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Rectifier Solutions for a Transverse Flux Permanent Magnet Machine

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Abstract

Due to their interesting characteristics including their high torque density, transverse flux permanent magnet machines are drawing attention, particularly as wind turbine generators. An uncontrolled bridge rectifier could not be used with a transverse flux generator due to its high stator reactance. The objective of this study is to examine the rectifier solutions for variable speed transverse flux generators, their advantages and their shortcomings. As conventional solutions, uncontrolled bridge-rectifier with passive compensation and PWM rectifiers are examined. Also, a new high efficiency rectifier topology is introduced which could be used with sources with variable frequency and high saturable source inductance including transverse flux machines.

Résumé

Un pont à diodes ne peut pas être utilisé avec une génératrice à flux transverse en raison de sa grande inductance statorique. Le but de cette étude est d'examiner différentes solutions pour des redresseurs utilisés avec une génératrice à flux transverse fonctionnant à vitesse variable en regardant leurs avantages et inconvénients. Comme solutions conventionnelles, des ponts à diodes avec compensation passive et des redresseurs à MLI sont étudiés. Aussi, une nouvelle topologie de redresseur à haut rendement sera présentée qui peut être utilisée avec des sources à fréquence variable ayant une inductance très élevée et saturable incluant les machines à flux transverse.
## List of Symbols and Abbreviations

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \rho )</td>
<td>air density</td>
</tr>
<tr>
<td>( v )</td>
<td>wind speed</td>
</tr>
<tr>
<td>( A )</td>
<td>rotor swept area</td>
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<tr>
<td>( C_p )</td>
<td>power coefficient</td>
</tr>
<tr>
<td>( \lambda )</td>
<td>tip-speed ratio</td>
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<tr>
<td>( \omega )</td>
<td>turbine angular velocity, radian frequency</td>
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<tr>
<td>( \phi )</td>
<td>phase angle between ( V_g ) and ( E_g )</td>
</tr>
<tr>
<td>( \delta )</td>
<td>power angle</td>
</tr>
<tr>
<td>( \theta_f )</td>
<td>firing angle</td>
</tr>
<tr>
<td>( \omega_g )</td>
<td>radian frequency of the source EMF</td>
</tr>
<tr>
<td>( \alpha )</td>
<td>damping factor</td>
</tr>
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<td>( \zeta )</td>
<td>damping ratio</td>
</tr>
<tr>
<td>( p(v) )</td>
<td>rayleigh probability density function</td>
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<tr>
<td>( v )</td>
<td>wind speed in m/s</td>
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<tr>
<td>( u_d )</td>
<td>long term average wind speed</td>
</tr>
<tr>
<td>( \eta )</td>
<td>efficiency</td>
</tr>
<tr>
<td>( \mu )</td>
<td>overlap angle</td>
</tr>
<tr>
<td>( \nu_{\text{cut-in}} )</td>
<td>“cut-in” speed</td>
</tr>
<tr>
<td>( \nu_{\text{cut-out}} )</td>
<td>“cut-out” speed</td>
</tr>
<tr>
<td>( V_{\text{CF}} )</td>
<td>forced response for the capacitor voltage</td>
</tr>
<tr>
<td>( V_{\text{Out}} )</td>
<td>output voltage</td>
</tr>
<tr>
<td>( P_w )</td>
<td>available wind power</td>
</tr>
<tr>
<td>( P_{\text{out}} )</td>
<td>power measured at the output terminals</td>
</tr>
<tr>
<td>( P_{\text{in}} )</td>
<td>power measured at the input terminals</td>
</tr>
<tr>
<td>( I_{\text{LgF}} )</td>
<td>forced response for the inductor current</td>
</tr>
<tr>
<td>( R )</td>
<td>radius of the turbine</td>
</tr>
<tr>
<td>( \nu_{\text{nom}} )</td>
<td>wind speed at which the turbine produces its nominal power</td>
</tr>
<tr>
<td>( L_g )</td>
<td>stator inductance</td>
</tr>
<tr>
<td>( R_g )</td>
<td>equivalent resistance of the stator</td>
</tr>
<tr>
<td>( C_{\text{Out}} )</td>
<td>rectifier output capacitor</td>
</tr>
<tr>
<td>( R_{\text{Load}} )</td>
<td>load resistance</td>
</tr>
<tr>
<td>( C_{\text{base}} )</td>
<td>base value for capacitance</td>
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<tr>
<td>( X_{\text{base}} )</td>
<td>base value for reactance</td>
</tr>
<tr>
<td>( E_{g,\text{nom}} )</td>
<td>nominal value of ( E_g )</td>
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</table>
peak value of the source EMF voltage,
short circuit current
nominal value for \( I_{g} \) are chosen as base values for voltage and current
output voltage at no-load
shunt capacitor
series capacitor
base value for power
turbine output power
frequency
output power
source reactance
maximum average on-state current,
on-state voltage
maximum collector DC current
maximum collector-emitter voltage
collector-emitter saturation voltage
instantaneous current through switch \( X \)
instantaneous voltage through switch \( X \)
on-state voltage drop on the switch \( X \) (IGCT or thyristor)
forward voltage on the diode \( X \)
instantaneous capacitor voltage
instantaneous source current.
nominal value of source RMS current
RMS values of the source EMF voltage
Equivalent Series Resistance of the capacitor
slope resistance of the conducting switch
source voltage
RMS value of the fundamental component of the source current.
peak voltage on the compensation capacitor.
thyrister turn off time
rectifier’s output voltage
rectifier’s output current
inductor current at the end of a half period
inductor current at the beginning of a half period
peak Repetitive off-state Forward Voltage
peak Repetitive off-state reverse Voltage
active power output from the source
rectifier losses
losses in the thyristor half-bridge
diode half-bridge
bidirectional switch losses
capacitor losses
losses in the forced commutation circuit
losses in the IGBTs
diode losses
threshold collector-emitter voltage of the IGBT
\( E_{\text{on}} \) turn-on energy
\( E_{\text{off}} \) turn-off energy
EMF electromotive force
SGCT symmetric Gate-Commutated Thyristors
BSW bidirectional Switch
FCC forced commutation circuit
ETO emitter turn-off thyristor
TFPM transverse flux permanent magnet
DCM discontinuous conduction mode
PMSG permanent magnet synchronous generator
DSP digital signal processor
CCM continuous conduction mode
DISF distortion factor
dPF displacement factor
IGCT gate commutated thyristor
PWM pulse width modulation
RB reverse blocking
RPR rapid polarity reversal
GTO gate turn-off thyristor
ESI equivalent Series Inductance
HB half-bridge
TCHD total current harmonic distortion
ESR equivalent series resistance
CSR current source rectifier
VSR voltage source rectifier
CSI current source inverter
VSI voltage source inverter
# Table of Contents

1. INTRODUCTION .......................................................................................................................... 9

2. EXISTING AC/DC CONVERTER SOLUTIONS FOR A TRANSVERSE FLUX PERMANENT MAGNET MACHINE .................................................................................................................... 13
   2.1 WIND ENERGY CONVERSION BASICS .................................................................................. 13
      2.1.1 The extractable Wind power ......................................................................................... 13
      2.1.2 Efficiency of the Turbine ............................................................................................ 14
   2.2 TRANSVERSE FLUX PERMANENT MAGNET MACHINE ......................................................... 16
   2.3 DIODE BRIDGE RECTIFIER ................................................................................................... 18
      2.3.1 Single-phase bridge with constant voltage load in the presence of high supply impedance ........................................................................................................... 19
      2.3.2 Diode bridge rectifier with constant current (inductive) load in the presence of high supply inductance .............................................................................. 24
   2.4 PASSIVE COMPENSATION ..................................................................................................... 28
      2.4.1 Shunt Compensated VSR ............................................................................................... 29
      2.4.2 Series Compensated VSR ............................................................................................. 36
      2.4.2.1 Series Capacitor Compensation Considering Saturation ........................................ 38
      2.4.4 Comparison of the shunt and series compensated VSR ................................................. 42
      2.4.4.1 Comparison of the shunt and series compensated VSR considering saturation .... 43
      2.4.5 Shunt and series compensated CSR .............................................................................. 44
   2.5 PWM RECTIFIER ................................................................................................................... 48
      2.5.1 PWM Voltage Source Rectifier (VSR) .......................................................... 48
      2.5.2 PWM Current Source Rectifier (PWM-CSR) ............................................................. 50
   2.6 THE RAPID POLARITY-REVERSAL MIXED-BRIDGE RECTIFIER ........................................ 51

DISCUSSION ..................................................................................................................................... 52

3. THE RAPID POLARITY REVERSAL (RPR) MIXED-BRIDGE RECTIFIER .................................... 54
   3.1 GENERAL DESCRIPTION ....................................................................................................... 54
   3.2 THE CHOICE OF SWITCHES .................................................................................................. 56
      3.2.1 Switches for the Controlled Half-Bridge ........................................................................ 56
      3.2.2 Bidirectional Switch ...................................................................................................... 60
   3.3 ADVANTAGES OF THE PROPOSED TOPOLOGY ................................................................. 62
   3.4 CIRCUIT ANALYSIS .............................................................................................................. 62
      3.4.1 Forced commutation circuit (FCC) ................................................................................ 63
      3.4.2 Detailed Analysis ........................................................................................................... 64
      3.4.3 Analysis of the Resonant Circuit .................................................................................. 68
   3.5 RATING ANALYSIS .............................................................................................................. 74
      3.5.1 Thyristor and Diode Half-Bridges (T1,T2,D1,D2) ......................................................... 74
      3.5.2 Bi-directional Switch (T3, T4) ....................................................................................... 77
      3.5.3 Capacitor ......................................................................................................................... 77
   3.6 ANALYSIS OF THE CONTROL STRATEGY FOR THE RPR-CONVERTER ...................... 80
      3.6.1 Gate Control circuit for the Bidirectional Switch ......................................................... 80
      3.6.2 Gate Control for the controlled Half-Bridge ................................................................. 82
   3.7 PERFORMANCE OF THE RPR-CONVERTER ....................................................................... 82
Chapter 1

Introduction

Though the increasing concerns over environmental impact of consuming non-renewable sources of energy could explain the rapidly growing interest in renewable energy, another reason behind such a trend should be sought in the increasing prices of the rapidly exhausting fossil fuels as the main source of energy in the world. Wind energy as a huge, unlimited, free and non-polluting source of energy could actually be considered as the most attractive alternative among renewable energy sources, among several factors, due to rapid advances in the generator and converter technologies.

