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Power Loss Modeling of Isolated AC/DC Converter

Royal Institute of Technology

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Electrical Power Engineering

Power Loss Modeling of Isolated AC/DC Converter

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Naveed Ahmad Khan

Stockholm
Several research activities at KTH are carried out related to Isolated AC/DC converters in order to improve the design and efficiency. Concerning the improvement in the mentioned constraints, losses of the elements in the prototype converter are modeled in this thesis work. The obtained loss model is capable of calculating the losses under different circumstances. The individual contribution of losses for each element at different conditions can be obtained, which is further useful in improving the design and therefore, efficiency. The losses in different elements of the converter, including power semiconductor devices, RC-snubbers, transformer and filter inductor at different operating points can be computed by using the obtained model. The loss model is then validated by comparing the analytical results with the measurements.

The results based on developed loss model show consistency with the measured losses. The comparison at different conditions shows that, the difference between measured and analytical results ranges between 10% to 20%. The difference is due to those losses which are disregarded because of their negligible contribution. On the other hand, it is also observed that if the neglected losses are counted, the difference reduces up to 10%.
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<td>VSC</td>
<td>Voltage Source Converter</td>
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<td>PWM</td>
<td>Pulse Width Modulation</td>
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<td>$U_d$</td>
<td>Input DC link Voltage</td>
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<td>IGBT</td>
<td>Insulated Gate Bipolar Transistor</td>
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<tr>
<td>$l_e$</td>
<td>Effective magnetic path length</td>
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<td>$A_{core}$</td>
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$i_F$  Forward current of diode

$P_{\text{cond}_{\text{igbt}}}$  Conduction losses in IGBT

$P_{\text{cond}_{\text{diode}}}$  Conduction losses in diode

$E_{\text{off}}$  Turn-off energy loss

$f$  Fundamental frequency

$f_{\text{tr}}$  Transformer operating frequency

$W_{CS}$  Energy stored in capacitor

$\Delta t_{\text{accl}}$  Commutation time of cycloconverter valve

$P_\nu$  Loss density

$K, \alpha, \beta$  Coefficients of Steinmetz equations

$B_m$  Peak magnetic flux density

$R_{AC}$  AC resistance

$R_{DC}$  DC resistance

$\delta_w$  Skin depth

$d$  Diameter of conductor

$\rho$  Resistivity of material

$\mu_0$  Permeability

$F_{RS}$  Skin effect factor

$F_{RP}$  Proximity effect factor

$M$  Modulation Index

$\bar{U}_{\text{carrier}}$  Peak amplitude of carrier voltage

$\bar{U}_{\text{cyc},1}$  Fundamental component of cycloconverter voltage

$L_{f/ii}$  Filter Inductance
$C_{fit}$  Filter capacitance

$T_j$  Junction temperature of semiconductor devices

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1. Introduction

The thesis deals with the power loss modeling of “Two Stage Isolated AC/DC Converter” which is suitable for sustainable energy sources. This chapter provides a base for all material presented in the next chapters, which focuses on the loss modeling of different elements in the prototype converter.

1.1 Background

The thesis is related to the power loss modeling of the two stage isolated AC/DC converter. The power loss modeling of the power electronic converters is of vital importance because of its relation with efficiency, reliability, cost and size.

The prototype converter comprises a Voltage Source Converter (VSC) and a cycloconverter coupled by a medium frequency transformer, which are studied in this thesis. The prototype can be operated by applying the soft-switched commutation across all the valves at all the operating points leading to lower switching losses of the power semiconductor devices. The valves of the VSC in the prototype are equipped with lossless snubbers that reduce the switching losses across the IGBTs and the diodes. The cycloconverter in the prototype is equipped with RC-snubbers to reduce the stress on the valves during the commutation and other transients. Furthermore the medium frequency transformer has also lower losses because of its compact size. The concepts applied to the prototype are useful in loss reduction and further can result in higher efficiency and ultimately power density which are the main goals in the design of power electronic converters.

Despite the fact that various loss minimization techniques are employed, the prototype suffers lower efficiency at lower output power. There can be a number of reasons. In order to resolve the issue, power loss modeling of the prototype is required. After obtaining the proper loss model, the losses across the individual elements in the prototype can be obtained, which further helps in improving the design and efficiency.
1.2 Motivations and Objectives

Several research activities at KTH at the Department of Electrical Engineering have been carried out concerning isolated AC/DC converters with medium frequency transformer [1], [2]. The concept is attractive due to the fact that it allows soft switching in all the operating points that results in high efficiency and ultimately high power density. The prototype has been built and tested successfully by Dr. Staffan Norrgra in his PhD Dissertation [3], the idea of which is shown in figure 1-1.

![Prototype Converter](image)

The main objective of the thesis work is to develop a loss model for the prototype converter that is able to describe the losses in all the elements that the converter includes.

There are two major reasons that motivate this work,

- The prototype needs further optimization for lower output powers and it is needed to be seen whether there is a potential for further loss reduction, see figure 1-2 for previous work
- From previous work it was also observed that the measured and the calculated losses are not consistent indicating that the obtained loss model does not take into account all the causes of losses, Figure 1-3. Therefore, it is needed to improve the loss model by investigating those losses that are not identified.
The work breakdown structure, in order to obtain the loss model for the prototype converter is,

- The theoretical concepts of the converter are studied in order to understand the operation of each element within it.
- The possible losses that can exist in the prototype converter are investigated.
- The methods to compute those losses are gathered.
- The appropriate methods are then used to get the proper loss model.
- The analytical results are compared with the measured ones, in order to validate the obtained loss model.
1.3 Outline of Thesis
The Thesis has been divided into six chapters.

Chapter 2
The introduction and operation of the converter considering the losses are explained in this chapter.

Chapter 3
The methods to calculate the losses are explained in this chapter. Apart from this the mathematical description and the formulation of the losses is also presented.

Chapter 4
In this chapter, the MATLAB model developed for loss computation is presented and explained.

Chapter 5
This chapter presents the simulation results of the losses in each element of the converter. The effects of variations of different parameters on the losses are presented. Besides this, the comparison of measured and analytical results is also shown.

Chapter 6
Conclusion and future work are presented in this chapter.
2. Isolated AC/DC Converter

This chapter gives an overview and analysis of a prototype converter studied for power loss modeling. The operation of the converter is explained with emphasis on losses. Apart from this, implementation of the different components is also described.

2.1 Introduction to Converter

The prototype converter resides in the class of converters that perform bidirectional AC/DC operation with isolation by a medium frequency transformer. The fundamental structure is presented in figure 2-1. The transformer in the middle is magnetized by a medium frequency AC voltage, supplied by a Voltage Source Converter (VSC), which is equipped with snubber capacitors. On the other winding, the voltage is converted to a Pulse Width Modulation (PWM) voltage by a cycloconverter which is then applied to an inductive filter in order to achieve a desired AC voltage. This configuration is compatible with a commutation sequence that permits soft-switching conditions for all the power semiconductor valves at all the operating points.

Figure 2-1 Mutually Commutated Converter System

2.2 Converter Topology

The topology of the converter is shown in figure 2-2. The VSC equipped with the snubber capacitors in parallel to each valve is connected to one of the windings of the medium frequency single phase transformer. The other winding is connected to the 2-phase to 3-phase cycloconverter which is then connected to a passive line filter [1].
2.2.1 Voltage Source Converter (VSC)

In general, a VSC converts a DC voltage to an AC voltage of a desired frequency and amplitude. Usually PWM strategies, due to their simplicity, are used to control the switching of the IGBTs. Normally, the valves in the VSC are designed in order to block the voltage in one direction by using gate command and also to conduct a current in both directions.

The VSC in the studied converter is a bridge converter having two phase legs. Each phase leg has two valves. Each valve consists of an IGBT with an anti-parallel diode. Snubber capacitors are also used in parallel to each valve to reduce the switching stress to a safe level. The output is connected to the midpoint of each phase leg. A DC link capacitor is used at the input in order to by-pass the AC components during a power flow from AC to DC side.

In the studied converter, different control strategy has been adopted for the VSC to produce an “equal width square” wave at the output by a proper switching of the IGBTs. As a result, a square wave of constant frequency having levels $U_d$, 0 and $-U_d$ is produced. Where, $U_d$ is the amplitude of the DC-link voltage.

The valves in the VSC are implemented with “BSM150GB120DLC” IGBT module with the blocking voltage and rated current of 1200 V and 150 A respectively. The snubber capacitors are of the polypropylene type having a capacitance of 220nF per valve. The IGBT modules are connected to a DC link capacitor that has a capacitance of 2.9 mF.
2.2.2 Cycloconverter

In general, a cycloconverter converts an AC voltage directly to another AC voltage of desired frequency and amplitude without any intermediate DC stage. In this type of converter, bidirectional power flow is possible with an ability to operate with a load of variable power factors. Generally, the valves in the cycloconverter are designed to conduct the currents in one direction and to block voltage in both directions.

The cycloconverter in the studied converter is a 2-phase to 3-phase converter having three phase legs. Each phase leg has two valves. The bidirectional valves are made by connecting the IGBTs in the common-emitter fashion as shown in figure 2-2. RC-Snubbers are used in parallel with each valve in order to prevent the overvoltage during the commutation process.

The switching of the valves in each phase leg follows the PWM strategy, which results in a rectangular waveform having a variable width.

The valves in the cycloconverter are implemented with “BSM100GT120DN2” IGBT module having the blocking voltage and the rated current of 1200 V and 150 Amps respectively.

2.2.3 Medium Frequency Transformer

A medium frequency transformer is used in the studied converter to couple the VSC with the cycloconverter.

The transformer is of a toroidal type with a ferrite core from AVX, type B2. Four pieces of the toroidal cores are stacked together to make a core. Each piece has an outer diameter equal to 152mm, inner diameter 68mm and a width of 19mm. One of the windings is split into two in order to reduce the leakage inductance. The key data of the transformer is listed in table 2-1.

<table>
<thead>
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<th>Table 2-1 Medium Frequency Transformer Data</th>
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<tbody>
<tr>
<td>Transformer operating frequency</td>
</tr>
<tr>
<td>Transformer turns ratio</td>
</tr>
<tr>
<td>Number of turns winding 1</td>
</tr>
<tr>
<td>Number of turns winding 2</td>
</tr>
<tr>
<td>Core Cross Section</td>
</tr>
<tr>
<td>Core peak flux density</td>
</tr>
<tr>
<td>Magnetizing Inductance</td>
</tr>
<tr>
<td>Leakage Inductance</td>
</tr>
<tr>
<td>Total Weight</td>
</tr>
</tbody>
</table>
2.2.4 Output Filter

The filter in the studied converter has been implemented with a simple LC filter. It has an ability to smooth the load current by blocking the harmonics produced by the PWM voltage. Proper values of capacitance and inductance have been chosen to get the desired results. The parameters that specify the filter are given in table 2-2.

<table>
<thead>
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<td>Inductance (typical)</td>
</tr>
<tr>
<td>Capacitance</td>
</tr>
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</table>

The inductor in the filter has an iron core to improve the inductance, the key parameters of which are given in table 2-3. The core material for an inductor core is assumed in this case because the loss data for the real core is not available. However, the assumed core has the same lamination thickness as the one used in the actual core.

<table>
<thead>
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<th>Table 2-3 Filter Inductor Data</th>
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<tr>
<td>Core material</td>
</tr>
<tr>
<td>Core shape</td>
</tr>
<tr>
<td>Effective magnetic path length</td>
</tr>
<tr>
<td>Effective cross sectional area</td>
</tr>
<tr>
<td>Number of turns of winding</td>
</tr>
<tr>
<td>Total Weight</td>
</tr>
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</table>

2.3 Principle of Operation

To analyze the converter, a convenient method used by [1] is considered in which the coupling functions that relate the voltages and currents are used. For ease of analysis, all the voltages and currents are represented with respect to the transformer voltage [1].

The coupling function for a VSC to couple a DC-link voltage with a transformer winding is represented by \( k_d [3] \). For a full bridge VSC, \( k_d = +1 \), when the valves in one of the diagonals are conducting and \( k_d = -1 \), when the valves in the other diagonal are in conduction state. In this regard, the transformer voltage at any instant is given by equation-2.1.
\[ u_{tr} = k_d \cdot U_d \]  

(2.1)

where, \( U_d \) is a DC-link voltage.

