{\text{on}}}{1.5\text{s}})^{-0.3}. \]

(5.11)

Figure 5.7. Thermal cycle with \( \Delta T_j = 20 \, ^\circ\text{C}, T_{j,\text{max}} = 100 \, ^\circ\text{C}, t_{\text{cyc}} = 3 \, \text{s}, t_{\text{on}} = t_{\text{off}} = 1.5 \, \text{s}. \)

Figure 5.8. Power cycling capability of the IGBT modules, \( t_{\text{cyc}} \) is 3 s (Infineon, 2010). The component tolerates 30 K temperature variation in 3 s cycles only for one year with the 150 \( ^\circ\text{C} \) maximum temperature (10.5 M \( \approx 10^7 \) 3 s cycles in a year).
Figure 5.9. Power cycling capability of the IGBT modules: typical dependence on $t_{on}$ (Infineon, 2010).

In order to obtain the information about the thermal cycles that the IGBT experiences during operation in a wind turbine converter, the temperature load profile should be obtained by a general algorithm shown in Figure 5.10.

Figure 5.10. Flow chart of the temperature load profile generation of a semiconductor device, where $v_w$ is the wind speed, $P_o$ is the converter power, $P_{loss}$ is the power losses of semiconductor, and $T_j$ is the junction temperature.

In (Ma et al., 2015), a comprehensive procedure to generate the load profile of an IGBT module used in a wind turbine converter is introduced. It is proposed to generate the load profile and perform a lifetime analysis for different time intervals and with different time steps according to the causes of temperature cycles. Thus, the lifetime analysis is divided into long-, medium-, and short-term analyses. A long-term analysis is performed for a one-year period with a time step of several hours, and it allows considering the temperature cycles caused for instance by environmental conditions varying once in a few days or months such as the ambient temperature or the average wind speed. In this analysis, the inertia effects of the wind turbine, generator, cooling system, and the IGBT modules are ignored as their time constants are much less than one hour. A medium-term
Thermal modelling and reliability analysis of IGBT modules

Analysis is performed for a time interval of several hours with a one-second time step, and the focus is on analysing the thermal variations caused by the mechanical behaviour of the wind turbine system such as the pitch and speed control. A medium-term analysis requires consideration of the dynamics of the wind turbine, generator, and the cooling system, which have time constants of seconds to minutes while the dynamics of the IGBT modules having time constant less than a second can be ignored. Finally, a short-term analysis considers a one-second time interval with a time constant of millisecond, where the temperature changes are caused by the fast electrical disturbances such as the alternating current and the switching of the IGBTs. In this analysis, only the dynamics of the IGBT module is considered.

In order to investigate the influence of the cooling solutions with different thermal time constants on the IGBT reliability, a medium-term lifetime analysis is performed in this work.

5.5.1 Lifetime estimation algorithm

In this study, the one-year consumed lifetime of an IGBT of a grid-side converter in a direct-drive wind turbine is estimated considering the thermal cycles less than three hours with a one-second time step. For this purpose, a one-year wind profile measured in the South Karelia region in Finland is converted into a wind speed distribution shown in Figure 5.11. Then, the wind speed in a three-hour time interval (Figure 5.12) is generated using a von Karman spectrum for several values of the average wind speed within the cut-in and cut-out range (Munteanu et al., 2008). According to the IEC Class A, the turbulence intensity (TI) of 18% is used.

Then, the corresponding series of the converter power fluctuations are obtained for three-hour intervals (Figure 5.12). In order to completely reflect the power converter mission profile, the wind turbine power dynamics is processed with a time step of 1 s for three hours. With this time step, the analysis regards both the turbulent wind speed dynamics and the mechanical behaviour of the wind turbine.
Figure 5.11. Wind speed distribution at the height of 88 m over the earth surface from the South Karelia region in Finland.