A brief history of wind energy utilization
The first known use of wind energy has been boat sailing. The first documented indications of windmills and their use date from 10th century, in Persia [1]. Wind has been first used in the 19th century for generating electricity. By 1910 there were already several hundred units in Denmark with 5-25 kW capacities [2].

In the United States, already by 1930, there were a number of large-scale turbines in operation including a 1.2 MW machine which used a turbine of 53 m diameter. The increase of oil prices, which led to the energy crisis of 70's, attracted a lot of attention to wind energy and resulted in a rapid development in this field. During the 70's and 80's a series of large wind turbines were built. One of these (MOD-5A), had a power output of 7.3 MW and a diameter of 122 meters [2]. With the new technologies involved and market development, the cost of electricity produced from wind turbines has considerably
reduced in recent years and it is becoming more competitive as prices of fossil fuels are increasing.

**Turbines with gearbox vs. direct-drive concept**

The modern wind turbines could be divided into two categories:

1) Those with a gearbox between the turbine hub and the generator. Until the late 1990s these were mainly constant-speed wind turbines, however, nowadays most wind turbine manufacturers have changed to variable speed for power levels between 1.5 to 5 MW [4].

2) The direct-drive concept where the turbine hub is directly connected to the generator (fig. 1-1). The main advantages of the direct-drive concept could be the fact that it eliminates the lossy gearbox and its associated costs, including maintenance. The direct-drive could be also considered as a more reliable system.

![Diagram of Direct-drive grid connected wind energy conversion system]

**Generator Types**

In the first group the combination of the of gearbox and the doubly-fed induction generator offers the best performance [3,4]. In the direct drive group, however, two types of generators are mainly employed:

1. Synchronous Wound-rotor DC-excited Generators
2. Permanent Magnet Synchronous Generators (PMSGs)
Most direct drive turbines currently on the market are of the first type, however, according to [5] the direct drive systems using a wound-rotor synchronous machine cost more than the geared system. PM generators avoid the excitation losses in the wound rotor and do not need brushes or slip rings which means that they a very attractive choice for wind turbine generators. Besides, they also allow a smaller pole pitch which may possibly reduce the copper losses further [6]. However, even with PM machines, the cost of a direct drive system still remains a problem.

Among different topologies of PMSG, Transverse Flux Permanent Magnet (TFPM) machine has drawn much attention since it could, under certain conditions, substantially reduce the size of the generator and lower the cost of the machine active material for a given torque value compared to more conventional PM machine topologies [16].

**Outline of the thesis and the investigation approach**

The TFPM machine will be briefly discussed in chapter 2. Due to high values of stator inductance, the design of AC/DC converter for such a machine will not be a straightforward task. As solution for a variable speed machine with high stator reactance, PWM rectifiers are currently considered, however, they are expensive and involve high losses. Reducing rectifier losses and price will be the central objective of the current thesis. This thesis will investigate the existing rectifier solutions for TFPM machine and also examine the performance of a new alternative rectifier topology. In chapter 2, the uncontrolled bridge rectifiers for both constant voltage and current loads will be first considered and their limitations will be investigated using simulation in Matlab-Simulink. Passive compensation as the simplest solution to high stator inductance will be also discussed in chapter 2. Numerous simulation models will be used to compare series and shunt compensation and the effect of saturation in the variable speed machine. PWM rectifiers, as a second conventional solution to high source inductance, will be also discussed in chapter 2. In chapter 3 a new AC/DC converter topology (the rapid polarity reversal mixed bridge rectifier) will be introduced which could be considered as a high-efficiency alternative to these topologies. Special attention will be paid to the choice of
the switches for the converter and a detailed analysis of the converter will be presented. Analytical formulas will be developed which will be later used for numeric calculation. A comprehensive rating analysis for all converter components will follow. The capability of the converter to extract power despite the high source inductance will be also analyzed. A detailed loss analysis will be conducted and analytical formulas will be developed for all converter components. To investigate the efficiency of the converter, the losses will be numerically calculated in each component. To examine the feasibility of employing such a converter in direct-drive wind turbines, the real power curves for a MW range turbine, will be used along with numerical analysis to establish the efficiency-speed curves and the distribution of losses in converter components. The losses will then be compared to those of a two-level Sine-PWM rectifier obtained with PCIM simulator with identical turbine power curves. To see if it is economically justified to employ the new topology for a MW range wind turbine with TFPM generator, the annual energy production of the turbine for both converter options and their respective semiconductor prices will be compared.

Beside numerical calculation, the performance of the new topology will also be investigated using simulation Simulink. Finally, an experimental model of the proposed topology will be presented which uses a DSP controller. The converter waveforms along with efficiency measurements for the model will be also examined.
Chapter 2

Existing AC/DC Converter Solutions for a Transverse Flux Permanent Magnet Machine

The objective of this chapter will be to discuss the existing AC/DC converter solutions for a transverse flux permanent magnet machine. It will take a brief look into the machine itself in section 2.2. Section 2.3 examines the feasibility of employing uncontrolled rectifiers with a TFPM machine represented by its equivalent circuit. Passive compensation as the simplest way to overcome the limitations of a diode rectifier will be the subject of section 2.4. Simulation models will be used to compare the series and shunt compensation and show their advantages and disadvantages. PWM rectifiers will be reviewed in section 2.5. Before all, however, we will briefly address the basic concepts of wind energy conversion in section 2.1. This will be limited only to those necessary for our analysis in chapters 2 and 3.

2.1 Wind Energy Conversion Basics

2.1.1 The extractable Wind power
The available wind power can be expressed as:

\[ P_w = \frac{1}{2} \rho A v^3 \]  

(2.1)

Where \( \rho \) is air density, \( v \) is the wind speed, and \( A \) is the rotor swept area.
All of this power is not extractable, the actual captured value is:

\[ P = \frac{1}{2} C_p \rho A \nu^3 \]  

(2.2)

\( C_p \) is known as power coefficient and has a theoretical limit of 0.593.

2.1.2 Efficiency of the Turbine

The power coefficient \((C_p)\) versus tip-speed ratio \((\lambda)\) for various types of wind turbines is shown in Fig. 2-1. Tip speed ratio is a physical characteristic of the wind turbine and is defined as \(\lambda = \omega R/\nu\), where R is the radius of the turbine, and \(\omega\) is the turbine angular velocity.

![Diagram showing power coefficient \((C_p)\) versus tip-speed ratio \((\lambda)\) for various types of turbines.](image)

Fig. 2-1 The power coefficient \((C_p)\) versus tip-speed ratio \((\lambda)\) for various types of turbines [7].

The power curve for a 3 MW wind turbine is shown in Fig. 2-2.
Below “cut in” speed \( v_{\text{cut-in}} \) the generator does not produce any power. As the wind speed increases to a value higher than \( v_{\text{cut-in}} \), the output power increases until the rotor reaches its nominal value. In this interval, \( \lambda \) and as a result \( C_p \), will be kept at their optimal values by controlling the reaction torque of the generator and the wind turbine will have its highest efficiency. At a certain point \( (10 \text{m/s}) \), as the wind speed increases, rotor speed will be kept constant. As a consequence, both \( \lambda \) and \( C_p \) will decrease until wind reaches a speed \( (v_{\text{nom}}) \) at which the turbine produces its nominal power. Beyond \( v_{\text{nom}} \) the turbine control systems limit the power output to a constant value through controlling the reaction torque of the generator or through mechanical control like pitch angle adjustment as shown above. The “cut-out” speed (not shown in the figure) represents the wind speed at which the generator shuts down to protect the blades from possible damage. The turbine power curve of Fig. 2-2 will be used later in chapter 3 for efficiency comparison among different topologies in the MW range.
2.2 Transverse Flux Permanent Magnet Machine

The concept of the transverse flux machines has its origins in the beginnings of the 20th century, however it was Weh [9] who first developed the first prototypes of machines based on the transverse flux principle in 1980s [18]. Since then, it has constantly been the focus of attention due to its interesting characteristics as briefly mentioned in the previous chapter.