Similarly, the coupling function defined for each phase leg of the cycloconverter is represented by \( k_{aci} \) [1]. Where, \( k_{aci} = \pm \frac{1}{2} \), when the phase leg of the cycloconverter is connected to the upper and lower terminal of the transformer respectively. The cycloconverter voltage as a function of coupling function can be expressed as,

\[ u_{aci,i} = k_{aci} \cdot u_{tr} \cdot N_{tr} \]  

(2.2)

Similarly, the transformer current can be related to the cycloconverter current by using the coupling function as,

\[ i_{tr} = N_{tr} \cdot \sum_i k_{aci} \cdot i_{aci} \]  

(2.3)

The turns ratio of the transformer is assumed to be unity. The coupling function \( k_{aci} \) is one half because the output voltages are referred to the midpoint of the transformer windings connected to the cycloconverter [1].

From the aspect of losses when the coupling functions are non-zero, the power semiconductor devices are in on-state and present the on-state voltage during current flow resulting in conduction losses. These losses have a major contribution to the total losses and therefore need to be computed.

### 2.4 Commutation Sequence

The commutation process is also analyzed in a simple way by making few assumptions [3].

- The voltage applied to a DC link capacitor is assumed to be essentially constant.
- The inductor filter on the AC side is assumed to be large enough as it has the ability to maintain constant current in each modulation interval and can be modeled by a current source.
- The transformer is modeled by a leakage inductance \( L_A \), i.e. magnetizing current is neglected.

Both converters commutate alternately. VSC undergoes snubbered or zero voltage commutation [1] that reverses the voltage across the transformer and enables a source commutation (natural commutation) of the cycloconverter which in turn reverses the direction of the transformer current.
2.5 Commutation process with an aspect of losses

2.5.1 VSC commutation

The VSC commutation starts by turning-off the IGBTs, thereby diverting the current to the snubber capacitors to recharge them. The voltage across the transformer starts decreasing whereas the current remains constant due to the passive line inductance. When the potential of the phase legs has fully moved to the opposite DC rail, the diodes in the opposite valves start to handle the current. At this instant the power flow is from AC side to the DC side as explained briefly in [1]. The IGBT anti-parallel to the conducting diode can be turned on at a zero-voltage and zero-current conditions. The turn-off is made at a low voltage derivative [1] due to the snubber capacitors.

The VSC commutation time can be expressed as,

\[ \Delta t_{VSC} = 2. C_s \cdot \frac{v_d}{i_{tr}} \]  \hspace{1cm} (2.4)

The VSC commutation process is also illustrated together with different stages in figure 2-3.

![Figure 2-3 Commutation process in VSC [1]](image)

Loss perspective

IGBTs

The turn-off of the IGBTs in the VSC occurs at high current, therefore the turn-off losses in the IGBTs need to be considered, whereas the turn-on process takes place at zero voltage and zero current condition therefore the turn-on losses do not exist.
The VSC can be operated with or without snubber capacitors. In both cases the IGBTs turn-on at zero current and zero voltage conditions. On the other hand, the turn-off process takes place at higher current for both operations therefore the turn-off losses need to be computed in both cases.

The turn-off process of an IGBT without snubbers is shown in figure 2-4.

![Figure 2-4 Turn-off Process of IGBT](image)

Two distinct parts of the turn-off current have been observed. The rapid drop in the current during $t_{f1}$ interval corresponds to the turn-off of the MOSFET’s part of the IGBT whereas the tail in the current during the time interval $t_{f2}$ is due to a stored charge in the $n^{-}$ drift region. The tail in the current has a significant contribution to the power dissipation as the collector-emitter voltage is in its off-state value. The duration of the tail current also increases with increased temperature. The overlap between the voltage and current for the interval $t_{off}$ indicates the area of the power loss during the turn-off process.

In case of a turn-off with snubbers (soft-switching), the voltage rises linearly with slow rate which depends upon the current that flows through the capacitor to recharge it, figure 2-5. This condition reduces the overlap between the voltage and current and hence the switching losses. Though the overlap between the current and voltage is reduced, still the losses need to be considered due to the tail current that overlaps with the voltage during the voltage rise.
The stress on a switch as well as electromagnetic interference can be reduced by using the soft switching strategy. The switching frequency and the power density can be increased as well.

**Diodes**

Usually the turn-on losses of the diode are negligible as it turns-on quickly [4]. The diode however, exhibits reverse recovery losses during turn-off. But in this case they are negligible because the current flowing through it goes to zero naturally and at this moment IGBT is turned on at zero voltage and zero current condition. As a result it faces only the on-state voltage of the IGBT during a turn-off process (reverse recovery current flow) that results in a negligible amount of reverse recovery losses.

**VSC-Snubbers**

In the VSC, lossless snubbers are used as IGBTs turn-on at zero voltage and zero current. During the VSC commutation, the capacitors in one of the diagonals of the VSC transfer their energy completely to the capacitors in the other diagonal after which the IGBTs are turned-on. Therefore the losses due to these snubbers do not exist.

VSC commutation causes voltage reversal at the transformer terminal leading to a source commutation (natural commutation) in the cycloconverter.

**Cycloconverter Snubbers**

The VSC commutation establishes a condition of voltage reversal across the RC snubbers in the cycloconverter. The voltage across the charged snubbers changes from $\pm U_d$ to $\mp U_d$ respectively that corresponds to the discharge and then recharge of the snubbers. As resistive snubbers are mounted, the stored energy in the capacitors dissipates across the resistors.
The VSC commutation ends up with the turn-off losses of the IGBTs in VSC and RC-snubber losses in the cycloconverter.

### 2.5.2 Cycloconverter Commutation

In cycloconverter commutation, one phase leg undergoes commutation at a time. It starts by turning-on the non-conducting valve in the direction of the current through the phase terminal with the following condition to be fulfilled [1]

$$k_{ac}, u_{tr}, i_{acl} < 0$$  \hspace{1cm} (2.5)

The voltage appears across the leakage inductance, and the incoming valve takes over the current [1]. The current in the initially conducting valve goes to zero, and the switch in the valve can be gated off. The current derivative depends upon the transformer’s leakage inductance and is relatively low. The commutation time for each phase leg can be estimated by

$$\Delta t_{Ai} = L_{L}, \frac{i_{acl}}{u_{d}}$$  \hspace{1cm} (2.6)

At the end of commutation for each phase leg, the following condition becomes valid

$$k_{acl}, u_{tr}, i_{acl} > 0$$  \hspace{1cm} (2.7)

Figure 2.6 highlights the cycloconverter’s commutation process [1].

![Figure 2-6 Commutation of one phase leg [1]](image)

When all the phase legs have been commutated the following condition becomes valid [1]

$$u_{tr}, i_{tr} = N_{tr}, \sum_{i} k_{acl}, u_{tr}, i_{acl} = N_{tr}, \sum_{i} \frac{1}{2} |u_{tr}| |i_{tr}|$$  \hspace{1cm} (2.8)

After the cycloconverter’s commutation, the VSC commutation occurs again and the process repeats.
Loss perspective

*IGBTs*

In the studied cycloconverter, during the turn-on process the rate of rise of the current is slow due to the leakage inductance of the transformer that reduces the overlap between the current and the voltage. Consequently, the switching losses are also reduced. The transition of voltage and current is shown in figure 2-7. The power dissipation encloses a very small area of negligible amount during each transition therefore the turn-on losses in this case can be neglected.

![Figure 2-7 Turn-on behavior of IGBT in cycloconverter](image)

The IGBTs turn-off naturally and are gated off when the current through them goes to zero; therefore turn-off losses in the cycloconverter do not exist.

*Diodes*

Normally reverse recovery losses occur in the diodes during the turn-off period. These losses exist in this case as well because the diode voltage acquires substantial negative values during turn-off. These losses are due to the negative current that flows as a result of recombination of the carriers to remove the stored charge in the drift region. This stored charge is called recovered charge $Q_{rr}$ which is represented by a shaded area in figure 2-8. During the negative current flow it holds the on-state voltage whereas, when the carriers are swept out completely, the negative voltage begins to rise and reaches substantial value. Figure 2-8 highlights the turn-off process of the diode.
The diode turn-off losses depend upon the charge that is removed with the reverse recovery current. A fast decrease in the forward current gives a short time for the recombination of charges therefore a large amount of charge must be removed in a short time. Whereas, a slow decrease in the current gives the carriers a long time to recombine. This results in low recovery current and hence lower losses.

**RC-Snubbers**

The turn-on/off of IGBTs in the cycloconverter also affects the snubbers in the parallel. The turn-on of the valves creates a short circuit condition for the snubber capacitors in the parallel that result in the complete discharge of the stored energy. Whereas when the valves turn-off, the parallel snubbers recharge completely. This discharge and recharge results in power dissipation across the resistances and incoming valves.

### 2.5.3 Resonant commutation

The control strategy in the converter is developed in such a way that the installed capacitors in the VSC should not dissipate their energy across the valves during VSC commutation. At low currents, the capacitors can take long time to recharge or discharge, and the IGBTs can be turned-on before the commutation process ends. Consequently, hard switching can occur and capacitors can dissipate their stored energy across the IGBTs. In order to avoid this, a resonant commutation is introduced at low currents. According to this, all the valves in the cycloconverter are turned-on at low currents during the VSC commutation, thereby short-circuiting the secondary side of transformer winding. This results in a high instantaneous current flow that can recharge the capacitors in short time. In this way hard turn-off can be avoided and IGBTs can be turned-on at zero voltage and zero current conditions. The resonant commutation process is explained explicitly in [3].

![Diagram](image-url)
2.5.4 Voltage and Currents

The transformer voltage and current along with the cycloconverter output voltages for each phase are shown in figure-2.9 [1]. The instant of commutation for VSC and cycloconverter are also shown in this figure.

The commutation times have been exaggerated in the figure for clarity purpose. In the real implementation however, the time is of small fractions [1].

2.5.5 Modulation

In the studied converter, proper operation of the transformer has been made possible by avoiding low frequency components or the DC component in the voltage applied to the transformer. This has been achieved by making the VSC commutation at constant intervals i.e. a square wave voltage is applied to the transformer. [1]

In the cycloconverter, a simple and a straight forward PWM strategy has been used. Two saw-tooth carriers based on the load current are compared with the reference signal in order to get the switching instants. The desired PWM signals have been achieved by making the commutation of the cycloconverter’s phase legs at appropriate instants i.e. between the two successive VSC commutations. In this way, the width of the PWM pulses can be chosen freely [1].
2.5.6 Medium frequency transformer with loss aspects

The medium frequency transformer couples the VSC and the cycloconverter. The windings are excited by an equal width AC Square wave that causes magnetization and demagnetization of the core and therefore results in the hysteresis and eddy current losses. Due to current flow through the windings, copper losses also occur.

2.5.7 LC filter with loss aspects

An LC filter has been used at the output of the cycloconverter to transform PWM voltage to a sinusoidal voltage by filtering the generated harmonics. The output of the converter will be sinusoidal under linear load or steady state condition and can be distorted in the case of non-linear load because of the slow system response.

Since the inductor in the filter also includes a core, core losses along with the winding losses also exist.

2.6 Summary

The operation of the converter summarizes the following with regard to the losses.

The following losses in the prototype converter need to be computed:

- Conduction losses of power semiconductors in the VSC and cycloconverter
- Turn-off losses of the IGBTs in the VSC and the diodes in the cycloconverter
- RC-Snubber losses in the cycloconverter
- Core losses and winding losses in the transformer and the filter reactor
3. Evaluation of Losses

This chapter presents the models to find out the contribution of losses of each component mounted in the prototype converter. The components in which the losses exist are treated separately. Furthermore, the methodology to compute the possible losses from the point of view of the prototype converter is presented.

3.1 Introduction

In the prototype converter, power semiconductor devices as well as inductive elements are mounted. These elements are not ideal therefore whenever current flows through them, it leads to a voltage drop across them, which in turn causes power dissipation.