Figure 5.12. Turbulent wind speed with the 10 m/s average value and the corresponding converter power.
When the converter power is determined, the IGBT power loss is taken from the 3D look-up table where the IGBT power loss for a certain converter power and junction temperature is given. This look-up table is obtained by simulating the converter model with a detailed switching behaviour at different converter power and junction temperature values. Power losses include conduction and switching losses. The usage of the look-up table speeds up the simulations significantly and allows obtaining the temperature load profiles for different average wind speed values (Bryant et al., 2007).

In order to transfer the power loss of the IGBT to the temperature, a thermal model is needed. In this study, a Cauer thermal model is used (Figure 5.13). For the modelling, the four-layer Foster models (Table 5.1) of the IGBT and the diode are mathematically transformed into single-layer Foster models and then to single-layer Cauer models (Table 5.3) by an algorithm presented in (Bagnoli et al., 1998).

Once the temperature load profile of the IGBT in the three-hour interval is obtained, the thermal cycles can be counted by a rainflow algorithm that allows obtaining the information such as a mean value ($T_{j,m}$), amplitude ($\Delta T_j$), and period ($t_{cycle}$) of the thermal cycles. A detailed description of the algorithm is given in (Nieslony, 2009). The rainflow algorithm is suggested as the optimal counting method for the lifetime estimation of power modules as presented in (Mainka et al., 2011), where a comparison of different counting methods is given. The lifetime models of the power devices provided by the
manufacturer are applied to map the $k_{th}$ counted thermal cycle to the number of cycles that the IGBT is expected to have before a failure ($N_{k,cyc}$) (Infineon, 2010). The consumed lifetime by each cycle $CL_k$ is calculated as a reciprocal of $N_{k,cyc}$. Finally, Miner’s rule is applied to calculate the consumed lifetime $CL_{3m/s,3h}$ by the total number of thermal cycles ($K$) within a three-hour interval at a certain average wind speed (Miner, 1945)

$$CL_{3m/s,3h} = \sum_{k=1}^{K} CL_k.$$

(5.12)

When the consumed lifetime for the three-hour intervals at average wind speeds from 3 m/s to 14 m/s ($CL_{x m/s,3h}$) are evaluated, the total one-year consumed lifetime ($CL_{1\text{year}}$) can be calculated by

$$CL_{1\text{year}} = \frac{365 \times 24}{3} (W_{3m/s} \cdot CL_{3m/s,3h} + W_{4m/s} \cdot CL_{4m/s,3h} + \cdots + W_{14m/s} \cdot CL_{14m/s,3h}),$$

(5.13)

where $W_{x m/s}$ is the weighting coefficient taken from the wind speed distribution in Figure 5.11. The estimated lifetime $LT$ is calculated as

$$LT = 100 \frac{CL_{1\text{year}}}{CL_{x m/s,3h}}.$$

(5.14)

5.5.2 Results of the lifetime estimation

For the lifetime estimation, forced air cooling solution and two liquid cooling solutions are considered. The thermal resistance of the cooling solution required for the converter under study is estimated based on a static thermal model (Figure 5.2) assuming that the devices of one phase are placed on a single heat sink ($T_{x1} - T_{x6}$ IGBT modules). The required thermal resistance can be obtained by using forced convection. The heat sink base is $320 \times 280$ mm$^2$, and the fin height is 5 mm. The thermal resistance is given for an air flow rate of 2.5 m/s (Wakefield-vette, 2015a). The thermal capacitance is usually not specified by manufacturers and is calculated from the thermal time constant corresponding to forced convection cooling solutions, as shown in Figure 5.6. The converter power can be increased from 1 p.u. to 1.6 p.u. by switching from a forced convection heat sink to liquid-cooled cold plates. The base of the exposed tube cold plate is $304.8 \times 196.9$ mm$^2$, and the liquid flow rate is 3.8 l/min (Wakefield-Vette, 2015b). The allowed power increase without exceeding the maximum junction temperature is calculated as presented in Section 5.2. The use of liquid-cooled cold plates allows
reducing the thermal resistance by four times compared with forced convection. However, the thermal capacitance is also reduced. A further increase in the converter power is possible by using integrated liquid-cooled cold plates as it is presented in Table 5.3. In this study, the thermal resistance of the integrated liquid-cooled cold plate is decreased by 30% in comparison with a non-integrated solution (Figure 5.6), and the TIM is eliminated. The thermal parameters of the cooling solutions used to generate the temperature load profile of the IGBT are listed in Table 5.3 together with the achieved power increase. The air or liquid temperature is assumed to be 40 °C in the simulations. The generated temperature load profiles of IGBT $S_{11}$ at the 10 m/s wind speed is shown in Figure 5.14. The IGBT $S_{11}$ is found to be the most loaded device in the converter under study.