Transverse flux machines may find their application where high torque and low to medium rotational speeds are required [21]. In the Transverse Flux Permanent Magnet (TFPM) machines the magnetic flux flows in a direction perpendicular to the direction of rotation and their force density is strongly dependent on pole pitch and air gap thickness [20]. TFPM machines are generally divided in two main families. Those with surface mounted permanent magnets (flat magnets rotor topology) as shown in Fig. 2-2R (a) or with permanent magnets inserted in between the flux concentrating poles (concentrating flux rotor topology) as shown in Fig. 2-2R (b) [14]. The stator may be single sided or double sided. The double sided machine may have single or double stator winding. Various core geometries are also possible. A more detailed discussion could be found in [25].

Fig. 2-2R: Single sided TF machine a) with flat permanent magnets rotor and b) with concentrating flux rotor [14]
Some of the advantages of this topology are as follows:

(i) The pole pitch could be made very small.
(ii) Much higher force densities are possible compared to conventional synchronous permanent magnet machines [16].
(iii) Very simple armature coils are employed and the total conductor length is relatively short [15] which results in lower copper losses.
(iv) The phases in a TFM are magnetically independent and this decoupling in structure simplifies the control method [15].
(v) Since stator winding has a simple mechanical structure, higher levels of isolation could be achieved [17].

The main disadvantages of TFMs are:

(vi) Low power factor due to high values of stator inductance.
(vii) The complexity of construction due to the high accuracies required and complex structure of the machine.

Other problems encountered are torque ripples and fluctuations in the normal force value. They can be reduced by modification of the geometry of the magnetic path and by the application of an appropriate current waveform to the machine [19]. Finding the appropriate AC/DC converters to overcome the problems resulting from the high stator inductance will be the main concern of this work in chapters two and three.

The simplified equivalent circuit of the machine is shown in Fig. 2-3, where \( E_g \) is the RMS value of the machine EMF voltage, \( L_g \) is the stator inductance, and \( R_g \) is the equivalent resistance of the stator.
Under nominal conditions, current in a TFPM machine will be sufficiently high to reach core saturation [25]. This will be an important consideration when looking for an AC/DC converter for the TFPM generator. The saliency effect on the stator inductance is neglected in the rest of the work and the EMF voltage will be assumed to be sinusoidal. While these assumptions could not be applied to all TFPM machine topologies, they could be considered as reasonable based on what is found in literature on TFPM machine [25].

2.3 Diode Bridge Rectifier

An uncontrolled or plain diode rectifier is the most commonly used circuit in power electronics. It is cheap, requires no control, and has low losses. Diode rectifiers used as AC/DC converters in a wind energy conversion system often require a smooth current or voltage in the dc-link to feed a Current Source Inverter (CSI), a Voltage Source Inverter (VSI) or a DC-DC converter. That is why the analysis of rectifiers with constant current load and constant voltage load has been included in section 2.3.1 and 2.3.2.
TFPM machine is basically a single-phase machine, so this work will consider the single-phase rectifiers only. However, it is possible to build multi-phase machines by using independent single-phase units sharing the same shaft.

2.3.1 Single-phase bridge with constant voltage load in the presence of high supply impedance

The single-phase bridge with constant voltage load (also called a Voltage Source Rectifier (VSR)) is the most widely used rectifier circuit in power supplies. However, the case of high source inductance is not such a common situation and needs special attention. The term constant voltage load (also referred to as capacitive load) implies that the load voltage should be almost constant within an ac cycle (but it may vary over time). A very large capacitor will be considered as an approximation to constant load voltage. Depending on the source inductance and the output current of the rectifier, current in the inductor may be continuous or discontinuous.

Discontinuous Conduction Mode (DCM)

Fig. 2-5 shows the simulation waveforms for the bridge rectifier of Fig. 2-4 for a source reactance of 2.04 pu. The load resistance has been chosen so that the rectifier operates in the discontinuous conduction mode. Ideal diodes were assumed and the following parameters were used:
\[ E_{g,\text{nom}} = 60 \ \text{V}_{\text{rms}} \ @ \ f_{\text{nom}} = 50\text{Hz} \]

\[ I_{\text{nom}} = 2 \ \text{I}_{\text{SC}} = 15 \ \text{A} \]

\[ X_g = 8.168 \ \Omega \ @ 50\text{Hz} = 2.04 \ \text{pu (unsaturable)} \]

\[ R_g = 0 \ \Omega \]

\[ C_{\text{Out}} = 25.7 \ \text{pu} \]

\[ R_{\text{Load}} = 20 \ \text{pu} \]

where \( I_{\text{nom}} \) represents the nominal current of the source, \( f_{\text{nom}} \) is the nominal frequency, \( X_g \) is the source reactance, and \( \text{I}_{\text{SC}} \) is the short circuit current of the source. \( C_{\text{Out}} \) is the rectifier output capacitor and \( R_{\text{Load}} \) is the load resistance. The base value for capacitor (\( C_{\text{base}} \)) was chosen to be as \( C_{\text{base}} = \frac{(2\pi f_n)}{L_g} \). The base value for reactance was chosen to be \( X_{g,\text{base}} = \frac{E_{g,\text{nom}}}{I_{\text{nom}}} \), where \( E_{g,\text{nom}} \) is the nominal value of \( E_g \).

Fig. 2-5: The waveforms for the (a) source EMF voltage, (b) source current and (c) the input voltage of the rectifier of Fig. 2-3 in DCM
Considering the positive half-cycle of the source EMF voltage, diodes D1 and D2 begin to conduct at $\theta_1$ when $v_g$ exceeds the output voltage $v_{Out}$. We have:

$$\sin(\theta_1) = \frac{v_{Out}}{E} \quad (2-3)$$

where $E$ is the peak value of the source EMF voltage.

In the absence of source reactance, the source current will become zero as $v_g$ gets smaller than $V_{Out}$ and the inductor current will be highly discontinuous, however, due to the presence of the source reactance, D1 and D2 will continue to conduct until the current falls to zero at $\theta_2$ and both diodes will cease to conduct. This mode of operation will be referred to as DCM and will prevail as far as the condition $\theta_2 - \theta_1 < \pi$ applies.

**Boundary Condition**

As the inductor current increases, $\theta_2$ will also increase until we have a condition where $\theta_2 - \theta_1 = \pi$ and the inductor current becomes continuous. As one diode conducts in an arm, the other one in the same arm will be reverse biased by the output voltage. As the polarity of source current changes, an instant commutation between the two diodes will occur and the input voltage will also change polarity. This means that input voltage ($v_g$) and the source current will be in phase. The input voltage ($v_g$) and the source current ($i_{L,g}$) can be approximated by their fundamental components $v_{g1}$ and $i_{L,g1}$ [22]:

$$v_g \approx v_{g1} = \frac{4V_{Out}}{\pi} \sin(\omega t - \delta) \quad (2-4)$$

$$i_{L,g} \approx i_{L,g1} = \sqrt{2} I_{L,g1} \sin(\omega t - \delta) \quad (2-5)$$

where $\delta$ is the phase angle between $V_g$ and $E_g$ and $I_{L,g1}$ is the RMS value of the fundamental component of the source current.

The phasor diagram has been drawn in Fig. 2-6.
We have:

$$\cos(\delta) = \frac{V_g}{E_g} = \frac{4}{\pi} \times \frac{V_{\text{out}}}{E} \quad (2-6)$$

Since we have $\theta_i \approx \delta$, the equations (2-3), (2-4), and (2-6) give the boundary condition for CCM as being:

$$\tan(\delta) \approx \frac{\pi}{4} \text{ or } \delta \approx 38^\circ \quad (2-7)$$

**Continuous Conduction Mode (CCM)**

Beyond this limit $\delta$ will increase (active power will decrease), however, the rectifier will remain in the continuous conduction mode and equations (2-4), (2-5), and (2-6) will remain valid. For the fundamental of the source current we could write:

$$I_{Lg1} = \sqrt{E_g^2 - V_g^2} = \sqrt{E^2 - \left(\frac{4V_{\text{out}}}{\pi}\right)^2} \quad \frac{X_g}{\sqrt{2}} \quad (2-8)$$

The analytical approach will become much more complicated in the DCM and therefore for the analysis in the rest of this chapter, simulation will be used.

In order to give a better idea of the limitations of a diode rectifier for a variable speed wind turbine-generator, the diode rectifier in Fig. 2-4, has been simulated with a source
of variable frequency (0-50Hz). The source EMF voltage was assumed to be proportional to the source frequency. Fig. 2-7 shows the power output versus load resistance and source EMF voltage for $X_g=2$ pu. The source EMF voltage has been chosen as base value for the voltage and $I_{nom}=2I_{SC}$ is chosen as the base value for the current where

$$I_{SC} = \frac{E_{g,nom}}{X_g}.$$  

![Diagram](image-url)

Fig. 2-7: Power output versus load resistance and source EMF voltage for $X_g=2$ pu

$P_{Out,max}$ is in the (y, z) plane (i.e. $E_g=1$ pu) and will be equal to 0.2 pu for a load resistance of 1.95 pu (corresponding to a source current of 0.36 pu). This clearly shows the limitation of a plain diode-rectifier with constant voltage load to extract the source power.
2.3.2 Diode bridge rectifier with constant current (inductive) load in the presence of high supply inductance

Unlike the case with constant load voltage, analysis of the bridge rectifiers with constant load current in the presence of source inductance could be found in most texts (among others [10], [11], and [12]). The term constant current load (also referred to as inductive load) implies that the load current should be almost constant within an ac cycle (but it may vary over time). The current in a large inductance could be considered a good approximation to constant current. Due to the presence of source inductance, the commutations in the rectifier will not be instantaneous and will require a finite time which will be represented by commutation or overlap angle (\(\mu\)).