From the discussion in chapter 2 it is concluded that the following losses should be computed in the prototype converter,

- Power semiconductor losses
  - Conduction losses in the VSC and the cycloconverter
  - Switching losses: Turn-off losses in the VSC (in case of hard and soft switching) and diodes (reverse recovery) in the cycloconverter
- RC-Snubber losses in the cycloconverter
- Transformer losses (winding and core)
- Inductor losses (winding and core)

3.2 Power semiconductor Losses

The power semiconductor losses can be estimated by calculating the energy loss per pulse and then adding them. A couple of methods have been developed to predict the losses in the power semiconductor devices and are well known. Loss calculation is possible either by a complete numerical simulation of a circuit by using special simulation tools with the integrated loss computation tools [4]. The other way is to determine the electrical behavior of a circuit analytically, i.e. currents and voltages to calculate the power losses [4]. In the manufacturers datasheet useful information about the loss calculation is also available by using which; the losses in the semiconductor devices can be computed efficiently. In this work, a simple method
Power Loss Modeling of Isolated AC/DC converter

using the data sheet parameters is considered to compute the losses in the power semiconductors.

There are two kinds of losses in the power semiconductors, namely static losses also called conduction losses, and the transient losses known as switching losses.

### 3.2.1 Conduction Losses

In the conduction state both the IGBTs and the diodes have a voltage drop across them. The product of the voltage drop and the conducting current determines the conduction losses [6].

The power semiconductor device can be modeled as a voltage source in series with a linear resistor [7], [8]. The simplified model is valid for both the IGBTs and the diodes. The on-state voltage of the IGBTs and diodes can be expressed as,

\[ V_{CE}(i_{CE}) = V_{CEO} + r_i i_{CE} \]  \hspace{1cm} (3.1)

Similarly for the diodes

\[ V_F(i_F) = V_{FO} + r_d i_F \]  \hspace{1cm} (3.2)

Whereas \( i_{CE} \) and \( i_F \) are the currents flowing through the IGBTs and diodes respectively. \( V_{CEO} \) and \( V_{FO} \) are the threshold voltages across the IGBTs and diodes while \( r_i \) and \( r_d \) are the slope resistances.

The on-state voltage can be obtained from the datasheet which is available as a function of a forward current. The average conduction losses can then be obtained by using a direct method of the product of instantaneous current and the corresponding on-state voltage by averaging over one fundamental period.

\[ P_{cond_{igbt}} = \frac{1}{T} \int_{t=0}^{T} V_{CE}(i_{CE}).i_{CE} \, dt \]  \hspace{1cm} (3.3)

\[ P_{cond_{diode}} = \frac{1}{T} \int_{t=0}^{T} V_F(i_F).i_F \, dt \]  \hspace{1cm} (3.4)

Where, \( T \) is the fundamental period.

### 3.2.2 Switching Losses

Switching losses occur when the semiconductor devices undergo into a transition state due to finite time taken by a charge to respond to the applied voltages [9].
From the consideration in chapter 2, the switching losses in the prototype converter that should be considered are the turn-off losses of the IGBTs in the VSC and the reverse recovery losses of the diodes in the cycloconverter.

**IGBTs**

The turn-off losses in the IGBT are measured from the instant when the collector-emitter voltage starts rising, up to the instant when the collector current drops to zero. The dissipated energy is given as follows:

\[ p_{sw} = v_{CE}(t)i_{IGBT}(t) \]  \hspace{1cm} (3.5)

\[ E_{off} = \int_{t_1}^{t_2} p_{sw} \, dt \]  \hspace{1cm} (3.6)

Where, \( t_1 \) is the instant at which the collector-emitter voltage begins to rise and \( t_2 \) is the instant at which the collector current drops to zero.

The device turns-off several times during a whole fundamental cycle. As a result the turn-off switching losses are the sum of energy loss per pulse over a one fundamental period which is expressed as,

\[ P_{switch_{igbt}} = f \sum_{x=1}^{f_{tr}/f} E_{off} \]  \hspace{1cm} (3.7)

Where \( f \) is the fundamental frequency, \( f_{tr} \) is the switching frequency, which is the transformer’s operating frequency in this case.

In the data sheet the turn-off energy values as a function of device current are available. These values can be used to calculate the turn-off losses. The energy values for every instant of a turn-off for a specified current can be obtained from the datasheet and the total losses can be computed by using equation (3.7).

The VSC in the prototype is also operated with the “soft turn-off” of the IGBTs. In the data sheet, the energy values are given only for “hard turn-off” therefore the energy values need to be computed for the soft turn-off operation. The required energy values are already obtained by Norrnga in his Doctoral Thesis [3] for the prototype converter which can be used in this case. By using those energy values the turn-off losses can be calculated by using equation (3.7).

**Diode reverse recovery losses**

In order to compute the reverse recovery losses the waveforms of the voltage and current are required. However, these losses can also be calculated by using the information available in the data sheet.
The energy loss can be expressed by considering figure 3-1 as,

$$E_{rr} = \int_{t_1}^{t_3} i_{rr} u_{diode} dt$$  \hspace{1cm} (3.8)$$

Where, $u_{diode}$ is the voltage across the diode and $i_{rr}$ is the current during the turn-off process.

![Figure 3-1 Diode reverse recovery losses [5]](image)

From the instant $t_1$ to $t_2$ the loss contribution is negligible due to small voltage drop, whereas from $t_2$ to $t_3$ the diode acquires a high voltage and has a major contribution in the recovery losses. The energy loss can therefore be approximated as

$$E_{rr} \approx U_{diode} \int_{t_2}^{t_3} i_{rr} dt \approx 0.5 U_{diode} Q_{rr}$$  \hspace{1cm} (3.9)$$

The recovery charge $Q_{rr}$ represented by the shaded area in figure 3-1 is available in the data sheet and can be used in equation 3.9 to calculate the energy loss.

The obtained energy values from (3.9) are for conventional test voltages and currents and should be adjusted according to the specified application by [7], [8],

$$E_{rec} = E_{rr} \frac{V_p}{V_{test}} \frac{i_p}{I_{test}}$$  \hspace{1cm} (3.10)$$

Here, $E_{rr}$ is the energy loss for the specific test voltage and current. $V_p$ and $i_p$ are the actual commutation voltage and current, $V_{test}$ and $I_{test}$ are the test commutation voltage and current respectively.

As the recovery losses also depend upon the rate at which the current changes therefore the linear dependence of the current derivative is also need to be taken into account. By doing so equation 3.10 becomes [10],

$$E_{rec} = E_{rr} \frac{V_p}{V_{test}} \frac{i_p}{I_{test}} \frac{dV_p/dt}{dI_{test}/dt}$$  \hspace{1cm} (3.11)$$
After calculating the energy, the total power loss for each diode can be calculated by using equation 3.7.

**Total semiconductor losses in VSC and Cycloconverter**

The VSC is a single phase converter and it has two phase legs, whereas the cycloconverter is a three phase leg converter.

The VSC has a total of four valves each of them includes an IGBT and a diode therefore the total losses in it will be

\[ P_{total_{SC}} = 4 \left( P_{\text{switch}_{\text{IGBT}}} + P_{\text{cond}_{\text{diode}}} + P_{\text{cond}_{\text{IGBT}}} \right) \]  \hspace{1cm} (3.12)

Similarly, there are six valves in the cycloconverter each of them includes two IGBTs and two diodes therefore the total losses in it will be

\[ P_{total_{Cyc}} = 12 \left( P_{\text{cond}_{\text{diode}}} + P_{\text{cond}_{\text{IGBT}}} + P_{\text{switch}_{\text{diode}}} \right) \]  \hspace{1cm} (3.13)

Hence the total semiconductor losses in the prototype converter will be

\[ P_{total} = P_{total_{Cyc}} + P_{total_{SC}} \]  \hspace{1cm} (3.14)

**3.2.3 Thermal Characteristics of Power Semiconductors**

The dissipated power in the power semiconductor devices turns into heat which in turn increases the junction temperature inside the chip [11]. This results in the degradation of the characteristics and reduces the lifetime. The junction temperature can be reduced by permitting the heat to escape outside. The ability of the device to dissipate the heat is measured by a thermal resistance. If the heat flow in the thermal network is considered equivalent to a current in the electrical network, the heat dissipation channel can be represented by a model shown in figure 3-2. The entire thermal resistance from the chip junction to the ambient air can be designated as \( R_{\theta JA} \) and for an equivalent circuit it can be written as [5],

\[ R_{\theta JA} = R_{\theta JC} + R_{\theta CS} + R_{\theta SA} \left[ ^\circ C/W \right] \]  \hspace{1cm} (3.15)

\( R_{\theta JC} \): Junction to case resistance in \(^\circ\)C/W.

\( R_{\theta CS} \): Case to sink resistance in \(^\circ\)C/W.

\( R_{\theta SA} \): Sink to ambient resistance in \(^\circ\)C/W.
Thermal resistance from junction to case ($R_{\theta JC}$)

It is an internal thermal resistance from the chip junction to case of the package. It is determined by a package design and a frame material. It is measured at a case temperature of $T_{\text{case}} = 25 \, ^\circ\text{C}$ condition and can be illustrated by the following formula [11], [5].

$$R_{\theta JC} = \frac{(T_J-T_{\text{case}})}{P_D} \tag{3.16}$$

$P_D$: Total dissipated power

$T_J$: Junction temperature

$T_{\text{case}}$: Case temperature

$T_{\text{case}} = 25 \, ^\circ\text{C}$ is a condition with infinite heat sink.

The values for $R_{\theta JC}$ for the diodes and the IGBTs are given by the manufacturer and can be obtained from the datasheets.

Thermal resistance from case to sink ($R_{\theta CS}$)

It is a thermal resistance from case to the heat sink. It varies with the package and the type of insulators that are used. The thermal resistance of the insulators is available in the handbook [5] and also in the datasheets provided by the manufacturer.

Thermal resistance from sink to ambient ($R_{\theta SA}$)

It is a thermal resistance from the heat sink to the ambient. It can be determined by a geometric structure, surface area and quality of the heat sink [11]. The temperature of the heat sink rises during the operation of the power semiconductor devices due to the heat flow from the junction to an ambient. The increased temperature of the heat sink can be reduced either by natural convection cooling or by forced air. Forced air cooling has an advantage over natural cooling because of small size heat sinks and low thermal resistance. Moreover, reduced thermal time constant of the heat sink [5]. Usually the resistance of the heat sink is provided by the
manufacturer. In case the thermal resistance is not available, it can be calculated by using the method explained in chapter 29 of [5]. In the prototype converter the heat sinks are used with fans (forced air cooling). The description related to the fans and the thermal resistance of the heat sink (with and without fan) for both the cycloconverter and the VSC is available in the datasheet attached to the appendix.

3.3 RC-Snubber Losses

The RC-snubbers are mounted in the cycloconverter in parallel with each valve constituting a capacitor in series with a resistor. Whenever the voltage varies across the snubber capacitors they either discharge or recharge accordingly leading to energy dissipation. In the prototype converter, the following are instants where the losses due to these snubbers occur.

- VSC commutation i.e. transformer voltage reversal
- Cycloconverter commutation

3.3.1 VSC Commutation

The state of the switches in the cycloconverter does not change during the VSC commutation. The voltage across the non-conducting valves changes from $\pm U_d$ to $\mp U_d$. The potential difference occurs due to a difference between the rise/fall time of the transformer voltage and a voltage across the snubber capacitors that leads to a current flow through the snubber resistances and the valves which results in the power dissipation across them. The operating waveforms during the VSC commutation for one of the phase legs are shown in figure 3-3.