Table 5.3. Parameters of four cooling solutions.

<table>
<thead>
<tr>
<th>Cooling solution</th>
<th>Thermal resistance, K/W</th>
<th>Thermal capacitance, J/K</th>
<th>Time constant, s</th>
<th>Converter power, p.u.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Forced convection heat sink</td>
<td>0.075</td>
<td>2000</td>
<td>150</td>
<td>1</td>
</tr>
<tr>
<td>Liquid-cooled cold plate</td>
<td>0.018</td>
<td>1000</td>
<td>18</td>
<td>2.4</td>
</tr>
<tr>
<td>Integrated liquid-cooled cold plate</td>
<td>0.013</td>
<td>385</td>
<td>5</td>
<td>3</td>
</tr>
</tbody>
</table>

As shown in Figure 5.14, the amplitude of the junction temperature change is much higher when the cooling solutions with smaller time constants are used. High junction temperature changes are caused by changes in the case temperature, which varies rapidly when cooling solutions with small time constants are used. The lifetime of the IGBT with different cooling solutions is estimated by the algorithm described in Section 5.5.1, and the results are presented in Figure 5.15 and Figure 5.16.
Figure 5.14. Temperature load profile of the IGBT at the 10 m/s wind speed with different cooling solutions.

Figure 5.15. Influence of the cooling solution applied to the consumed lifetime of the IGBT in one year considering medium-term thermal behaviour.
Figure 5.16. Estimated lifetime of the IGBT with different cooling solutions considering only medium-term thermal behaviour.

The difference in terms of lifetime is significant for different cooling solutions, and small inertia of the liquid cooling solution leads to a lower reliability of the IGBT. However, the use of a more efficient cooling solution allows an increase in the converter power. It can be said that the air cooling solution is beneficial in terms of reliability for the converter under study because of the high thermal inertia. However, in high-power applications, the amount of losses usually generated cannot be dissipated by air cooling solutions. Thus, in order to improve the reliability when liquid cooling solutions are applied, the heat sink temperature should be kept constant during load variation. The cooling solutions that maintain constant temperature of the heat sink are preferred for applications with varying loads. For example, two-phase cooling solutions are known to be able to provide a constant heat sink temperature, which is equal to the temperature of the vaporizing liquid used. However, the two-phase cooling solution is expensive and sensitive to mechanical vibrations, which cause evaporator dry-out and system overheating. On the other hand, liquid cooling solutions provide an opportunity of using thermal control to reduce the amplitude of the temperature change during the device operation by maintaining a constant heat sink temperature; if the heat sink temperature of the liquid-cooled cold plate is kept constant and equal to 70 °C, the lifetime is estimated to be 221 years, and thus, the influence of the load variation on the lifetime of the IGBT is negligible.

The lifetime of the IGBT can also be increased by lowering the maximum junction temperature, and thus, oversizing the cooling solution. However, by this measure, the power density of the converter is sacrificed.
5.6 Control of the cooling system

It is possible to maintain a constant heat sink temperature by controlling the fluid flow rate in the liquid cooling solutions, and consequently, changing the convection coefficient and thermal resistance of the cooling solution. The change in thermal resistance, again, changes the thermal time constant. For example, a control strategy is proposed in (Wang et al., 2014a) to control the temperature of the junction during load variation. The strategy is based on the estimation of the losses generated in the power device. In this doctoral dissertation, the use of the gradient heat flux sensor (GHFS) in order to control the heat sink temperature is proposed. The benefit of the use of a heat flux sensor is in eliminating the power loss estimation as they can be directly measured.