![Single phase diode bridge rectifier with constant current load](image)

Fig. 2-8: Single phase diode bridge rectifier with constant current load

For a single phase bridge (Fig. 2-8) the commutation angle could be calculated as follows [10]:

\[
\cos \mu = 1 - \frac{\sqrt{2} \omega L_s}{E_{g,\text{nom}}} I_{\text{Out}}
\]

(2-9)

where \(\omega\) is the angular frequency of the source.

As \(I_{\text{Out}}\) increases, the commutation angle will also increase. Fig. 2-9 shows the output voltage and the input current waveforms for \(X_g=0.5\) pu.
Fig. 2-9: Output Voltage ($v_{Out}$) and input current ($i_L$) waveforms for the rectifier in Fig. 2-8, $X_g=0.5$ pu

We could rewrite Eq. (2-9) as follows:

$$\cos \mu = 1 - \frac{\sqrt{2} I_{Out} X_{g, pu}}{I_{nom}}$$  \hspace{1cm} (2.10)

Fig. 2-10 shows the commutation angle versus $I_{Out} / I_{nom}$ for $X_g=0.5$, 1, and 2 pu. A commutation angle equal to $180^\circ$ means that the source will be short circuited during the whole cycle.

The output voltage can be calculated from [10]:

$$V_{Out} = V_{Out(0)} - \frac{2\omega L_s I_{Out}}{\pi}$$  \hspace{1cm} (2.11)

to:

$$V_{Out(0)} = \frac{2\sqrt{2}}{\pi} E_g$$  \hspace{1cm} (2.12)
In the above calculation the voltage drop on the diodes is neglected.

Eq. (2-11) could be rewritten as:

\[ V_{Out} = \frac{2\sqrt{2}}{\pi} E_g \left( 1 - \frac{I_{Out} X_{g,pu}}{\sqrt{2} I_{nom}} \right) \]  

(2.13)

The output voltage versus normalized output current \( (I_{Out} / I_{nom}) \) for \( X_g = 0.5, 1, \) and \( 2 \) pu are shown in Fig. 2-11 where \( V_{Out(0)} \) has been chosen as base value for the output voltage.

The output power of the rectifier will be:

\[ P_{Out} = I_{Out} V_{Out} \]

\[ P_{Out} = P_{Out(L_g=0)} \left( 1 - \frac{I_{Out} X_{g,pu}}{\sqrt{2} I_{nom}} \right) \]  

(2.14)

where \( P_{Out(L_g=0)} = \frac{2\sqrt{2}}{\pi} E_g I_{Out} \).

\( P_{Out(L_g=0)} \) represents the output power of an ideal diode rectifier with no source inductance, so:

\[ \frac{P_{Out}}{P_{Out(L_g=0)}} = \frac{V_{Out}}{V_{Out(0)}} \]  

(2.15)

and its characteristic versus output current will be the same as Fig. 2-12.

In order to give a better idea of the limitations of a diode rectifier as an AC/DC converter for a variable speed wind turbine-generator, the diode rectifier in Fig. 2-8 has been simulated with a source of variable frequency (0-50Hz). The source voltage \( (E_g) \) was assumed to be proportional to the source frequency. Fig. 2-12 shows the power output versus source current and voltage for \( X_g = 2 \) pu (@ 50 Hz).
Fig. 2-10: The commutation angle versus normalized output current ($I_{\text{Out}}/I_{\text{nom}}$) for $X_g=0.5, 1, \text{and } 2 \text{ pu.}$

Fig. 2-11: Output voltage versus normalized output current ($I_{\text{Out}}/I_{\text{nom}}$) for $X_g=0.5, 1, \text{and } 2 \text{ pu.}$
Fig. 2-12: Output power versus source current and voltage for $X_g=2$ pu

$P_{\text{Out, max}}$ is in the $(y, z)$ plane (i.e. $E_g=1$ pu) and will be equal to 0.159 pu for a current of 0.354 pu. As could be seen, like the uncontrolled VSR bridge, the CSR bridge will seriously limit the output power of rectifier. That is why in the next section we will investigate the passive compensation.

### 2.4 Passive Compensation

Passive compensation is the simplest solution for a source with high source reactance. In this section the advantages and disadvantages of such a method will be discussed.
2.4.1 Shunt Compensated VSR

For the purpose of analysis, the circuit in Fig. 2-13 will be considered.

Initially, we suppose that diodes D1 and D2 are conducting. If the output capacitor value is large enough, the voltage on the input capacitor ($C_{sh}$) will be fixed by the output voltage ($V_{out}$). As the source emf voltage gets smaller than $V_{out}$, the source current will decrease until it falls to zero and the diodes D1/D2 will be blocked. Current in the inductor will change sign and capacitor $C_{sh}$ will begin to discharge through the inductor. As the capacitor voltage ($v_g$) gets more negative than $-V_{out}$, D3 and D4 begin to conduct and the capacitor’s voltage will be fixed by $-V_{out}$.

Simulation was used to find the operating conditions at which maximum power could be extracted from the source at nominal frequency for the circuit of Fig. 2-13. The following parameters were selected:

$E_{g,nom}=60$Vrms @ $f_{nom}=50$Hz

$I_{nom}=2I_{SC} =15$ A

$X_g=8.168$ $\Omega$ @50Hz =2.04 pu (unsaturable)

$R_g=0.4$ $\Omega=0.1$ pu

$C_{base}=389.7$ $\mu$F

$C_{out}=25.7$ pu
$E_{g_{nom}}$ and $I_{nom}$ were used as the base values for voltage and current respectively. The base values for power ($P_{base}$) and for the compensation capacitor ($C_{base}$) were chosen as follows:

$$P_{base} = (E_{g_{nom}} I_{nom} - R_g I_{nom}^2)$$

$$C_{base} = \frac{(2\pi f_{nom})}{L_g}$$

The parameters for the diode rectifier were chosen to be typical datasheet values.

For the analysis in section 4.2, the displacement factor ($\cos(\varphi)$) will be defined as that of the sinusoidal source $E_g$, i.e. $DPF = \cos(\varphi) = \frac{P_g}{E_g I_e (rms)}$ and distortion factor (DISF) will be defined as $DISF = \frac{I_{L_g (rms)}}{I_{L_e (rms)}}$, where $P_g$ is the active power extracted from the source.

The capacitance $C_{Sh}$ varied from 0.1 to 1.2 pu with 0.1 pu steps and load resistance varied between 0.5 to 7 pu with 0.5 pu steps. The results for output power are shown in Fig. 2-14. The results included are only those for which the RMS value of the source current does not exceed $I_{nom} = 2I_{SC}$.

As could be seen, the peak power occurs at higher values of load resistance as $C_{Sh}$ increases. For the parameters chosen, a capacitor value equal to 0.7 pu will give the highest output power. Fig. 2-15 shows the waveforms for the source EMF voltage, source current, and the input voltage of the rectifier in such a condition. It should be noted that the peak power value will be an approximation whose precision depends on the step size chosen for the $C_{Sh}$ and $R_{Load}$.

To find out the output power in applications where we have a variable EMF voltage and frequency (as in the case of a variable speed wind generator), the capacitor was chosen to be 0.7 pu and the EMF voltage and frequency were varied simultaneously between 0.3 and 1.4 pu. Load resistance varied between 0.5 to 7 pu. The results for output power are partially shown in Fig. 2-16. As before, the results included are only those for which the RMS value of the source current does not exceed $I_{nom} = 2I_{SC}$. 
Fig. 2-14: 2D representation of the simulation results for the shunt compensated VSR at nominal frequency, $X_g=2.04$ pu @50Hz

Fig. 2-15: The waveforms for the (a) source EMF voltage, (b) source current and (c) the input voltage of the rectifier for the 0.7 pu shunt capacitor.
Again, the 1 pu source EMF and frequency will give the highest power and the peak power will shift to higher values of $R_{\text{Load}}$ as $E_g$ increases.

To see the consequences of the compensation for a variable speed wind turbine-generator, the peak extractable power at each speed (frequency-EMF voltage) is presented in Fig. 2-17 (line).

Fig. 2-17: Maximum extractable output power versus source frequency (line) for unsaturable source inductance and a frequency proportional to wind speed ($C_{\text{Sh}}=0.7$ pu), the optimal turbine power curve (dotted).
The second curve (dotted) shows the optimal turbine power ($P_{\text{Turbine}}$) curve, for which the following assumptions were made:

\[ C_p = \text{constant} \]

\[ E_s \propto f \] \hspace{1cm} (2.16)

So we have:

\[ \omega \propto v \] \hspace{1cm} (2.17)

\[ P_{\text{Turbine}} \propto v^3 \propto f^3 \] \hspace{1cm} (2.18)

It is also assumed that the turbine is dimensioned so that at nominal speed, $P_{\text{Turbine}}$ is equal to maximum extractable output power.