If the voltage across one of the snubber capacitors during the VSC commutation is $u_{snb}$ and the current that flows is $i_{snb}$, the dissipated energy can be found out as follows:
The instantaneous power can be expressed as,

\[ p(t) = u_{snb} \cdot i_{snb} \]  
(3.17)

The current through the snubber capacitor,

\[ i_{snb} = C_s \frac{du_{snb}}{dt} \]  
(3.18)

Using 3.18, 3.17 becomes

\[ p(t) = u_{snb} \cdot C_s \frac{du_{snb}}{dt} \]  
(3.19)

The power as a rate of energy can be written as

\[ p(t) = \frac{dE}{dt} = u_{snb} \cdot C_s \frac{du_{snb}}{dt} \Rightarrow dE = C_s \cdot u_{snb} \cdot du_{snb} \]  
(3.20)

Then integrating we have

\[ E_{snb, vsc} = C_s \int_{v_1}^{v_2} u_{snb} \, du_{snb} = C_s \cdot \left( \frac{(v_2 - v_1)^2}{2} \right) \]  
(3.21)

Similarly the lost charge can be obtained by,

\[ Q_{snb, vsc} = C_s \cdot \Delta U_{snb} = C_s (v_2 - v_1) \]  
(3.22)

Where,

\( C_s \): is the snubber capacitance

\( v_1 \): is the initial voltage, either \( \pm U_d \)

\( v_2 \): is the final voltage, either \( \mp U_d \)

\( u_{snb} \): is the instantaneous voltage across the snubber capacitor during commutation

\( i_{snb} \): is the instantaneous current through the snubber capacitor during discharge/recharge.

\( Q_{snb, vsc} \): represents the lost charge

\( \Delta U_{snb} \): voltage change across the capacitor during the commutation process

The energy loss for one of the snubber capacitors during the VSC commutation by using (3.21) will then be,

\[ E_{snb, vsc} = C_s \cdot \frac{(2U_d)^2}{2} \]  
(3.23)
Similarly the lost charge by using (3.22),

$$Q_{snb_{vsc}} = C_s \cdot (2U_d) \quad (3.24)$$

During the VSC commutation, three snubber capacitors discharge and recharge again completely. This happens twice in one complete transformer cycle. Therefore the snubber losses due to the VSC commutation can be estimated as

$$E_{CC_{vsc}} = 6E_{snb_{vsc}} \quad (3.25)$$

Similarly the lost charge can be estimated as,

$$Q_{RC_{vsc}} = 6Q_{snb_{vsc}} \quad (3.26)$$

### 3.3.2 Cycloconverter Commutation

The cycloconverter commutation occurs between the two successive VSC commutations and begins by turning-on the non-conducting valves thereby creating a short circuit condition for the charged snubbers parallel to the corresponding valves. The charged capacitors dissipate their stored energy across the respective resistance and the valve thereby discharging completely, Figure 3-4 (left). Similarly the snubbers across those valves that turn from conducting to a non-conducting state recharge themselves completely and equal amount of energy that is stored, dissipates across the corresponding resistances and the incoming valves, Figure 3-4 (left).

![Figure 3-4 Behavior of snubbers during CC commutation left side (recharge and discharge completely) right side (partial discharge of snubbers)](image)

The voltage across the valve that turns from a non-conducting to a conducting state changes from $\pm U_d$ to 0 and the snubber discharges completely. Whereas for the same phase leg the
other valve turns from a conducting to a non-conducting state leading to a recharge of the snubber capacitor completely with a voltage change from 0 to \( \pm U_d \). The energy loss for each capacitor can be estimated by using equation (3.21) as

\[
E_{snb,cc} = C_s \int_0^{U_d} u_{snb} \, du_{snb} = C_s \frac{(U_d)^2}{2}
\]  

(3.27)

The lost charge due to each snubber can be expressed as,

\[
Q_{snb,cc} = C_s \cdot \Delta U_{snb} = C_s \cdot U_d
\]  

(3.28)

Due to the cycloconverter commutation, three of the mounted snubbers discharge completely whereas other three recharge completely. For a complete transformer cycle the process repeats twice. Therefore by using equation (3.27), the total energy loss for one complete transformer cycle will be

\[
E_{cc} = 12 \cdot E_{snb,cc}
\]  

(3.29)

Similarly the lost charge due to all the snubbers will be,

\[
Q_{RC,cc} = 12 \cdot Q_{snb,cc}
\]  

(3.30)

### 3.3.3 Additional Losses due to cycloconverter commutation

Due to the leakage inductance both the valves in a phase leg that undergoes commutation, remain in on-state during the commutation time. This situation creates a short circuit condition leading to a discharge of the snubbers mounted in the other phase legs as illustrated in figure 3-4 (right). The duration of the commutation period is shorter than the discharge time therefore it results in a partial discharge of the snubbers leading to the additional snubber losses. At the end of commutation, the partially discharged snubbers begin to recharge again to the original level. Assuming \( U_{acc} \) as a voltage up to which the capacitors discharge partially and again recharge to the original level as in figure 3-5, the energy loss of one of the capacitors can be estimated as

\[
E_{cc,add} = C_s \int_{\pm U_d}^{\pm U_{acc}} u_{snb} \, du_{snb} + C_s \int_{\pm U_{acc}}^{\pm U_d} u_{snb} \, du_{snb}
\]  

(3.31) a

\[
E_{cc,add} = \frac{C_s (U_{acc} - U_d)^2}{2} + \frac{C_s (U_d - U_{acc})^2}{2}
\]  

(3.31) b

\[
E_{cc,add} = C_s \cdot (U_d - U_{acc})^2
\]  

(3.31) c

The lost charge due to one snubber will be will be,

\[
Q_{cc,add} = 2 \cdot C_s \cdot (U_d - U_{acc})
\]  

(3.32)
Since the snubbers in the two phase legs discharge partially during each cycloconverter commutation, the additional energy that will dissipate across the snubber resistances during each cycloconverter commutation can be expressed as,

\[ E_{cc,\text{add}} = 2 \cdot C_s \cdot (U_d - U_{acc})^2 \]  \hspace{1cm} (3.33)

The lost charge will be,

\[ Q_{cc,\text{add}} = 4 \cdot C_s \cdot (U_d - U_{acc}) \]  \hspace{1cm} (3.34)

The partial discharge of the capacitors occurs twice in one complete transformer cycle. Therefore, by using equation 3.33, the additional snubber losses due to all the cycloconverter commutations for one complete transformer cycle will be

\[ E_{cc,\text{add}} = 4 \cdot \sum_{i=1,2,3} (C_s \cdot (U_d - U_{acc})^2) \]  \hspace{1cm} (3.35)

Similarly the lost charge will be,

\[ Q_{RC,cc,\text{add}} = 8 \cdot \sum_{i=1,2,3} (C_s \cdot (U_d - U_{acc})) \]  \hspace{1cm} (3.36)

The total loss due to the snubbers for one fundamental cycle using equation 3.25, 3.29 and 3.36 can be written as

\[ P_{\text{tot}} = \frac{1}{T} \sum_{x=1}^{f_1} (E_{cc,\text{add}} + E_{\text{Snb,cc}} + E_{cc,\text{vse}}) \]  \hspace{1cm} (3.37)

\( f_1 \): is the fundamental frequency

\( f_{tr} \): is the transformer operating frequency

\( T \): is the fundamental period
3.4 Losses in Transformer and Reactor

The possible losses in the transformer and reactor are the load losses namely winding losses or copper losses and the no-load losses i.e. iron losses constituting hysteresis and eddy current losses.

The copper losses are due to the resistance that the inductor includes.

The hysteresis losses are related to the power which is required to magnetize and demagnetize the core. On the other hand, the eddy current losses are the result of small currents that appear in the core due to the magnetic AC field.

3.4.1 Core losses

There are a number of methods available in the literature, which deal with the magnetic loss determination [12]. These methods can be divided into three main models [12]

- Hysteresis Models
- Loss separation approach
- Empirical methods

The Hysteresis models are based on the Jiles-Atherton or Preisach Models. One of them uses a macroscopic calculation whereas other introduced a statistical approach for the description of time and space distribution of a domain-wall motion [12].

The Loss separation model has introduced the contribution of three different effects to the magnetization losses namely static hysteresis loss, eddy current loss and excess- eddy current loss. The accuracy of this model is improved after Bertotti’s physical explanation of excess loss.

Empirical approaches are based on the Steinmetz equation commonly known as curve-fitting expression for the measured data.

Although hysteresis models and loss separation models can be considered for the adequate results but they need extensive computations. On contrary, the empirical methods based on the Steinmetz equations utilize the manufacturers provided data with a simple expression to determine the magnetic losses.

Original Steinmetz equation

Steinmetz Introduce a general expression to characterize the core losses which is usually used in the design of magnetic power devices namely transformers, electrical machines and
inductors. It is also known as Original Steinmetz Equation (OSE), based on the curve fitting approach of the measured data under the sinusoidal excitation [12], [13],

\[ P_v = K f^\alpha \hat{B}_m^\beta \]  

(3.38)

Whereas, \( K, \alpha \) and \( \beta \) are the coefficients based on the material characteristics. \( \hat{B}_m \) is the peak magnetic flux density and \( f \) is the frequency.

However, a major drawback of equation-3.38 is that it is valid only for sinusoidal waveforms [12] [13]. However, in power electronic converters inductors and transformers are usually subjected to non-sinusoidal rectangular waveforms having positive, negative and zero amplitudes. This results in the ramping up/down and a constant flux in the cores [13]. It has also been proven that for a non-sinusoidal waveform, the losses are high as compared to a sinusoidal waveform, though the frequency and peak flux density are same [14]. The limitation of the OSE is overcome by the introduction of the modified forms of Steinmetz equation which are useful for the wide variety of waveforms. These are Modified Steinmetz equation, Generalized Steinmetz equation and Improved Generalized Steinmetz equation which are explained with mathematical expressions below.

**Modified Steinmetz equation**

In [15], Modified Steinmetz Equation (MSE) is considered effective due to the fact of relating the rate of change of magnetic induction with the core losses. In MSE, the frequency parameter in equation 3.38 is replaced with an equivalent frequency component which can be expressed as,

\[ f_{eq} = \frac{2}{\Delta B^2 \pi^2} \int_0^\pi \left( \frac{dB(t)}{dt} \right)^2 dt \]  

(3.39)

Where, \( \Delta B \) is the peak-to-peak magnetic induction.

By using the MSE, the losses can be estimated as

\[ P_v = (K f_{eq}^{\alpha-1} \hat{B}_m^\beta) f \]  

(3.40)

Where, \( f \) is the fundamental frequency.

**Generalized Steinmetz Equation**

MSE is useful for non-sinusoidal waves, but in [16] it is shown that the MSE does not show consistency with the simple Steinmetz equation i.e. mismatch between the OSE and MSE for the sinusoidal waveforms. Therefore the Generalized Steinmetz Equation (GSE) is considered
more appropriate. This is due to the fact that GSE considers both the rate of change of magnetic induction and the instantaneous value of magnetic induction $B(t)$, proposing

$$P_v = \frac{1}{T} \int_0^T k_1 |\frac{dB(t)}{dt}|^\alpha |B(t)|^{\beta-\alpha} dt$$  \hspace{1cm} (3.41)

**Improved Generalized Steinmetz Equation**

Since there are possibilities of minor asymmetrical hysteresis loops due to different excitation conditions especially in the case of inductor if used as a filter [17]. In [14] a method based on the Improved Generalized Steinmetz Equation (IGSE) is presented to cope with the minor hysteresis loops. In IGSE the instantaneous value of magnetic induction is replaced with its peak to peak value thus taking into account the time history of the material as well as the instantaneous value of magnetic induction.

The form of equation is [14]

$$P_v = \frac{1}{T} \int_0^T k_i |\frac{dB(t)}{dt}|^\alpha (\Delta B)^{\beta-\alpha} dt$$  \hspace{1cm} (3.42)

Where

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos(\theta)|^\alpha 2^{\beta-\alpha} d\theta}$$  \hspace{1cm} (3.43)

### 3.4.2 Core Losses in Transformer

The operating waves of the transformer are shown in figure 3-6. The windings of the transformer are excited by an equal width AC square wave form represented by $u_{tr}$. The resulting magnetic flux density is an equal width triangular waveform that can be obtained by the following equation,

$$B(t) = \frac{1}{N_{turn} A_{core}} \int u_{tr} \, dt$$  \hspace{1cm} (3.44)

$N_{turn}$ : Number of turns

$A_{core}$ : Effective Area of Core

$u_{tr}$ : Instantaneous transformer voltage

In the case of square wave, the amplitude of the voltage remains constant therefore the peak magnetic flux density can be obtained by [12],

$$\hat{B}_m = \frac{u_{dc}}{(4N_{turn} A_{core} f)}$$  \hspace{1cm} (3.45)
$U_{dc}$: is the amplitude of the applied voltage

$f$: Frequency of applied voltage

![Figure 3-6 Magnetic Flux density and Transformer applied voltage](image)

**Proposed method for core losses**

As the resulting flux density waveform in the transformer, excited by a square wave voltage, is an equal width triangular waveform. The area occupied by it will be approximately equal to the sinusoidal waveform of the same frequency, figure 3-7. In this case, for simplified calculations, the OSE can be used to calculate the core losses.