The range of thermal resistance variations varies for different cooling solutions, but even with a small range of thermal resistances it is possible to smooth the heat sink temperature variation during the load changes and thereby extend the lifetime. As a case study, the control of fluid flow in a liquid-cooled cold plate is considered. The thermal resistance of the cooling solution varies from 0.018 K/W to 0.18 K/W.

The control strategy proposed here is to calculate the required thermal resistance of the cooling solution \( R_{(h-a)} \) for a particular power loss level of the devices on the heat sink \( P_{\text{loss}} \) in such a way that the temperature difference between the heat sink and the ambient is kept equal to the reference value \( \Delta T_{(h-a),\text{ref}} \)

\[
R_{(h-a)} = \frac{\Delta T_{(h-a),\text{ref}}}{P_{\text{loss}}} \tag{5.15}
\]

and then, to determine the required speed of liquid \( v_l \). The block diagram of the proposed feed-forward control is shown in Figure 5.17. The reference \( \Delta T_{(h-a),\text{ref}} \) is assumed to be 30 °C.

With the proposed control scheme, the lifetime of the IGBT is increased from ten years for the integrated liquid-cooled solution without control to 19.1 years. The generated temperature load profiles of IGBT \( S_{x1} \) with and without a temperature control at a 10 m/s wind speed are shown in Figure 5.18.
Figure 5.17. Feed-forward control scheme. The controller generates a speed reference for the cooling fluid to smooth the variations in the heat sink temperature $T_h$. $T_a$ is the ambient temperature, $\Delta T_{(j-h)}$ is the temperature difference between the junction and the heat sink, and $\Delta T_{(h-a)}$ is the temperature difference between the ambient and the heat sink.

Figure 5.18. Temperature load profile of the IGBT cooled by liquid cold plates at a 10 m/s wind speed with and without the temperature control. The temperature differences in the junction get significantly lower with controlled cooling. However, the average temperature also rises slightly but not harmfully.

The losses of the devices can be directly measured by a heat flux sensor.
5.7 Gradient heat flux sensor

The GHFS used for local heat flux measurements is based on the transverse Seebeck effect, in which the thermo-electromotive force $E$ (thermo-EMF) is proportional to the temperature gradient $\nabla T$ in the direction normal to the applied heat flux vector $q$

$$E = S \cdot \nabla T,$$  \hspace{1cm} (5.16)

where $S$ is the Seebeck tensor.

In the GHFS, the temperature gradients are generated along and across the applied heat flux because of the anisotropy of the thermal conductivity, electric conductivity, and thermo-EMF of the material used. In such sensors, the sensitivity is proportional to the length of the sensor and not to its thickness, because the main direction of the thermo-EMF is along the GHFS. Thus, the thickness of the sensor can be minimized still keeping the sufficient signal strength. A small thickness allows application of the GHFS to measure the local heat flux coming from the IGBT module. The sensor can be placed between the baseplate of the IGBT module and the surface of the heat sink where the TIM is located; Figure 5.19. The thin GHFS has a fast response time in the range of $10^{-8}$–$10^{-9}$ seconds. The details of the GHFS operation and calibration are given in (Murashko et al., 2014), (Mityakov et al., 2012), (Sapozhnikov et al., 2012), (Jussila et al., 2013).

Figure 5.19. Placement of the GHFS under the IGBT chip for the measurement of the local heat flux coming from or to the IGBT module.
The length of the sensor $l_s$ is 20 mm, the width $w_s$ is 5 mm, and the thickness is 0.15 mm. The heat flux per unit area of the GHFS is

$$q_{in} = \frac{e}{w_s \cdot l_s \cdot S_0}, \quad (5.17)$$

where $e$ is the thermo-EMF generated by the sensor, and $S_0$ is the sensitivity of the sensor.