Comparing the two curves it could be seen that the peak extractable output power, at each turbine speed, will be always greater than the turbine optimal power. This means that despite the fixed capacitor chosen, it is still possible to adjust the load current at each speed so that maximal power could be extracted from the wind.

**2.4.1.1 Shunt Compensated VSR Considering Saturation**

In order to show the effect of the saturation, the unsaturable inductor was replaced with a saturable inductor model which has the same unsaturable inductance (2.04 pu), however, its inductance falls to 0.627 pu at the nominal current. Similar to the previous case, the output power was measured while varying $C_{\text{Sh}}$ and $R_{\text{Load}}$ at nominal source EMF voltage and frequency. For source currents smaller or equal to $I_{\text{nom}}$, $C_{\text{Sh}}=0.6$ pu was found to give the highest output power. Fig. 2-18 shows the source current waveform at which maximum power could be extracted with the saturable inductor (Fig. 2-18 (c)) compared to that of the unsaturable inductor (Fig. 2-18 (b)).
Fig. 2-18: (a) The source EMF Voltage, source current (b) for the unsaturable inductor with $C_{sh}=0.7$ pu, (c) for the saturable inductor with $C_{sh}=0.6$ pu (c).

The displacement factor ($\cos(\phi)$) was found to be equal to unity for both cases (with $C_{sh}=0.675$ pu for unsaturable inductor and $C_{sh}=0.625$ pu for saturable inductor), however, the output power at nominal speed was approximately 3 percent lower when the saturable inductor was used. This could be explained by the higher current distortion due to saturation and slightly reduced efficiency. Again, the simulation with a variable speed (frequency-EMF voltage) and variable load was used to find the operating conditions at which peak power could be delivered at each frequency. DPF ($\cos(\phi)$) and distortion factor characteristics versus source frequency (source EMF) are shown in Fig. 2-19. It is interesting to see that distortion will be small during the whole range, though it will be considerably lower, at low speeds, due to the fact that a lower load current is required (saturation has a less pronounced effect) and the rectifier will still remain in the continuous conduction mode. The displacement factor, on the contrary, will decrease, as expected, at low speeds due to overcompensation which means increased reactive power generation at low speeds. However, as could be seen in Fig. 2-20, the available wind power will still be less than the power which could be delivered by the rectifier.
Fig. 2-19: \( \cos (\phi) \) (DPF) and distortion factor (DISF) characteristics versus source frequency (EMF voltage) for shunt compensated VSR with saturable source inductance.

Fig. 2-20: Maximum extractable output power versus source frequency (line) for saturable source inductance and a frequency proportional to wind speed \( (C_{Sh}=0.625 \text{ pu}) \), the optimal turbine power curve (dotted).
2.4.2 Series Compensated VSR

As a second option, we may compensate the large stator reactance by a series capacitor. To see the effects of series compensation, similar to the shunt compensation case, simulation was used to find the operating conditions at which maximum power could be extracted with series compensated rectifier of Fig. 2-21. The circuit parameters were similar to the previous case.

Fig. 2-21: Series compensated diode-bridge rectifier with constant voltage load

The capacitance \( C_s \) varied from 0.6 to 1.2 pu and load resistance varied between 0.5 to 10 pu. The results for output power are shown in Fig. 2-22. The results included are only those for which the RMS value of the source current does not exceed \( I_{\text{nom}} = 2I_{SC} \).

Fig. 2-22: Simulation results for the series compensated VSR.
As could be seen, the maximum power point shifts to lower values of load resistance as $C_s$ is increased and the maximum output power in each curve is less sensitive to the variation of the capacitance compared to the shunt capacitor case. A capacitor value equal to 1 pu will give the highest output power at nominal source current ($I_{Lg}=1$ pu). Fig. 2-23 shows the waveforms for the source EMF voltage, source current and the input voltage of the rectifier in such a condition.

For a varying EMF voltage and frequency (representing a variable speed machine), the capacitor was chosen to be 1 pu and the EMF voltage and frequency were varied simultaneously between 0.4 and 1.2 pu. Load resistance varied between 0.5 to 7 pu. The results for output power are shown in Fig. 2-24. As before, the results included are only those for which the RMS value of the source current does not exceed $I_{nom}=2I_{Sc}$.

![Waveforms](image.png)

Fig. 2-23: The waveforms for the (a) source EMF voltage, (b) source current and (c) the input voltage of the rectifier for the 1 pu series capacitor.
Again, the 1 pu source EMF and frequency will give the highest power and the peak power will shift to lower values of $R_{load}$ as machine speed increases.

The peak extractable power at each frequency (EMF voltage) is presented in Fig. 2-25 (line) along with optimal turbine power curve (dotted). As could be seen, while the VSR with series compensation could compensate the source reactance at nominal conditions, it could not extract the optimal power from the wind at turbine speeds below 0.9 pu.

![Fig. 2.24: Simulation results for the series compensated VSR, $C_s=1$ pu](image)

**2.4.2.1 Series Capacitor Compensation Considering Saturation**

In order to show the effect of the saturation, the unsaturable inductor was replaced with the saturable inductor model used in section 2.4.1.1. The output power was measured while varying $C_s$ and $R_{load}$ at nominal source EMF voltage-frequency. For source currents smaller or equal to $I_{nom}$, $C_s=1.325$ pu was found to give the highest output power. Fig. 2-26 shows the source current waveform for nominal conditions at which maximum power could be extracted from the source with saturable inductor (Fig. 2-26(c)) compared to that of the unsaturable inductor (Fig. 2-26(b)).
Fig. 2-25: Maximum extractable output power versus source frequency for unsaturable source inductance and a frequency proportional to wind speed (line), the optimal turbine power curve (dotted), $C_S=0.975$ pu.

Fig. 2-26: (a) The source EMF Voltage and source current for the unsaturable inductor with $C_S=0.975$ pu (b) and for the saturable inductor with $C_S=1.32$ (c), $E_g=1$ pu, $f_g=1$ pu.
Again, the simulation with a variable speed (frequency-EMF voltage) and variable load was used to find the peak power at each frequency. Examining the source current waveform for peak power at various speeds shows that, while the rectifier will be in continuous conduction mode at nominal conditions (Fig. 2-26(c)), at 0.3 pu speed, it will operate in discontinuous conduction mode as shown in Fig. 2-27 (b).

![Graph](#)

Fig. 2-27: (a) The source EMF Voltage and (b) source current for the series compensated VSR with saturable inductor, $C_S=1.32$ pu, $E_g=0.3$ pu.

DPF ($\cos(\phi)$) and distortion factor (DISF) characteristics versus source frequency (EMF voltage) are shown in Fig. 2-28. It is interesting to see that distortion factor will increase below 0.9 pu speed as the effect of saturation decreases, however, it will begin to decrease below 0.6 pu speed as the rectifier enters into the discontinuous conduction mode. Since for the two cases (saturable and unsaturable inductances), the optimal capacitors were used to compensate the effective inductance at nominal conditions, the
displacement factor ($\cos(\varphi)$) will be equal to unity for both cases. However, as speed decreases, $\cos(\varphi)$ will be higher for the saturable inductance since the effective value of the saturable inductance will increase and the larger selected capacitor will become a better choice for the lower frequencies.

Fig. 2-28: $\cos(\varphi)$ (DPF) and distortion factor (DISF) characteristics versus source frequency (EMF voltage) for series compensated VSR with saturable source inductance

Despite the fact that, saturation will improve the power transfer capability of the series compensated rectifier as the speed falls below its nominal value, (as could be seen in Fig. 2-29, where the peak extractable power at each frequency (EMF voltage) is presented along with the data from curve 2-25.), it could not still deliver the optimal power from the wind turbine.
Fig. 2-29: Maximum extractable output power versus source frequency for saturable source inductance (line, $C_S=1.325$ pu), unsaturable (line-dotted, $C_S=0.975$ pu), the optimal turbine power curve (dotted, calculated for the unsaturable source inductance)

### 2.4.4 Comparison of the shunt and series compensated VSR

As already shown in sections 2.4.1 and 2.4.2, the series compensation will supply a considerably smaller output power in the whole speed range. Table 2-1 summarizes the operating conditions for both shunt and series compensation circuits at their maximum output power where $\eta = \frac{P_{out}}{P_{in}}$ represents the efficiency, $P_{out}$ and $P_{in}$ are the converter output and input power respectively, $I_{C,rms}$ is the RMS current in the compensation capacitor, $C_{Out}$ stands for the output capacitor, and $V_{C,max}$ is the peak voltage on the compensation capacitor. For a better precision, much smaller step sizes has been chosen for $C_{Sh}$, $C_S$ and $R_{Load}$ compared to the previous simulations.
In the case of the shunt capacitor, output voltages are higher and the current in the diode rectifier is lower compared to the series compensation. This could explain much higher efficiencies in the former case. For the purpose of efficiency calculation, the capacitor ESR (Equivalent Series Resistance) has been neglected. While capacitor ESR might not affect the efficiency if the more expensive film type capacitors are used, the general purpose filter capacitors have much higher ESR values which means that the efficiency will further deteriorate due to higher capacitor currents in the series compensation.