![Figure 3-7 Comparison of Triangular and sinusoidal waveform area](image)

### 3.4.3 Core Losses in Inductor

The equivalent model of one of the phases of the converter is shown in figure 3-8 along with the operating waveforms. The AC-filter inductor current contains a high frequency ripple component as well as a low–frequency output component. The voltage across the inductor varies depending upon the PWM voltage and the output voltage. Based on the inductor voltage the magnetic flux density $B(t)$ can be calculated according to the following equation,

\[
B(t) = \frac{1}{N_{\text{turn}} A_{\text{core}}} \int u_{\text{ind}} \, dt \tag{3.46}
\]
Where,

\[ N_{\text{turn}} \]: Number of turns
\[ A_{\text{core}} \]: Effective area of core
\[ u_{\text{ind}} \]: Instantaneous applied voltage

The current ripple corresponds to the small magnetic hysteresis loops along with the major loop as shown in figure-3.9. The small hysteresis loops are called dynamic minor loops. The area surrounded by each minor loop corresponds to the iron losses in one switching period. The number of minor loops depend upon the switching frequency, and their area depends upon the value of \( \Delta B \) and \( \Delta H \). Furthermore, the minor loops are generated at various places together with the variation in the magnetic field \( H \), figure 3-9. Since the core losses are associated with hysteresis loops, therefore, both the minor loops and a major loop correspond to the core losses [18].
Proposed method for Core loss calculation

The flux density waveform is not a pure sinusoidal waveform and includes a high frequency ripple therefore the OSE is not valid in this case. The modified forms of Original Steinmetz equation considers instantaneous flux density thus these methods can be used in this case. From the literature review in [14], [15] and [16] about the modified forms of Steinmetz equation, following conclusions are made.

By using the modified empirical methods, the MSE and IGSE show good loss determination. However, MSE does not show consistency with the OSE at high frequencies even for the sinusoidal waveforms [16] i.e. it overestimates the equivalent frequency and underestimates the power losses. Another drawback of the MSE is that, it exhibits anomalies for the high harmonic contents [16].

The anomalies of the MSE are overcome by introducing the IGSE, which is shown by [14]. The IGSE shows consistency with OSE even at high frequencies and also shows acceptable results for a wide variety of flux density waveforms. Furthermore, by using the IGSE, a method in [14] is presented to deal with the minor hysteresis loops that shows its validity for the waveforms containing high harmonic contents.

The comparison of the methods is beyond the scope of this thesis work. Therefore the literature review is a suitable tool to choose the appropriate method which directs towards IGSE for satisfactory results. Hence IGSE is chosen in this work to calculate the iron losses in the reactor.
3.4.4 Copper Losses

In modern Power Electronic converters, high switching frequency is used. This results in reduced transformer and inductor size. This however introduces skin, proximity and fringing flux effects [19]. Also because of non-sinusoidal current waveforms, additional copper losses occur due to harmonics. The winding offers different resistances for each frequency component, it is therefore important to accurately estimate the resistance of windings subjected to the high frequency currents. The actual resistance of the winding, when is subjected to a high frequency current is called AC-resistance. Under high frequency excitation, the copper losses can therefore be expressed as,

\[ P_{cu} = I_{rms}^2 R_{AC} \]  \hspace{1cm} (3.47)

Whereas \( R_{AC} \) is the ac-resistance of the windings and can be expressed as

\[ R_{AC} = F_{ac} R_{DC} \]  \hspace{1cm} (3.48)

\( R_{DC} \) is the DC-resistance that remains constant

\( F_{ac} \) is the factor that shows the increase in the DC resistance due to skin and proximity effects caused by the high frequency components.

The DC resistance of the windings can be approximated as

\[ R_{DC} = \frac{\rho + l}{A_{cond}} \]  \hspace{1cm} (3.49)

Where

\( \rho \) is the resistivity in Ohms-mm\(^2\)/m

\( l \) is the length of conductor in metres

\( A_{cond} \) is the cross sectional area of the conductor in mm\(^2\)

The copper losses due to the dc current and all the harmonics of the inductor current are expressed as [19],

\[ P_w = P_{wdc} + P_{wAC} = R_{dc} I_{dc}^2 + \sum_{n=1}^{\infty} R_{wn} I_n^2 \]  \hspace{1cm} (3.50)

Where, \( R_{wn} \) is the AC winding resistance at the \( n \)th frequency component and \( I_n \) is the r.m.s value of \( n \)th component.

\[ P_w = R_{dc} I_{dc}^2 \left( 1 + \sum_{n=1}^{\infty} \frac{R_{wn}}{R_{dc}} \left( \frac{I_n}{I_{dc}} \right)^2 \right) \]  \hspace{1cm} (3.51)
Power Loss Modeling of Isolated AC/DC converter

\[ P_w = R_{dc}I_{dc}^2 \left( 1 + \sum_{n=1}^{\infty} F_{Rn} \left( \frac{l_n}{l_{dc}} \right)^2 \right) \]  \hfill (3.52)

\[ F_{Rn} = \frac{R_{wn}}{R_{dc}} = \sqrt{n} X \left[ F_{RS} + \frac{2(n_f^2-1)}{3} F_{RP} \right] \]  \hfill (3.53)

Where, \( N_i \) is the number of layers and \( X \) for a round conductor is defined as

\[ X = \left( \frac{\pi}{4} \right)^{\frac{3}{2}} \frac{d}{\delta_w} \frac{\delta}{\sqrt{p}} \]  \hfill (3.54)

d is the diameter of conductor

\( p \) is the distance between the centers of the adjacent conductors

\( \delta_w \) represents the skin depth at \( n \)th frequency

\[ \delta_w = \frac{\rho_w}{\sqrt{n\mu_0 f_n}} \]  \hfill (3.55)

\( \rho_w \) is the resistivity of material, normally copper is used.

\( \mu_0 \) is the permeability of vacuum

\( f_n \) is the frequency of \( n \)th harmonic

\( F_{RS} \) is the skin effect factor that according to [19] can be approximated by

\[ F_{RS} = \frac{\sinh(2X\sqrt{n}) + \sin(2X\sqrt{n})}{\cosh(2X\sqrt{n}) - \cos(2X\sqrt{n})} \]  \hfill (3.56)

\( F_{RP} \) is the proximity effect factor that according to [19] can be approximated by

\[ F_{RP} = \frac{\sinh(X\sqrt{n}) - \sin(X\sqrt{n})}{\cosh(X\sqrt{n}) + \cos(X\sqrt{n})} \]  \hfill (3.57)

The methods explained in this chapter along with the set of equations are used in the coming chapters to calculate the losses.
4. Model of Prototype Converter

This chapter describes the converter model that is developed to compute the losses. The theory related to its implementation is described. Apart from this, the different outputs based on the developed model are also shown.

4.1 Introduction

To model the losses in the prototype converter, an appropriate model of the converter capable of reproducing the voltages and currents associated with elements is required. It is always desirable to avoid complex models. Therefore the prototype converter is implemented based on the simplified analysis described in chapter 2 and also in [3]. The prototype is modeled in MATLAB/Simulink by dividing it into various blocks. These are,

- Voltage Source Converter (VSC)
- Cycloconverter
- Transformer
- RC-snubbers
- Filter inductor.

4.1.1 VSC

In the VSC block, the voltage and current at the output of the VSC are obtained by using equation 2.1 and 2.3 respectively. The current through the IGBTs and diodes of the VSC during the conduction state is obtained by using equation 4.1 and 4.2 which are,

\[ i_{\text{IGBT}} = k_T i_{tr} \]  \hspace{1cm} (4.1)
\[ i_{\text{diode}} = k_F i_{tr} \]  \hspace{1cm} (4.2)

Where \( k_T \) and \( k_F \) are the coupling functions. Based on the control strategy, explained in [1], \( k_F \) & \( k_T \) = 1, when the diodes and IGBTs are conducting and \( k_F \) & \( k_T \) = 0 when both of them are in off-state.

The current obtained by 4.1 and 4.2 is used to compute the conduction losses in the IGBTs and diodes respectively. This block also includes the switching loss calculation for the VSC.
4.1.2 Cycloconverter

The cycloconverter is based on the carrier based PWM control strategy, which is explicitly explained in [1]. The voltage at the output of each phase of the cycloconverter, referred to the midpoint of the transformer winding is obtained by using equation 2.2.

The current through the IGBTs and diodes during the conduction state are obtained by equation 4.3 as,

\[ i_{sw,cyc} = k_{sw,cyc} \cdot i_{cyc,i} \]  \hspace{1cm} (4.3)

Where, \( i_{cyc,i} \) is the current at the output of the cycloconverter and depends on the load. \( k_{sw,cyc} \) is a coupling function and is ‘1’, when the diodes and IGBTs conduct and is ‘0’ when both devices are in off state.

The cycloconverter block also includes the calculation of conduction loss for the IGBTs and diodes as well as reverse recovery losses for the diodes.

The voltage and current waveforms during the switching state for both the VSC and cycloconverter are not obtained in this model. This is because the switching losses can be computed precisely by using the energy loss values given in the device datasheet. The required current at the instant, when switching starts/ends, is obtained in this model. This current is then used to obtain the energy values from the datasheet to compute the switching losses.

4.1.3 Transformer

In the transformer block, the voltage and the current for the transformer are obtained by using equation 2.1 and 2.3 respectively. It also includes the calculation of the core losses as well as the copper losses for the transformer.

4.1.4 RC Snubbers

In the RC-snubber block, the equations presented in section 3.3 are used. This block is implemented to compute the RC-snubbers losses during the VSC commutation and cycloconverter commutation.

4.1.5 Output Filter

An LC-filter is used at the output of the cycloconverter to filter the harmonics in the voltages and currents. The branches of LC-filter for all the three phases are shown in figure 4-1.
The Star connected load is considered at the output. The LC filter block is implemented in the MATLAB/Simulink by using the differential equations that represent the voltages and currents. The output voltages are referred to the midpoint of the transformer. The corresponding equations can be derived by considering figure 4-1.

\[
\begin{align*}
    u_{ac,i} &= L_{fil} \frac{dlcyc,i}{dt} + u_{load,i} + u_{no} \\
    u_{no} &= \frac{1}{3} (u_{ac,1} + u_{ac,2} + u_{ac,3}) \\
    i_{fil,i} &= C_{fil} \frac{du_{load,i}}{dt} \\
    u_{load,i} &= u_{ac,i} - L_{fil} \frac{dlcyc,i}{dt} - u_{no}
\end{align*}
\] (4.4)

Where, $i$ represents the phase number i.e. $i = 1,2,3$.

The above equations were used in MATLAB/Simulink to get the voltages and currents across the filter inductor, filter capacitor and the loads. Apart from waveforms, this block is also implemented to compute the copper and core losses in a three phase filter reactor.
4.2 MATLAB Implementation

The model of the prototype converter was implemented in MATLAB by using the set of equations explained above. The blocks for the VSC, transformer, cycloconverter and the filter were generated.

The model of the converter showing different blocks is shown in figure 4-2 and 4-3.

4.2.1 Simulation

The main data of the prototype converter, used for the simulations is listed in table 4-1.
Table 4-1 Prototype converter parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Switching frequency</td>
<td>$f_{\text{switch}} = 12\ kHz$</td>
</tr>
<tr>
<td>Output frequency of the voltage</td>
<td>$f_{\text{load}} = 50\ Hz$</td>
</tr>
<tr>
<td>Filter Inductance</td>
<td>$L_{\text{fil}} = 1.2\ mH$</td>
</tr>
<tr>
<td>Filter capacitance</td>
<td>$C_{\text{fil}} = 33\ \mu F$</td>
</tr>
<tr>
<td>DC link voltage</td>
<td>$U_d = 600V$</td>
</tr>
<tr>
<td>$\bar{U}_{\text{control}}$</td>
<td>220V</td>
</tr>
<tr>
<td>$M$</td>
<td>0.73</td>
</tr>
<tr>
<td>$R_{\text{load}}$</td>
<td>2.07 $\Omega$ Star connected</td>
</tr>
</tbody>
</table>

Based on the control strategy, figure 4-4 shows the coupling functions for the output voltage of the cycloconverter and the current through the IGBT and diode for one of the valves at a low frequency ratio of $m_f = 20$.