### 5.7.1 Experiment

The GHFS has been tested with the circuit shown in Figure 5.20. Two IGBT modules $T_1$ and $T_2$ operated in a complimentary manner are placed on the heat sink as shown in Figure 5.21, and a fan (the flow rate is 2.6 m$^3$/min) is installed on one end of the aluminium heat sink. The dimensions of the heat sink base are 180 × 125 mm and the height is 120 mm. The dimensions of the fan are 120 × 120 × 38 mm. The IGBT module shown in Figure 2.16 is used, and the characteristics of the module are listed in Table 5.1 and Table A.2 of Appendix A. The signals for the opening and closing of the IGBTs are generated by carrier-based PWM; the unidirectional carrier sine wave frequency is 0.05 Hz and the unidirectional triangular wave frequency is 1 kHz. The GHFS is installed between the baseplate of $T_2$ and the heat sink as presented in Figure 5.22, and the thermocouple is placed close to the GHFS to measure case temperature.

![Figure 5.20. Schematic of the circuit used for the experiment.](image-url)
Figure 5.21. Photograph of the test set-up.

Figure 5.22. Photograph of the GHFS and the thermocouple installed on the baseplate of IGBT module $T_2$.

The heat flux measured by the GHFS and the case temperature measured by the thermocouple in the experiment are shown in Figure 5.23, where $U_i$ is changed from 12
V to 22 V at 1980 s, and the current of the circuit \( I \) is from 15 A to 30 A. The heat flux and the case temperature in the steady-state are shown in Figure 5.24 (\( U_s \) is 12 V). The average heat flux in the steady state is measured to be 1070 \( \text{W/m}^2 \) when \( U_s = 12 \text{ V} \) and \( I = 15 \text{ A} \) and 2020 \( \text{W/m}^2 \) when \( U_s = 22 \text{ V} \) and \( I = 30 \text{ A} \). The area of the baseplate of the IGBT module is 0.0604 \( \times \) 0.1064 \( \text{m}^2 \). Thus, assuming that the heat flux is uniform, the losses measured at \( U_s = 12 \text{ V} \) are 6.9 W and at \( U_s = 22 \text{ V} \) are 13.0 W. The losses estimated based on electrical parameters equal 6.6 W and 12.9 W, respectively. The difference between the measured and estimated results can be explained by the fact that the heat flux is not uniform and has a higher value under the chip area.

Figure 5.23. Heat flux and case temperature measured in the experiment. \( U_s \) is changed from 12 V to 22 V at 1980 s.
Figure 5.24. Heat flux and case temperature measured from 1800 to 1900 s ($U_c$ is 12 V). The waveforms measured in this experiment are also shown in Figure 5.23.

The simple tests shown here demonstrate the opportunity to implement the proposed cooling control for the most efficient cooling cases; one might assume that this kind of a controlled cooling technology may become commercially available in the future.

5.8 Summary

An analysis of the effects of cooling on the estimated lifetime of an IGBT switch was carried out. It was shown that in the case of a very efficient and low time constant cooling, the power electronic switch may suffer from a significant lifetime reduction as the temperature cycling becomes excessive compared with a high-inertia cooling. In the case of a very efficient low time constant cooling, the situation may be improved by arranging control in the cooling. Controlled cooling needs a feedback signal from the heat flux of the switch. A GHFS may be used as such as a feedback element as it reacts fast to the changes in the heat flux.

In this case, the GHFS was used in quite an inefficient way. However, in the future, integrating such a sensor close to the power chip could make the controlled cooling even more efficient in reducing the thermal cycling stresses of an IGBT in cyclic loads such as wind power.
6 Conclusions

This doctoral dissertation presents a study on the design of the power converter main circuit. The main purpose was to develop methods and tools for the design of the converter with a high power density and reliability.

In this study, models to evaluate electromagnetic and thermal performances of the multilevel converter were developed. This work provides guidelines on the design process of the laminated busbar system with a minimized stray inductance of the converter commutation loops. The electromagnetic model of the converter busbar system and the IGBT modules was established to estimate the stray inductance and analyse the electromagnetic coupling between the converter components. A low-inductive busbar system for the three-level ANPC converter was introduced where the symmetrical arrangement of the components is applied to obtain equal stray inductance values in the three phases. The results of the induction estimation were validated by measurements. A circular layout of the busbar system was proposed in order to obtain an equal current sharing between the parallel-connected devices and implemented in the NDT set-up.