The size, and as a consequence, the price of a capacitor, are a function of $CV$ rather than $CV^2$. This means that the size of the capacitor needed in series compensation will be approximately 1.7 times greater than shunt compensation.

2.4.4.1 Comparison of the shunt and series compensated VSR considering saturation

Again, as shown in sections 2.4.1.1 and 2.4.2.1, compared to shunt compensated rectifier, the series compensated rectifier will supply a considerably smaller output power in the whole speed range if a source with saturable inductance is considered. Table 2-2 summarizes the operating conditions for both shunt and series compensation circuits with saturable source inductors at their maximum output power (at nominal speed).
Again, compared to series compensation, considering saturation, the shunt compensation needs a 2.2 times smaller capacitance and is more efficient given the same source EMF voltage and source nominal current.

2.4.5 Shunt and series compensated CSR

To see the effect of compensation on a current source rectifier, the same procedure in sections 2.3.1 and 2.3.2 was applied to a CSR with the same source specifications. A capacitor value equal to 0.825 pu was found to give the highest output power (0.95 pu) at nominal source current \(I_{Lg}=1\) pu for the series compensated rectifier (Fig. 2-30). The peak extractable power at each speed (EMF voltage-frequency) was measured. It was found that the series compensated capacitor will supply a considerably smaller output power compared to the optimal power curve in the whole speed range and therefore was no more investigated.
For the shunt compensated CSR, the 0.8 pu and 0.95 pu capacitors give the highest output power (0.99 pu and 0.96 pu) for the unsaturable and saturable inductances respectively (at nominal source current ($I_{Lg}=1$ pu)). The peak extractable power at each speed (EMF voltage-frequency) for both saturable and unsaturable inductances were compared to the desired optimal power curve. The results are shown in Fig. 2-31.

**Fig. 2-31:** Maximum extractable output power versus source frequency-EMF voltage for shunt compensated CSR with unsaturable source inductance (dashed-dotted, $C_{Sh}=0.8$ pu), with saturable source inductance (line, $C_{Sh}=.95$ pu), the desired optimal power curve (dotted)

While the results for the shunt compensated rectifier may suggest that shunt compensated CSR is an ideal solution for the high saturable source inductance, further investigations show important limitations. Table 2-3 summarizes the operating conditions for both unsaturable and saturable inductors at the maximum output power (at nominal speed). $I_{r,pp}$ is the peak to peak amplitude of the ripple in the output current.
The high value attained for the peak extractable power in the case of unsaturable inductor could be explained by the sinusoidal source current and unity displacement factor \(\cos\phi=1\) at the operating point (nominal speed) at which the capacitor was calculated. In the case of saturable inductor, the lower efficiency is due to higher output current which results in higher rectifier losses. In both cases the high capacitor RMS current will result in considerably high losses if low ESR capacitors are not used. Besides, the high values of ripple current despite the high output inductance (20 pu) could be also considered as a disadvantage.

While the source current for the unsaturable inductor is sinusoidal at nominal conditions (nominal speed), this will not be the case at lower speeds. The source current waveforms for the peak power at nominal speed (frequency-EMF voltage) (Fig. 2-32 (a)) is compared to that of 0.3 pu speed (Fig. 2-32 (b)). Both waveforms are for the saturable inductor, however, at 0.3 pu speed the saturation will not occur and the waveform will be similar to that of the unsaturable inductor.

Table 2-3

<table>
<thead>
<tr>
<th></th>
<th>(C_{sh})</th>
<th>(L_{Out})</th>
<th>(E_g)</th>
<th>(I_g)</th>
<th>(PF)</th>
<th>(DPF)</th>
<th>(\frac{I_{r,pp}}{I_{out}})</th>
<th>(\frac{I_{Out}}{I_n})</th>
<th>(P_{Out})</th>
<th>(I_c)</th>
<th>(\frac{V_{c,ms}}{E})</th>
<th>(\eta)</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Shunt Cap.</strong></td>
<td></td>
<td></td>
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<tr>
<td><strong>Unsat.</strong></td>
<td>0.8</td>
<td>20</td>
<td>1</td>
<td>1</td>
<td>.997</td>
<td>.997</td>
<td>0.1</td>
<td>0.43</td>
<td>0.99</td>
<td>0.92</td>
<td>2.36</td>
<td>99.1</td>
</tr>
<tr>
<td><strong>Inductor</strong></td>
<td></td>
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<tr>
<td><strong>Shunt Cap.</strong></td>
<td>0.95</td>
<td>20</td>
<td>1</td>
<td>1</td>
<td>.979</td>
<td>.998</td>
<td>0.0</td>
<td>0.63</td>
<td>0.96</td>
<td>0.8</td>
<td>1.77</td>
<td>98.4</td>
</tr>
<tr>
<td><strong>Sat.</strong></td>
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<td><strong>Inductor</strong></td>
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The high value attained for the peak extractable power in the case of unsaturable inductor could be explained by the sinusoidal source current and unity displacement factor \(\cos\phi=1\) at the operating point (nominal speed) at which the capacitor was calculated. In the case of saturable inductor, the lower efficiency is due to higher output current which results in higher rectifier losses. In both cases the high capacitor RMS current will result in considerably high losses if low ESR capacitors are not used. Besides, the high values of ripple current despite the high output inductance (20 pu) could be also considered as a disadvantage.

While the source current for the unsaturable inductor is sinusoidal at nominal conditions (nominal speed), this will not be the case at lower speeds. The source current waveforms for the peak power at nominal speed (frequency-EMF voltage) (Fig. 2-32 (a)) is compared to that of 0.3 pu speed (Fig. 2-32 (b)). Both waveforms are for the saturable inductor, however, at 0.3 pu speed the saturation will not occur and the waveform will be similar to that of the unsaturable inductor.
Fig. 2-32: Source current waveforms for the peak power (a) at nominal speed compared to that of (b) 0.3 pu speed

Though the peak output power at a speed of 0.3 pu might be more than the optimal turbine power at that speed, the effect of the current distortion on the torque ripple might not be negligible. For the reasons mentioned above, particularly the high current ripple and the need for high values of output inductance, a VSR with a DC-link inductance could be considered a better choice.
2.5 PWM Rectifier

PWM rectifier is another conventional solution for the high source inductance. The key advantages of PWM rectifiers are as follows:

- Any desired waveform for source current could be imposed (sinusoidal, rectangular, trapezoidal, etc.) despite the saturable stator inductance
- Capability to extract maximal power from the source at all load conditions.
- Adjustable dc link output current or voltage.

The main disadvantages are as follows:

- The switching and conduction losses are much higher than a diode rectifier.
- The price of the PWM rectifier is also considerably higher than a diode rectifier.

As PWM rectifiers are abundantly discussed in the literature [11, 42, 43], in the next section only a very brief description will be presented.

2.5.1 PWM Voltage Source Rectifier (VSR)

Fig. 2-33 shows the power circuit of a single-phase IGBT (Insulated Gate Bipolar Transistor) PWM-VSR and Fig. 2-34 shows its typical waveforms.

Fig. 2-33: The power circuit of a single-phase PWM-VSR
Fig. 2-34: The typical waveforms for a PWM VSR

As could be seen at the input of rectifier very high rates of $dv/dt$ are present. In order to have a smooth source current, a large input filter is required which is already present. This could be considered an important advantage of a VSR for the current application. The phase angle of the input voltage could be controlled by changing the phase angle of the modulating signal. In this way, the desired power factor could be achieved. The output voltage may be kept constant by changing the modulation index as the amplitude of the input voltage varies. Any kind of fast, forced commutated, reverse conducting switches may be used. IGBTs are more widely used, however they have high conduction losses.

Integrated Gate Commutated Thyristors (IGCTs) have lower conduction losses but their switching frequency is typically in the range of 500Hz, though this will not be an important limitation for the low speed MW range turbines.
Despite its advantages, PWM rectifier is not still the best choice if high-currents and high efficiencies are required. This will be later become clearer as PWM rectifier losses are discussed in chapter 3.

2.5.2 PWM Current Source Rectifier (PWM-CSR)

PWM CSR is the dual of PWM VSR and has similar waveforms and control strategy. Fig. 2-35 shows the power circuit of a single-phase IGBT -CSR.

![PWM CSR Circuit Diagram]

Fig. 2-35: The power circuit of a single-phase PWM-CSR

At the input of rectifier high rates of di/dt are required. So given the large source inductance, the PWM-CSR will not function properly if an input capacitance in parallel with the rectifier is not present. This could be considered a disadvantage of CSR compared to VSR for the considered application. The size of the capacitor will decrease as the switching frequency increases, however switching losses will also increase. Any kind of fast, forced commutated, reverse blocking switches may be used or reverse blocking diodes should be put in series with the switch. This results in relatively high semiconductor conduction losses [23].