Figure 4-4 PWM strategy with coupling functions of Cycloconverter Voltage and its Valve current

Similarly, the coupling functions of the VSC for the IGBT and diode current are shown in figure 4-5.
By using the coupling functions and the set of equations in the above sections, the voltages and currents were obtained for the VSC and cycloconverter.

### 4.4 Simulation Results

The simulation results are shown for the parameters listed in Table 4-2.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input DC-link Voltage</td>
<td>$U_d = 600V$</td>
</tr>
<tr>
<td>Amplitude of Carrier wave</td>
<td>$\bar{U}_{\text{carrier}} = 300V$</td>
</tr>
<tr>
<td>Reference Voltage</td>
<td>$\bar{U}_{\text{control}} = 220V$</td>
</tr>
<tr>
<td>Modulation Index</td>
<td>$M = 0.73$</td>
</tr>
<tr>
<td>Load Resistance</td>
<td>$R_{\text{load}} = 2.07\ \Omega$</td>
</tr>
<tr>
<td>Switching Frequency</td>
<td>$f_{\text{switch}} = 12\ kHz$</td>
</tr>
<tr>
<td>Frequency of Output Voltage</td>
<td>$f_{\text{load}} = 50\ Hz$</td>
</tr>
</tbody>
</table>

By using equation 2.2, the output voltage of the cycloconverter was obtained. Figure 4-6 shows the output voltage of the cycloconverter for a few transformer cycles.
By using equation 4.3, the current through the valves of the cycloconverter was obtained, Figure 4-7.

By using equation 2.3, the transformer current was obtained, Figure 4-8.
By using equation 4.1 and 4.2, the currents through the diodes and IGBTs in the VSC were obtained, Figure 4-9

Similarly, the current through the filter inductor was obtained, (ref. section 4.1.5). Figure 4-11 shows the current through the inductor.
The required currents and voltage waveforms obtained by the implemented MATLAB model are then used in chapter 5 to obtain the losses in all the components of the converter.
5. Calculation of Losses

The aim of this chapter is to calculate the losses in different parts of the converter together with their comparison with the measured results. The losses are computed by using the current and voltage waveforms along with the known properties of the different components. The obtained losses are then compared with the measured results.

The methods to compute the losses in the different components of the converter are explained in the previous chapters along with the converter model to obtain the waveforms of the currents and voltages. Those methods are used in this chapter to compute the losses. From the set of equations, it can be observed that the losses are voltage and current dependent. It is therefore important to observe the behavior of the losses under different circumstances.

The losses that are shown in this chapter are

- Power semiconductor losses (conduction and switching losses)
- Snubber losses
- Transformer (winding and core losses)
- Inductor losses (winding and core losses)

5.1 Power semiconductor losses

The semiconductor losses hold major contribution in the total losses of converter. The semiconductor losses were computed by using the simulation waveforms in MATLAB together with the methods presented in chapter 3. The power semiconductor losses are comprised of conduction and switching losses. The resulted losses for the semiconductors, mounted in the prototype converter are shown in table 5-1.

5.1.1 Conduction losses

To calculate the conduction losses, the on-state voltage as a function of forward current, available in the data sheet is used. The on-state voltages as a function of forward current for the IGBT and diode modules specified for the VSC and cycloconverter are shown in figure 5-1 and 5-2. The on-state voltages for the IGBTs and diodes were extracted from the data sheets by using the curve fitting approach. The conduction losses for power semiconductor devices in the
VSC and cycloconverter were obtained by using equation 3.3 and 3.4. The conduction characteristics were assumed at a constant junction temperature of $T_j = 125^\circ C$.

![Figure 5-1 On-state Voltage vs. forward current of Diode (left) and IGBT in Cycloconverter](image1)

![Figure 5-2 On-state Voltage vs. forward current of Diode (right) and IGBT in VSC](image2)

### 5.1.2 Switching Losses

The switching losses in the prototype converter that need to be computed are the turn-off losses of the IGBTs in the VSC and the diode reverse recovery losses in the cycloconverter. The methods to calculate these losses are explained in chapter 3.
**Switching losses in IGBTs**

The switching losses of the IGBTs are computed by taking into account two cases which are,

- Hard turn-off
- Soft turn-off

In case of a hard turn-off, the turn-off energy values as a function of current provided in the datasheet for a specific device are used. The turn-off energy curve for an IGBT module, specified for the VSC is shown in figure 5-3. The energy values were retrieved from the datasheet by using the curve-fitting approach. Equation 3.7 was used to calculate the total switching losses over one fundamental frequency cycle. The switching characteristics were assumed at a constant junction temperature of $T_j = 125^\circ C$.

![Figure 5-3 Hard Turn-off energy values provided by the manufacturer](image)

In case of a soft turn-off, the turn-off energy values available in [3] were used. The turn-off energy values for different snubber capacitors as a function of current are shown in figure 5-4. The total switching losses were computed by using equation 3.7.
Diode reverse recovery losses

The diode reverse recovery losses were also computed by using the datasheet values. The obtained recovered charge for the specific device at a forward current of 100 A, a voltage of -600 V and a junction temperature of $T_j = 125^\circ C$ is $Q_{rr} = 11 \mu C$. The energy loss based on the recovery charge was computed by using equation 3.9. The total losses due to reverse recovery were computed by using equations (3.7) and (3.11).

Total power semiconductor losses

The total semiconductor losses at an input DC-link voltage of $U_d = 600V$, and a modulation index of $M = 0.73$ for different load currents are listed in table 5-1.

<table>
<thead>
<tr>
<th>$R_{load}$ (Ω)</th>
<th>$I_{load}$ rms (A)</th>
<th>Conduction Losses (Watts)</th>
<th>Switching Losses (Watts)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>VSC</td>
<td>Cycloconverter</td>
</tr>
<tr>
<td>15.55</td>
<td>10</td>
<td>25</td>
<td>58.35</td>
</tr>
<tr>
<td>10.37</td>
<td>15</td>
<td>40.22</td>
<td>99</td>
</tr>
<tr>
<td>7.77</td>
<td>20</td>
<td>57.47</td>
<td>145.2</td>
</tr>
<tr>
<td>6.22</td>
<td>25</td>
<td>76.83</td>
<td>196.2</td>
</tr>
<tr>
<td>5.18</td>
<td>30</td>
<td>97</td>
<td>251.39</td>
</tr>
</tbody>
</table>
5.1.3 Thermal model

The junction to case resistance \( R_{\theta,JC} \) was obtained from the data sheet provided by the manufacturers. The case to sink resistance \( R_{\theta,CS} \) depends upon the insulation material and is also specified by the manufacturer. The sink to ambient resistance \( R_{\theta,SA} \) depends upon the area of the heat sink which is designed for a specific application. In this case, \( R_{\theta,SA} \) is also available for the heat sinks for both the cycloconverter and VSC. Initially, the junction temperature is assumed \( T_{jo} = 125^\circ C \) whereas, the ambient temperature \( T_{amb} = 25^\circ C \) is considered. The actual junction temperature was then obtained by using the idea illustrated in the flow chart, figure 5-5.

![Flow Chart]

Figure 5-5 Method to compute thermal model of semiconductors

**Cycloconverter**

The thermal resistances \( R_{\theta,JC} \) were extracted from the datasheet. \( R_{\theta,JC-IGBT} = 0.18^\circ C/W \) and \( R_{\theta,JC-diode} = 0.36^\circ C/W \) were found for the IGBT and the diode respectively. The case to sink resistance \( R_{\theta,CS} = 0.06^\circ C/W \) was found. The thermal resistance of the heat sink with the forced air cooling was obtained from the datasheet which is \( R_{\theta,SA} = 0.04^\circ C/W \). The junction temperature was then calculated, the outcomes at different load currents are listed in table 5-2.
Table 5-2 Junction temperature of IGBT and diode for cycloconverter at different load current

<table>
<thead>
<tr>
<th>$R_{load}$ ($\Omega$)</th>
<th>$I_{load}$ rms (A)</th>
<th>Total Losses $P_D$ (Watts)</th>
<th>$T_j$ ($^\circ$C)</th>
<th>$R_{th-JC}$ ($^\circ$C/W)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Diode</td>
<td>IGBT</td>
<td>Diode</td>
</tr>
<tr>
<td>15.55</td>
<td>10</td>
<td>23.25</td>
<td>35.1</td>
<td>26</td>
</tr>
<tr>
<td>10.37</td>
<td>15</td>
<td>38.8</td>
<td>60.13</td>
<td>27</td>
</tr>
<tr>
<td>7.77</td>
<td>20</td>
<td>56.4</td>
<td>88.7</td>
<td>28</td>
</tr>
<tr>
<td>6.22</td>
<td>25</td>
<td>75.6</td>
<td>120.6</td>
<td>29</td>
</tr>
<tr>
<td>5.18</td>
<td>30</td>
<td>95.8</td>
<td>155.5</td>
<td>30</td>
</tr>
</tbody>
</table>

VSC

In the VSC same method is used as for the cycloconverter. The parameters $R_{\theta,JC}$, $R_{\theta,CS}$ and $R_{\theta,SA}$ were extracted from the datasheet for the diode and IGBT. The losses in the VSC constitute the switching losses of the IGBTs and the conduction losses of the IGBTs and diodes. The junction temperatures are obtained at hard turn-off of IGBTs in VSC. The resulting junction temperatures and the corresponding resistances are listed in table 5-3.

Table 5-3 Junction temperature of IGBT and diode for VSC at different load current

<table>
<thead>
<tr>
<th>$R_{load}$ ($\Omega$)</th>
<th>$I_{load}$ rms (A)</th>
<th>Total Losses $P_D$ (Watts)</th>
<th>$T_j$ ($^\circ$C)</th>
<th>$R_{th-JC}$ ($^\circ$C/W)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Diode</td>
<td>IGBT</td>
<td>Diode</td>
</tr>
<tr>
<td>15.55</td>
<td>10</td>
<td>4.02</td>
<td>99.86</td>
<td>25.8</td>
</tr>
<tr>
<td>10.37</td>
<td>15</td>
<td>6.22</td>
<td>130</td>
<td>25.8</td>
</tr>
<tr>
<td>7.77</td>
<td>20</td>
<td>8.57</td>
<td>163.5</td>
<td>26</td>
</tr>
<tr>
<td>6.22</td>
<td>25</td>
<td>11.2</td>
<td>197</td>
<td>26</td>
</tr>
<tr>
<td>5.18</td>
<td>30</td>
<td>13.86</td>
<td>233</td>
<td>27</td>
</tr>
</tbody>
</table>

5.2 Snubber losses

The formulation of the RC-snubber losses is given in section 3.3. The snubber losses are required to compute during the VSC commutation and cycloconverter commutation.

The snubber losses due to the VSC commutation were obtained by using equation 3.25.

The snubber losses due to the cycloconverter commutation were divided into two parts, one due to complete and the other due to partial discharge and recharge of the snubbers.
The losses due to complete discharge/recharge of the snubbers were obtained by using equation 3.28. The additional snubber losses due to the cycloconverter commutation as explained in section 3.3 were obtained by using equation 3.35.

In a similar fashion, the lost charge due to the VSC commutation and cycloconverter commutation was obtained by using the equations presented in section 3.3. The lost charge during the commutation processes is shown in table 5-4.

The total snubber losses for one complete fundamental cycle were obtained by using equation 3.37 are shown in table 5-4 at different load currents at an input DC link voltage of $U_d = 600V$ and a modulation index of $M = 0.73$.