In addition to the electromagnetic analysis, the method of a comprehensive thermal analysis of the busbar system was introduced to ensure that the critical temperatures are not exceeded in the designed busbar system. The thermal analysis is based on the 3D LPTM covering all heat transfer mechanisms: conduction, convection, and radiation. The temperature-dependent thermal and electrical parameters of the busbar system materials and of the cooling medium are used to calculate the temperature-dependent Joule losses and thermal resistances of the LPTM. An iterative procedure is implemented for the power losses and the temperature estimation. The main sources of heat in the busbar system are found to be the Joule losses resulting from the contact resistances and the resistances of the IGBT module main terminals. In order to improve the thermal performance of the busbar system, a low contact resistance has to be ensured; this can be achieved, for example, by gold plating of the contact surfaces. The suitability of the thermal model is verified experimentally and by FEM simulations.

With the aim of investigating the effect of the cooling solution on the reliability of the IGBT module in applications with varying loads (e.g. wind turbines), a lifetime estimation algorithm was provided. The lifetime of the IGBT was estimated based on the mission profile of the converter for the South Karelia wind speed distribution. The results of the analysis showed that the thermal inertia of the cooling solution has a significant influence on the reliability of the IGBT module. Thus, in order to select a cooling solution for a specific application that allows achieving a high power capability and the required reliability, the lifetime prediction should be performed at the design stage. The cooling solutions that have a high thermal inertia and maintain a constant heat sink temperature at varying loads are preferred to decrease the thermal cycling stress of the IGBT and thereby improve reliability. In this work, a method was proposed to control the heat sink temperature of the liquid cold plate as a simple way to extend the IGBT lifetime. The method is based on the use of the GHFS to measure the heat flux, and the liquid flow rate.
is adjusted based on the measured heat flux to maintain a constant heat sink temperature. The simulation results show an improvement in the lifetime. Further work should be concentrated on the implementation of the control method in the prototype. The investigation of the option to place the GHFS inside the IGBT module is required as this can further lower the thermal cycling stress of the IGBT.

Furthermore, a complex optimization methodology that incorporates the design methods proposed in this doctoral dissertation has to be developed in order to optimize the converter performance for instance in terms of power density, reliability, and cost.
References


References


References


References


Appendix A: Prototype

Table A.1. Parameters of the ANPC converter.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Line-to-line voltage (rms), V</td>
<td>690</td>
</tr>
<tr>
<td>Phase current (rms), A</td>
<td>120</td>
</tr>
<tr>
<td>DC link voltage, V</td>
<td>1100</td>
</tr>
</tbody>
</table>

Table A.2. Parameters of the converter components.

<table>
<thead>
<tr>
<th>Component</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>IGBT module FZ400R17KE4 (Infineon)</td>
<td>Maximum collector-emitter voltage, V</td>
<td>1700</td>
</tr>
<tr>
<td></td>
<td>Peak collector current, A</td>
<td>800</td>
</tr>
<tr>
<td></td>
<td>Module stray inductance, nH</td>
<td>16</td>
</tr>
<tr>
<td>DC link capacitor B43386-S9418-Q1 Al-Cap (Epcos)</td>
<td>Rated voltage, V</td>
<td>400</td>
</tr>
<tr>
<td></td>
<td>Rated capacitance (100 Hz, 20 °C), mF</td>
<td>4.1</td>
</tr>
<tr>
<td></td>
<td>Stray inductance, nH</td>
<td>20</td>
</tr>
<tr>
<td></td>
<td>Maximum ripple current, A</td>
<td>48</td>
</tr>
</tbody>
</table>
Appendix B: Partial Inductance matrices

Table B.1. Self-partial and mutual-partial inductances of the components constituting commutation loop A of phase a (at 1.2 MHz). The partial inductances of four busbars (P, A1, NT, A2) and two IGBT modules (T_x1, and T_x5) are presented. The diagonal elements (L_{pi,i}) are the self-partial inductances of the components, and the off-diagonal elements (M_{pi,j}) are the mutual-partial inductances.