In the MVA range the operation in Medium Voltage (2.3kV – 7.2kV) is preferred, otherwise the conduction losses will result in low efficiencies. Reverse Blocking Gate
turn-off Thyristors (RB-GTOs) are the conventional choice for CSRs in the MVA range, however they have very high switching and conduction losses and also have high losses in their snubber and gate circuits. Besides, the Peak Repetitive off-state reverse Voltage ($V_{RRM}$) of the RB-GTOs is lower than their Peak Repetitive off-state Forward Voltage ($V_{DRM}$). As an improved version of GTO, IGCT is now considered more attractive for MVA applications [13] and is taking the place of GTO. IGCTs have higher current ratings than IGBTs. Though the ratings of High Voltage IGBTs (HV-IGBTs) are rapidly increasing, their condition losses are still considerably higher than IGCTs.

2.6 The Rapid Polarity-Reversal Mixed-Bridge Rectifier

Our discussion in the previous sections shows that though passive compensated uncontrolled bridge rectifier could be an interesting rectifier solution for a wind turbine application, its use will be limited to the applications where power requirements do not need to be proportional to machine speed. The PWM rectifier, on the other hand, has high conduction losses and is not recommended if high source currents are required. The need for a high efficiency CSR rectifier will be still more evident.

A high efficiency CSR for a variable frequency source with high input inductance should meet the following important requirements:

- It should be simple, reliable and have a reasonable price considering its efficiency.
- It should have low conduction and switching losses in the switches and low losses in the passive components.
- The saturation should not affect the functionality of the rectifier, its efficiency, or its capability to extract power from the source.
- Capability to extract maximal power from the source at all load conditions.

A new topology will be introduced in chapter 3 which will meet all these conditions. It is simple since it uses only a bidirectional switch and a capacitor to reverse the polarity of
the source current and needs a mixed bridge as rectifier (consisting of two forced commutated switches and two diodes). It may use thyristors and diodes as switches with the lowest conduction losses. The compensation capacitor will be switched only during short commutation time which means that unlike passive compensation, losses in the capacitor will be considerably lower. Finally, saturation will not affect its performance.

**Discussion**

The presence of high source inductance will seriously limit the extractable output power of an uncontrolled diode rectifier in both current source and voltage source rectifiers. Among the existing solutions to such a situation, passive compensation has the advantage of simplicity and low price. It was shown that both series and shunt methods could compensate the high source inductance at nominal machine speed for which the compensation capacitor was calculated. However, higher efficiencies could be attained with shunt compensation. It was also shown that at nominal speed, more active power could be extracted with shunt compensation than series compensation if saturation is considered. For a variable speed machine, shunt compensation provides more active power than series compensation at all speeds and the optimal power could be extracted from the wind, however, saturation will slightly reduce efficiency and power transfer capability of the rectifier. The series compensated rectifier could not extract the optimal turbine power below its nominal speed.

While shunt compensation will be the best choice for a CSR, at low speeds, the source current will be highly distorted. Besides, due to high ripple in the output current, large dc-link inductors will be required which will further increase the losses. This means that a VSR with dc link inductance will be a better choice for a CSR.

Passive compensated rectifier could not considered as an optimal choice in a variable speed system if it could not extract the maximum available power from the source.
PWM converters are very versatile rectifiers for variable frequency sources with high input inductance, however, they have high conduction losses which will result in low efficiencies particularly at high source currents. Their efficiency will be improved if IGCTs with low switching frequencies are used. Given the limitations of passive compensation and PWM rectifiers, alternative topologies are required.
Chapter 3

The Rapid Polarity Reversal (RPR) mixed-bridge Rectifier

In the previous chapter the conventional rectifier solutions were discussed. In this chapter a new topology will be introduced which could be used with sources with variable frequency and high saturable source inductance such as transverse flux machines. This chapter begins with a general description of the proposed topology. The choice of switches will be later discussed and a detailed circuit analysis will follow. The rating analysis and the performance analysis of the new topology will be also presented. Simulation and analytical calculation will be used to compare the losses of the new topology with that of a PWM rectifier. Finally, the simulation and experimental results will be presented.

3.1 General description

Fig. 3-1 shows the suggested AC/DC converter topology for a constant current load. The proposed topology allows a rapid reversal of the current polarity in a TFPM machine despite its large stator inductance. During current commutation, the rectifier is disconnected from the source and the capacitor is connected in series with the source through a bidirectional switch. The rectifier is a mixed bridge of two diodes and two controlled switches. The controlled switches are part of the rectifier and also function as switches for disconnecting the resonant circuit from the output current source during
commutation. D1 and D2 are also part of the rectifier, but function as freewheeling diodes during this period.

Fig. 3-2 shows the Source EMF Voltage and current. The switch conduction pattern is shown in Fig. 3-3.

![Diagram](image-url)

**Fig. 3-1** The proposed Resonant AC/DC Converter

**Fig. 3-2** Source EMF voltage (Eg) and Inductor current (I_Lg) for one cycle
3.2 The choice of switches

A variety of switches may be used for S1-S4 in Fig.1 including thyristors, IGBTs, IGCTs, Gate Turn-Off thyristors (GTOs), Emitter Turn-Off thyristors (ETOs), and Symmetrical Gate Commutated Thyristors (SGCTs).

3.2.1 Switches for the Controlled Half-Bridge

Due to the low switching frequency of the converter, the conduction losses will be the most important criteria for the selection of the switches for the controlled half-bridge. Table 3.1 compares the on-state voltages for four types of switches with roughly similar ratings where $I_{TAVM}$ is the maximum average on-state current, $V_{DRM}$ is the peak repetitive off-state forward voltage, $V_T$ is the on-state voltage, $I_{c(max)}$ is the maximum collector DC current, $V_{CE(max)}$ is the maximum collector-emitter voltage, and $V_{CE(sat)}$ is the collector-emitter saturation voltage.

It should be noted that thyristors, GTOs, and IGCTs are generally rated in terms of $I_{TAVM}$ for half-wave operation while IGBTs are rated in terms of $I_{c(max)}$. Though these will not be equivalent, it will be good enough for the purpose of this comparison. Another
important consideration is the fact that the characteristics of each switch type widely varies for the same ratings since:

1. It may be optimized by the manufacturer for either low switching or low conduction losses.
2. Different manufacturing technologies may be used for each switch which affects the conduction losses.

Table 3.1: The on-state voltages for four switch types

<table>
<thead>
<tr>
<th></th>
<th>5SGA 30J4502 GTO</th>
<th>CM900HB-90H IGBT</th>
<th>5SHX 26L4510 IGCT</th>
<th>5STP 12F4200 THYRISTOR</th>
</tr>
</thead>
<tbody>
<tr>
<td>I_{TAVM}</td>
<td>930 A</td>
<td>900 A</td>
<td>1010 A</td>
<td>1150 A</td>
</tr>
<tr>
<td>I_{C(max)}</td>
<td>4500 V</td>
<td>4500 V</td>
<td>4500 V</td>
<td>4200 V</td>
</tr>
<tr>
<td>V_{T(max)}</td>
<td>2.75 V (max.)</td>
<td>3.5 V (typical)</td>
<td>2.25 V (max.)</td>
<td>1.55 V (max)</td>
</tr>
</tbody>
</table>

As could be seen, the thyristor has the lowest conduction losses and the IGCT occupies the second place. Since this could be considered as a general trend, thyristors will be the first choice as switches for S1 and S2.

**Thyristors**

Low conduction losses is not the only merit of thyristors, they have also the lowest prices among controlled power switches. Another interesting feature of the thyristors is their reverse blocking capability. However, unlike GTOs and IGCTs, it impossible to turn off thyristors by extracting current through the gate. To turn off the thyristor the current flow from anode to cathode should be reduced to a value below its holding current and a negative voltage should be applied between the anode-cathode terminals for a time equal
to its turn off time ($t_q$) to allow the recombination of the base junctions stocked charges and to guarantee the return to the blocking state before a forward voltage is re-applied [30]. Commutation could be natural or externally generated by an auxiliary circuit. Typical thyristor blocking waveforms are shown in Fig. 3-4.

![Fig. 3-4 Typical thyristor blocking waveforms][27]

Turn-off time depends upon a number of circuit conditions including on state current prior to turn-off, rate of change of current during the forward to reverse transition, reverse blocking voltage, rate of change of reapplied forward voltage, the gate bias, and junction temperature. Increasing junction temperature and on-state current both increase turn-off time and have a more significant effect than any of the other factors. Negative gate bias will decrease the turn-off time [27].

Thyristors are generally divided in two groups: Phase Control thyristors and Inverter type thyristors (also called Fast Switching Thyristors). Fast thyristors have fast turn-off times (typically under 50 $\mu$s), however, their on-state voltage drop and their turn-off times vary in an inverse manner. Besides, Fast thyristors are not available at medium voltages (2.3kV – 7.2kV). In the current application, the thyristors in the controlled half-bridge should be forced commutated since, despite the current commutation between the
conducting thyristor and the input capacitor, the thyristor (S1 or S2) will not turn off naturally since the voltage on the rapidly charging capacitor will not permit the reverse bias of the thyristor during the turn-off time. This will become more clear during our detailed analysis in section 3.4.