<table>
<thead>
<tr>
<th>$R_{load}$ (Ω)</th>
<th>$I_{load}$ rms (A)</th>
<th>VSC commutation</th>
<th>Cycloconverter commutation</th>
<th>$Q_{total}$ (coulomb)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>$P_{loss, vsc}$ (Watts)</td>
<td>$Q_{RC, vsc}$ (coulomb)</td>
<td>$Q_{RC, vsc}$ (coulomb)</td>
</tr>
<tr>
<td>15.55</td>
<td>10</td>
<td>389</td>
<td>0.012</td>
<td>194</td>
</tr>
<tr>
<td>10.37</td>
<td>15</td>
<td>389</td>
<td>0.012</td>
<td>194</td>
</tr>
<tr>
<td>7.77</td>
<td>20</td>
<td>389</td>
<td>0.012</td>
<td>194</td>
</tr>
<tr>
<td>6.22</td>
<td>25</td>
<td>389</td>
<td>0.012</td>
<td>194</td>
</tr>
<tr>
<td>5.18</td>
<td>30</td>
<td>389</td>
<td>0.012</td>
<td>194</td>
</tr>
</tbody>
</table>

**5.3 Transformer Losses**

The transformer includes copper losses and the core losses. The methods to find these losses are explained in section 3.4.

**5.3.1 Copper Losses**

The copper losses were computed by using equation 3.50. Since the current flowing through the windings also contains harmonic components. Therefore, for the ease of calculations, the AC resistance of each frequency component was obtained by using the measured values given in [3]. The AC resistances of transformer winding at different frequencies are shown in figure 5-6.
The amplitude spectrum including the fundamental and the dominant harmonic components of the transformer current, the winding resistances for the corresponding frequency components and the resulting power losses are shown in figure 5-7.

The total copper losses of the windings are the sum of the losses due to each frequency component. It can be observed that the major contribution is due the fundamental component (6 kHz) of the current whereas the harmonic components show much smaller contribution.

The copper losses in the transformer windings as a function of load current are shown in table 5-5.
Table 5-5 Copper losses in Transformer at $U_d = 600\,V$ and $M = 0.73$

<table>
<thead>
<tr>
<th>$R_{load}$ (Ω)</th>
<th>$I_{load, rms}$ (A)</th>
<th>Copper Losses (Watts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>15,55</td>
<td>10</td>
<td>4.16</td>
</tr>
<tr>
<td>10,37</td>
<td>15</td>
<td>9</td>
</tr>
<tr>
<td>7,77</td>
<td>20</td>
<td>15</td>
</tr>
<tr>
<td>6,22</td>
<td>25</td>
<td>22.2</td>
</tr>
<tr>
<td>5,18</td>
<td>30</td>
<td>27.7</td>
</tr>
</tbody>
</table>

5.3.2 Core Losses

The methods to compute the core losses are presented in section 3.4. The core losses are computed by using the original Steinmetz equation, although it is valid only for sinusoidal waveforms. This is because the resulting flux density waveform based on equation 3.44 is a triangular waveform with constant frequency, which will acquire almost the same area as the sinusoidal waveform of same frequency does. Therefore, for simplified calculations, the OSE is considered.

The peak flux density was obtained equal to $\hat{B}_m = 0.33\,T$ by using equation 3.45 for the parameters listed in table 5-6.

Table 5-6 Parameters to calculate peak flux density

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$N_{turn}$</td>
<td>24</td>
</tr>
<tr>
<td>$A_{core}$</td>
<td>$32,cm^2$</td>
</tr>
<tr>
<td>$f$</td>
<td>6 kHz</td>
</tr>
<tr>
<td>$U_{dc}$</td>
<td>600 V</td>
</tr>
</tbody>
</table>

The Steinmetz coefficients in the OSE were obtained by using the curve fitting approach in which OSE was compared with the datasheet values (attached to Appendix), Figure 5-8.
The coefficients obtained for the specified core of the transformer are listed in table 5-7.

<table>
<thead>
<tr>
<th>Coefficients</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( K )</td>
<td>( 5.6 \times 10^{-3} )</td>
</tr>
<tr>
<td>( \alpha )</td>
<td>1.36</td>
</tr>
<tr>
<td>( \beta )</td>
<td>2</td>
</tr>
</tbody>
</table>

The core losses for the parameters listed in table 5-6 are shown in table 5-8.

<table>
<thead>
<tr>
<th>Method</th>
<th>Core Losses (Watts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>OSE</td>
<td>92.8</td>
</tr>
</tbody>
</table>

### 5.4 Inductor losses

The inductor losses comprise of core losses and the winding losses. The method to compute these losses are explained in section 3.4.

#### 5.4.1 Copper Losses

The current that flows through the reactor winding has a high frequency ripple. Therefore it is important to analyze the frequency components of the current. This is because the resistance of the winding increases due to skin and proximity effects at higher frequencies. The appropriate calculation of copper losses will then include the losses due to fundamental
component as well as due to high frequency components. In figure 5-9, the frequency components of the current are shown. The fundamental frequency component is dominant, whereas the components at high frequencies have lower amplitudes.

![Frequency components of reactor current](image)

Figure 5-9 Harmonic components of reactor current at $I_{load} = 30$ A rms

For appropriate calculation of copper losses, the resistances of the windings at fundamental as well as at high frequencies are required. They can be either measured, or can be obtained analytically. The accurate measurement of resistance is not possible in this case because the windings are wound on the core and also cannot be separated. If for example LCR meter is used, the core losses will be included in the measurements that will result in the inaccurate resistance values.

Secondly, the resistance can be obtained analytically but the required parameters for these calculations are insufficient.

Since the amplitudes of high frequencies components of the current are very low, the resulting losses because of them will also be low. Therefore it is assumed that these losses can be disregarded. But their contribution in the total losses will result in more precise results.

The resulting copper losses for all the three phases, by considering only the fundamental component in the reactor windings are shown in table 5-9.
Table 5.9 Total copper losses in the reactor for all the three phases due to fundamental component of current at $U_d = 600V$ and $M = 0.73$

<table>
<thead>
<tr>
<th>$R_{load}$ (Ω)</th>
<th>$I_{load}$ rms (A)</th>
<th>Copper Losses (Watts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>15.55</td>
<td>10</td>
<td>11.1</td>
</tr>
<tr>
<td>10.37</td>
<td>15</td>
<td>25</td>
</tr>
<tr>
<td>7.77</td>
<td>20</td>
<td>44.4</td>
</tr>
<tr>
<td>6.22</td>
<td>25</td>
<td>70</td>
</tr>
<tr>
<td>5.18</td>
<td>30</td>
<td>100</td>
</tr>
</tbody>
</table>

### 5.4.2 Core Losses

The methods to compute the core losses are explained in chapter 3. The required instantaneous flux density was obtained by using equation 3.46, Figure 5-10.

The Steinmetz coefficients for the reactor core were obtained by using curve fitting method in which the OSE was compared with the provided data (attached in appendix), Figure 5-11. The resulting coefficients are listed in table 5-10.
Table 5-10 Steinmetz coefficients for specified inductor core

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K$</td>
<td>$2.1 \times 10^{-4}$</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>1.72</td>
</tr>
<tr>
<td>$\beta$</td>
<td>1.3</td>
</tr>
</tbody>
</table>

The core losses calculated by using the IGSE are shown in table 5-11 at $I_{\text{load}} = 30 \text{ Amp rms}$ and $V_{\text{load}} = 220 \text{ V}$.

Table 5-11 Core losses in the reactor core

<table>
<thead>
<tr>
<th>Method</th>
<th>Core Losses (Watts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>IGSE</td>
<td>151</td>
</tr>
</tbody>
</table>

5.5 Comparison of results

The losses calculated in the different components of the converter are summed up and then compared with the measurements. The total losses in the converter were measured in [20] at different conditions which are,

- Load current variation
- Modulation index variation
- Input DC link voltage variation

The analytical results are then compared with the measurements to validate the analytical loss model at different conditions.
5.5.1 Variable Load current

The comparison of the measurements and the analytical results is done under the following conditions:

- Load Current: Varied by varying the load resistance
- Modulation Index, Input DC-link voltage and output voltage: Constant
- Mode of operation: Soft-switched VSC with snubber capacitors
- Transformer disconnected

The results are shown in table 5-13 and comparison of measurements and analytical results is shown in figure 5-12 for the parameters listed in table 5-12.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$U_d$</td>
<td>600 V</td>
</tr>
<tr>
<td>$U_{load}$</td>
<td>220 V</td>
</tr>
<tr>
<td>$f_{sw}$</td>
<td>12 kHz</td>
</tr>
<tr>
<td>$M$</td>
<td>0.73</td>
</tr>
<tr>
<td>$R_{load}$</td>
<td>Star connected; variable</td>
</tr>
</tbody>
</table>

Table 5-13 Comparison of Measured and analytical results for parameters in Table 5-12

<table>
<thead>
<tr>
<th>$R_{load}$ (Ω)</th>
<th>$I_{load}$ (rms)</th>
<th>$P_{semi}$ (Watts)</th>
<th>$P_{snub}$ (Watts)</th>
<th>$P_{reac}$ (Watts)</th>
<th>$P_{total}$ (Watts)</th>
<th>$P_{msrd}$ (Watts)</th>
<th>Difference (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>15,55</td>
<td>10</td>
<td>117</td>
<td>584</td>
<td>134</td>
<td>835</td>
<td>1020</td>
<td>18</td>
</tr>
<tr>
<td>10,37</td>
<td>15</td>
<td>185.2</td>
<td>586</td>
<td>148</td>
<td>919</td>
<td>1150</td>
<td>20</td>
</tr>
<tr>
<td>7,77</td>
<td>20</td>
<td>262.6</td>
<td>588</td>
<td>186</td>
<td>1036</td>
<td>1280</td>
<td>19</td>
</tr>
<tr>
<td>6,22</td>
<td>25</td>
<td>345</td>
<td>591</td>
<td>211</td>
<td>1147</td>
<td>1400</td>
<td>18</td>
</tr>
<tr>
<td>5,18</td>
<td>30</td>
<td>434</td>
<td>594</td>
<td>251</td>
<td>1278</td>
<td>1510</td>
<td>15</td>
</tr>
</tbody>
</table>
From table 5-13, the contribution of the losses in different elements of the converter can be seen. At low currents, the major contribution of losses is due to RC-snubbers, and these losses increase slightly with the increased current. On the other hand, reactor and semiconductor losses also show significant contribution and increase with the increased load current.

The difference between the measured and analytical calculations is between 15 ~ 20%. This is because some losses due to insignificant contribution are not included. Those are the turn-on losses of the IGBTs in the cycloconverter. Additionally, the AC-resistance of reactor winding is not available; therefore the copper losses due to high frequency components are also not included.

In this case, the transformer is disconnected and replaced with the coil of fewer turns. Referring to section 2.5.2, if the transformer is removed, the current derivative will increase due to the absence of leakage inductance. As a result, the overlap between the voltage and current will increase, consequently the turn-on losses will increase. Since the resonant commutation is also employed, this principle includes additional turn-on of IGBTs in the cycloconverter during VSC commutation and results in additional turn-on losses. Therefore these losses will contribute to some amount.

If the turn-on losses of IGBTs in cycloconverter and additional winding losses of reactor are added-up to the total losses the difference will be reduced.
In the case of turn-on losses, the above mentioned reason is validated in section 5.5.3 in which transformer is connected and converter is operated without VSC snubbers. Although these losses are not counted, yet the difference between the measured and analytical results is less as compared to the difference in table 5-13.