<table>
<thead>
<tr>
<th>Inductance, nH</th>
<th>P</th>
<th>A1</th>
<th>NT</th>
<th>A2</th>
<th>T_x1</th>
<th>T_x5</th>
</tr>
</thead>
<tbody>
<tr>
<td>P</td>
<td>81.3</td>
<td>3.3</td>
<td>–59.6</td>
<td>–15.9</td>
<td>3.4</td>
<td>2.2</td>
</tr>
<tr>
<td>A1</td>
<td>3.3</td>
<td>43.5</td>
<td>–32.6</td>
<td>–3.2</td>
<td>–2.0</td>
<td>0.0</td>
</tr>
<tr>
<td>NT</td>
<td>–59.6</td>
<td>–32.6</td>
<td>92.4</td>
<td>6.3</td>
<td>–3.8</td>
<td>–4.8</td>
</tr>
<tr>
<td>A2</td>
<td>–15.9</td>
<td>–3.2</td>
<td>6.3</td>
<td>7.2</td>
<td>0.0</td>
<td>0.0</td>
</tr>
<tr>
<td>T_x1</td>
<td>3.4</td>
<td>–2.0</td>
<td>–3.8</td>
<td>0.0</td>
<td>13.9</td>
<td>0.6</td>
</tr>
<tr>
<td>T_x5</td>
<td>2.2</td>
<td>0.0</td>
<td>–4.8</td>
<td>0.0</td>
<td>0.6</td>
<td>13.4</td>
</tr>
</tbody>
</table>
Appendix C: Thermal model parameters

Table C.1. Parameters for the calculation of the constriction resistance (Slade, 2014).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Hardness of copper $H$, N/mm²</td>
<td>4 – 9</td>
</tr>
<tr>
<td>Hardness of aluminium $H$, N/mm²</td>
<td>1.8 – 4</td>
</tr>
<tr>
<td>Resistivity of copper $\rho$, Ohm·m</td>
<td>1.68·10⁻⁸</td>
</tr>
<tr>
<td>Resistivity of aluminium $\rho$, Ohm·m</td>
<td>2.82·10⁻⁸</td>
</tr>
<tr>
<td>Radius of the bolt $r_b$, m</td>
<td>3·10⁻³</td>
</tr>
<tr>
<td>Applied torque $T$, N·m</td>
<td>4 (max 5)</td>
</tr>
<tr>
<td>Nut factor $k_n$</td>
<td>0.3</td>
</tr>
</tbody>
</table>

Table C.2. Thermophysical properties of air at atmospheric pressure (Incropera, 2006).

<table>
<thead>
<tr>
<th>Temperature, K</th>
<th>Kinematic viscosity $\nu$·10⁶, m²/s</th>
<th>Thermal diffusivity $\alpha$·10⁶, m²/s</th>
<th>Thermal conductivity $k$·10³, W/(K·m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100</td>
<td>2.00</td>
<td>2.54</td>
<td>9.34</td>
</tr>
<tr>
<td>150</td>
<td>4.426</td>
<td>5.84</td>
<td>13.8</td>
</tr>
<tr>
<td>200</td>
<td>7.590</td>
<td>10.3</td>
<td>18.1</td>
</tr>
<tr>
<td>250</td>
<td>11.44</td>
<td>15.9</td>
<td>22.3</td>
</tr>
<tr>
<td>300</td>
<td>15.89</td>
<td>22.5</td>
<td>26.3</td>
</tr>
<tr>
<td>350</td>
<td>20.92</td>
<td>29.9</td>
<td>30.0</td>
</tr>
<tr>
<td>400</td>
<td>26.41</td>
<td>38.3</td>
<td>33.8</td>
</tr>
<tr>
<td>450</td>
<td>32.39</td>
<td>47.2</td>
<td>37.3</td>
</tr>
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