5.5.2 Low Input DC-link Voltage $U_{dc} = 300 V$

In this section, the input DC-link voltage is reduced to $U_{dc} = 300 V$. The comparison of the analytical and measured results is done under the following conditions

- Load Current: Varied by varying the load resistance
- Modulation Index, Input DC-link voltage and output voltage: Constant
- Mode of operation: Soft-switched VSC with snubber capacitors
- Transformer disconnected

The results were compared for the parameters listed in table 5-15. The comparison of the results is shown in table 5-16 and figure 5-13.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$U_{id}$</td>
<td>300 V</td>
</tr>
<tr>
<td>$U_{load}$</td>
<td>110V</td>
</tr>
<tr>
<td>$f_{sw}$</td>
<td>12 kHz</td>
</tr>
<tr>
<td>$M$</td>
<td>0.74</td>
</tr>
<tr>
<td>$R_{load}$</td>
<td>Star connected; variable</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>$R_{load}$ (Ω)</th>
<th>$I_{load}$ (rms)</th>
<th>$P_{semi}$ (Watts)</th>
<th>$P_{snub}$ (Watts)</th>
<th>$P_{reac}$ (Watts)</th>
<th>$P_{total}$ (Watts)</th>
<th>$P_{msrd}$ (Watts)</th>
<th>Difference (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>7.7</td>
<td>10</td>
<td>96.4</td>
<td>147</td>
<td>50</td>
<td>297</td>
<td>390</td>
<td>24</td>
</tr>
<tr>
<td>5.18</td>
<td>15</td>
<td>159.2</td>
<td>148</td>
<td>64</td>
<td>372</td>
<td>480</td>
<td>23</td>
</tr>
<tr>
<td>3.89</td>
<td>20</td>
<td>228</td>
<td>150</td>
<td>90</td>
<td>465</td>
<td>590</td>
<td>21</td>
</tr>
<tr>
<td>3.12</td>
<td>25</td>
<td>302</td>
<td>152</td>
<td>111</td>
<td>560</td>
<td>700</td>
<td>20</td>
</tr>
<tr>
<td>2.59</td>
<td>30</td>
<td>389</td>
<td>155</td>
<td>146</td>
<td>676</td>
<td>800</td>
<td>15</td>
</tr>
</tbody>
</table>
5.5.3 Modulation Index Variation

In this case, the modulation index is varied under the following conditions:

- Modulation Index and Load current: Varied by varying the output voltage
- Input DC-link Voltage: Constant
- Mode of operation: Hard-switched VSC without snubber capacitors
- Transformer connected

The resulting losses are shown in table 5-17 and figure 5-14 for the parameters listed in table 5-12 with the variable output voltage and the modulation index.

Table 5-17 Comparison of losses at $R_{\text{load}} = 1.7 \, \Omega$ with transformer connected and hard turn-off of IGBTs in VSC

<table>
<thead>
<tr>
<th>$M$</th>
<th>$I_{\text{load}}$ (rms)</th>
<th>$P_{\text{semi}}$ (Watts)</th>
<th>$P_{\text{tran}}$ (Watts)</th>
<th>$P_{\text{snub}}$ (Watts)</th>
<th>$P_{\text{reac}}$ (Watts)</th>
<th>$P_{\text{total}}$ (Watts)</th>
<th>$P_{\text{msrd}}$ (Watts)</th>
<th>Difference (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.12</td>
<td>15</td>
<td>241</td>
<td>95.5</td>
<td>586</td>
<td>264</td>
<td>1186</td>
<td>1290</td>
<td>8</td>
</tr>
<tr>
<td>0.15</td>
<td>20</td>
<td>324</td>
<td>101</td>
<td>588</td>
<td>312</td>
<td>1325</td>
<td>1450</td>
<td>8.6</td>
</tr>
<tr>
<td>0.19</td>
<td>25</td>
<td>413</td>
<td>109</td>
<td>591</td>
<td>337</td>
<td>1450</td>
<td>1620</td>
<td>10.4</td>
</tr>
<tr>
<td>0.23</td>
<td>30</td>
<td>511</td>
<td>114</td>
<td>594</td>
<td>372</td>
<td>1588</td>
<td>1780</td>
<td>10.7</td>
</tr>
</tbody>
</table>
In this case the difference between the measured and analytical results is very low. This is because the snubbers in the VSC are removed as a result of which, the resonant commutation is not employed. This means the additional turn-on losses of cycloconverter will not exist. Also the transformer is connected; the current derivative will reduce due to the leakage inductance. Ultimately, the turn-on losses will be reduced to a much lower value.

On the other hand it is also observed that, by reducing the modulation index at the same current, the computed losses in the semiconductors and snubbers remain the same, whereas the losses are increased in the reactor. The increase in the reactor losses is due to the fact that the width of the PWM voltage varies by reducing the modulation index that increases the magnitude of the ripple of the flux density, which expands the vertical and horizontal dimensions i.e. $\Delta B$ and $\Delta H$ of each minor hysteresis loop, Figure 5-15. Consequently the area of the minor loops increases, which in turn increases the iron losses in the reactor core. This phenomenon is observed both analytically and from the measurements and can be understood by comparing the reactor losses in table 5-13 at M=0.73 and table 5-17 (with low M) at the same load currents and the Input DC-link voltage.
Results and Discussion

The analytical calculations were carried out to observe the behavior of the power semiconductor losses at different conditions.

The conduction losses were found to be increasing with the increased load current at constant modulation index and Input DC-link voltage. This is due to the fact that the on-state voltage is directly proportional to the device current, equation 5.1. Also, the parameters \( V_{Th} \) and \( r_{on} \) in equation 5.1 are constants and do not vary with currents and voltages therefore the conduction losses were found purely current dependent and independent of input DC-link voltage and modulation index. This phenomenon was observed by varying modulation index and DC-voltage in tables 5-13, 5-16 and 5-17.

\[ v_{on-state} = V_{Th} + r_{on}i_s \]  

(5.1)

The switching losses were found to be increasing with increased input DC-link voltage and the load current. This is due to the fact that the input voltage and current appear across devices during transition state and cause increase in the losses with increased voltage and current.

The snubber losses were found to be increasing with increased input voltage because the energy stored across them depends upon that input voltage. These losses were also found increasing slightly with increased load current at constant DC link voltage because of the commutation time of phase legs that corresponds to partial discharge and recharge of the snubber capacitors.
The winding losses in the transformer and reactor were found to be increasing with increased load current due their direct proportionality with current, Equation 5.2.

\[ P_{winding} = I_{rms}^2 R_{winding} \]  

(5.2)

The core losses in the transformer were computed by using the Original Steinmetz Equation. On the other hand the core losses in the reactor were found by using the Improved Generalized Steinmetz equation.

The core losses in the reactor were found to be increasing with the increased load current because of the peak flux density which is directly proportional to the current, Equation 5.3.

\[ \bar{B}_m = \frac{L_{ind} I_{ind}}{N_{turn} A_{core}} \]  

(5.3)

The core losses in the reactor were also found to be increasing with reduced modulation index at the same load current and input DC-link voltage. This is due to the increase in the ripple that leads to the increase in the area of each minor hysteresis loop and in turn the core losses.

The comparison of the analytical and measured results shows difference. The reasons could be as follows:

- The loss data of the core of the reactor was unknown, therefore a core having a same lamination thickness was assumed. It is obvious that the assumed core will have different loss data therefore results can be varied but not very high.
- The losses in the filter capacitors and input DC-link capacitor are not computed because of their negligible contribution.
- The turn-on losses in the cycloconverter’s IGBTs are not taken into account. Although a soft-switching principle is used, yet low voltage and current overlaps that offers small amount of losses.
6. Conclusion

In this diploma work, the power losses in a prototype converter are investigated and modeled. The losses are obtained by using MATLAB/Simulink at different conditions of load, modulation index and Input voltage. The obtained results are compared with the measurements at different conditions, and it is found that the model shows an adequate results.

In chapter 5, the losses in the components offering major contributions to the total losses are presented. The comparison of losses based on the described methods and the measured results shows the validity of achieved loss model. Furthermore, the contribution of each element is also obtained which is useful to optimize the elements in the converter.

The comparison of improved and old loss model is shown in figure 6-1. The figure shows consistency in the measured and new loss model results therefore it can be concluded that all the losses are taken into account in the developed loss model.

![Comparison of old and new loss model with measured losses](image)

*Figure 6-1 Comparison of Analytical (old and new model) and measured results without VSC snubbers*

In figure 6-2, the contribution of different losses in the converter is shown and can be observed that the snubbers and power semiconductors have a significant contribution in the total losses. However, the reactor and transformer also show noticeable contribution.
Based on the comparison of results at different conditions, it can be concluded that this loss model can be further used to optimize the converter.

**Future Work**

After investigating the losses and their dependency upon different parameters, the following tasks can be considered for future work.

- Optimization of the prototype converter.
- Investigations about the elements to improve the converter design.
Works Cited


[18] Toshiba Schimizu Kwanryol Kim, "Dynamic Iron Loss measurement method for AC filter Inductor on a PWM inverter", Metropolitan University, Tokyo, Japan.


Three Phase SPWM VSC using Switching Function Concept", Amir Kabir University of Technology, Tehran Polytechnic, Tehran, Iran.

Appendix

VSC Data Sheet


Cycloconverter Data Sheet


Loss Curves of Transformer
Reactor Core Loss Data

**Typical data for SURA® M235-35A**

<table>
<thead>
<tr>
<th>T</th>
<th>W/kg at 50 Hz</th>
<th>VA/kg at 50 Hz</th>
<th>A/m at 50 Hz</th>
<th>W/kg at 100 Hz</th>
<th>W/kg at 200 Hz</th>
<th>W/kg at 400 Hz</th>
<th>W/kg at 1000 Hz</th>
<th>W/kg at 2500 Hz</th>
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<tbody>
<tr>
<td>0.1</td>
<td>0.02</td>
<td>0.05</td>
<td>24.7</td>
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<td>14.3</td>
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<td>1.88</td>
<td>7.45</td>
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<td>4.73</td>
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<td>1950</td>
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<td>12000</td>
<td>9.70</td>
<td>24.5</td>
<td>82.5</td>
<td>336</td>
<td>1524</td>
</tr>
</tbody>
</table>

| Loss at 1.5 T, 50 Hz, W/kg | 2.25 |
| Loss at 1.0 T, 50 Hz, W/kg | 0.92 |
| Anisotropy of loss, %     | 10   |

Magnetic polarization at 50 Hz

- H = 2500 A/m, T
- H = 3000 A/m, T
- H = 10000 A/m, T

Coercivity (DC), A/m

- 35

Relative permeability at 1.5 T

- 610

Resistance, μΩcm

- 59

Yield strength, N/mm²

- 460

Tensile strength, N/mm²

- 580

Young’s modulus, RD, N/mm²

- 185 000

Young’s modulus, TD, N/mm²

- 200 000

Hardness HV5 (VPN)

- 220

**RD** represents the rolling direction
**TD** represents the transverse direction
Values for yield strength (0.2% proof strength) and tensile strength are given for the rolling direction
Values for the transverse direction are approximately 5% higher

June 2008

---

Power Loss Modeling of Isolated AC/DC converter
Heat Sink

Cycloconverter

For all isolated power modules

P 16

Features
- Intended for all isolated power modules: SEMIPACK, SEMITRON, SEMIFONT, SKIM, SEMIX, SKIP
- Excellent efficiency/volume ratio
- Best suited fans: SKF 16-A and SKF 16-B
- Available in various lengths

P16 general profile dimensions (w = 23.5 kg/m)

RC Circuit for SEMIPACK 1/2/3
Type RC48B or RC47B
(Check SEMIPACK for further details/specifications)

P 16 standard accessories
Power Loss Modeling of Isolated AC/DC converter

**VSC**

<table>
<thead>
<tr>
<th>Standard lengths</th>
<th>n</th>
<th>b/d Ø (mm)</th>
<th>( P_{\text{th} \text{IC}} ) (KWs)</th>
<th>( P_{\text{th} \text{IC}} ) with Fan (KWs)</th>
<th>w (kg)</th>
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<td>0,43 (150W)</td>
<td>0,147</td>
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<tr>
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<td>0,36 (180W)</td>
<td>0,12</td>
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<td></td>
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</tbody>
</table>

Heatsink

For isolated power modules

- Intended for isolated power modules, SEMIPACK (1 to 4) and SEMITRANS 2 range
- Integrated rails allow for easy mounting of the modules
- Available in various lengths
- Best fitted fan: SKF 3-230-01
- Mounting bar rails available (see sketches)

Dimensions in mm

P 3 general profile dimensions (w = 17,6 kg/m)

Protection grill + Fan SKF 3-230-01

RC network for SKKT/H 19 or 26

RC network for SEMIPACK 1/2/3

Mounting bars (to be inserted into the rails)

(Please contact SEMIKRON for further details on the above accessories)

P 3 standard accessories

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