GTO
As a member of the thyristor family, GTOs may be preferred to thyristors because they are self-commutated devices and unlike thyristors do not require an external forced commutation circuit. However, due to the following reasons, they are not recommended for this application:

- Higher conduction losses compared to thyristors.
- Gate currents are higher which means higher losses in the Gate drive circuit
- Their reverse blocking capability is less than their forward blocking capability.

IGBT
Unlike Thyristors, IGBTs have no reverse blocking capability. They should be used with a series RB (Reverse Blocking) diode. The RB-IGBT version has an integrated reverse blocking diode. Such a combination will have a very high on-state voltage drop. Besides, IGBTs have the highest prices among the switches considered so they will not be an interesting choice for the proposed topology. In the future, however, due to rapid progress in HV-IGBTs, they might become more interesting as their prices fall and their saturation voltage decreases.

IGCT
IGCT is a more recent member of thyristor family, introduced by ABB in 1996. Its on-state voltage drop could be quite close to that of thyristors. It has the following advantages for this application:

- Requires no forced commutation circuit which means more reliability and simplicity.
- The optically isolated gate unit is already integrated into the device.
Its drawback in this applications is that its reverse blocking version (reverse blocking diode integrated into device), has much higher on-state voltage drop. However, as will be shown in section 3.8.2, it is still considered as an interesting choice for this application in the MW range.

**Other Switches**

Emitter Turn-off (ETO) thyristors are not yet commercially available and will not be examined here, however, according to [40], they has similar conduction losses to IGCTs. Symmetric Gate-Commutated Thyristors (SGCTs) have reverse blocking capability and are already used in certain commercial current source inverters, however, their availability is limited and their specification sheets do not yet exist. Both are forced commutated switches and might be interesting choices in the future.

The ratings for different commercially available switches are shown in figure 3.5 and 3.6. As Fig. 3-5 shows, for the low frequency applications, the power rating of the thyristors is far superior to other switches. Fig. 3-6 shows, their voltage and current ratings. As could be observed, the voltage and current ratings for the thyristors are twice higher than their closest rivals.

### 3.2.2 Bidirectional Switch

For the bidirectional switch (S3, S4), thyristors are preferred since the commutation could be natural. Besides the conduction time in the bidirectional switch is limited to the resonance interval, which means that the current rating of the switch does not need to be as high as the main switches, it should rather have high pulse current ratings and high di/dt. Pulsed power thyristors are designed for such specifications and are available through a number of manufacturers like Dynex. They could also be operated in GTO mode with the appropriate commutating gate drive.
Fig. 3-5 Ratings for Power Switching Devices [31]

Fig. 3-6: Current and voltage ratings for Power Switching Devices [31]
3.3 Advantages of the Proposed Topology

The main advantages of this configuration for sources with high saturable inductance are:

- The capability to extract the maximal power from the source despite the high inductance.
- Thyristors and diodes could be used as main switches. They have the lowest conduction losses.
- Core saturation does not affect the functionality of the rectifier.
- Unlike a PWM rectifier, losses will be limited to conduction losses in the switches.
- Due to high efficiencies, cooling requirements for the switches are lower.
- Requires no dv/dt snubber (no snubber losses).
- Could be used with a low loss CSI inverter.

The detailed loss calculation and comparison to PWM converter will be discussed in detail in sections 3.8.1 and 3.8.2.

3.4 Circuit Analysis

Fig. 3-7 shows the required control blocks including the bidirectional switch gate control, thyristor half-bridge gate control, and the forced commutation circuit for the thyristor (switches S1, S2) version of the converter used in the experimental circuit which will be presented later. This analysis will be also valid for the IGCT version if the thyristor turn-off interval is ignored.
3.4.1 Forced commutation circuit (FCC)

The forced commutation circuit is not required if S1 and S2 are IGCTs, GTOs and IGBTs. If S1 and S2 are chosen to be thyristors, they will not turn off naturally (as mentioned before). The forced commutation circuit may be realized in different ways. One possible solution is the circuit shown in Fig. 3-8. The advantage of such a circuit is its simplicity. The power rating of external power supplies ($E_1$ and $E_2$) are very small compared to the power extracted from the source ($e_g$). The losses in the FCC are limited to conduction losses and are proportional to input current of the rectifier and the turn-off time of the thyristors in the controlled half-bridge. This circuit will be used later as FCC for the experimental circuit and its use may be justified only in low voltage ratings (under 1200V), where IGBTs are generally low priced and Pulse IGBTs may be used whose pulse current ratings are much higher than their average and rms ratings. For high power, either IGCTs should be used, or for better efficiency, thyristor based FCCs should be designed.
3.4.2 Detailed Analysis

The following analysis applies to thyristor version (switches S1, S2) of the rectifier with the forced commutation circuit of Fig. 3-8, however it will be valid for the IGCT version of the rectifier or any other forced commutation circuit with slight modifications. In the rest of this chapter, \( i_X \) and \( v_X \) will be used as symbols for the instantaneous current and voltage through switch \( X \). \( V_{T(X)} \) is the on-state voltage drop on the switch \( X \) (IGCT or thyristor). \( V_{F(X)} \) is the forward voltage on the diode \( X \). \( v_c \) is the instantaneous capacitor voltage. \( i_{ig} \) is the instantaneous source current. \( I_{nom} \) is the nominal value of source RMS current. \( E \) and \( E_g \) are the peak and RMS values of the source EMF respectively. \( v_g \) is the output voltage of the source. \( v_{out} \) and \( I_{out} \) are the rectifier’s output instantaneous voltage and DC current respectively.

The analysis begins with interval 0 (Fig. 3-9) which ends before \( t_0 \). 

**Interval 0: Normal Conduction Mode**

T1 and D2 conduct. T3 and T4 are blocked.

\[
i_{T1}=i_{D2}=I_{out} \tag{3-1}
\]

\[
v_c=0 \tag{3-2}
\]

We could write:

\[
-v_{out}=-v_g^+ (V_{T(T1)}+V_{F(D2)}) \tag{3-3}
\]

and

\[
v_{out}=V_{F(D2)}+V_{D1} \tag{3-4}
\]
So we have:

\[ v_{D1} = -v_g + V_{T(T1)} \]  

\[ (3-5) \]

Fig. 3-9: Current path for intervals 1 to 3
Thyristor turn-off Interval: \((t_0-t_q \rightarrow t_0)\)

At \(t=t_0-t_q\), Tr1 is activated for a period of time equal or larger than the turn-off time of T1 (\(t_q\)).

\(E_t\) is chosen to be greater than the total voltage drop on Da1, Tr1, and D1 (at nominal current):

\[|E_t| > V_F(Da1) + V_T(Tr1) + V_F(D1)\]  \((3-6)\)

Since:

\[v_g = -|E_t| - V_F(Da1) - V_T(Tr1)\]  \((3-7)\)

we have:

\[v_g < -V_F(D1)\]  \((3-8)\)

According to (3-5), this will forward bias D1 and an instantaneous commutation between Tr1, T1, and D2 will take place. Input current will flow through Tr1 and the output current will flow through D1 and D2 (freewheeling diodes):

\[i_{Lg} = i_{Tr1} = i_{Da1}\]  \((3-9)\)

\[I_{Out} = i_{D1} = i_{D2}\]  \((3-10)\)

\[i_{T1} = 0\]  \((3-11)\)

\[v_{T1} = v_g + V_F(D1) < 0\]  \((3-12)\)

The last two conditions will ensure the blocking of the thyristor at \(t=t_0\).

Transition Interval 0: Resonance mode \((t_0 \rightarrow t_0+)\)

At \(t=t_0\) T3 is triggered and Tr1 is deactivated. To insure the continuity of the source current, deactivation of Tr1 should be delayed with respect to triggering of T3. To determine the proper timing, the turn-on time of the thyristor (\(t_{on}\)) and the turn-off delay time of Tr1 (\(t_{off}\)) should be considered.

Since \(v_g\) is initially negative and, no current will flow through T3 until Tr1 begins to turn off and \(v_g\) increases beyond the threshold voltage of T3 (it is supposed that \(v_C(t_0) = 0\)). As a result, at the end of this interval, \(i_{Lg}\) will flow through T3 and begin to charge capacitor C. \(I_{out}\) will continue to flow through the freewheeling diodes.
**Interval 1** Resonance mode \((t_0 \rightarrow t_1)\)

During the resonance mode the energy stored in the source inductance will be transferred to the capacitor and current will fall to zero.

During this interval we have:

\[
i_{T3} = i_C = i_{Lg} \quad (3-13)
\]
\[
i_{T1} = i_{T2} = 0 \quad (3-14)
\]
\[
i_{D1} = i_{D2} = I_{Out} \quad (3-15)
\]

At the end of this interval, capacitor \(C\) will be charged to \(V_{C_{max}}\) and current in the inductor falls to zero.

**Interval 2** Resonance mode \((t_1 \rightarrow t_2)\)

During the resonance mode the energy stored in the capacitor will be transferred to the source inductance and the polarity of the source current will be reversed (compared to its initial polarity at the beginning of interval 1). At the beginning of this interval we have:

\[
V_{T4} = V_{C_{max}} - V_g > 0 \quad (3-16)
\]

So \(T4\) will conduct as it is triggered and \(C\) will begin to discharge. As a result current in the inductor increases.

We have:

\[
v_g = V_C - V_{T(T4)} \quad (3-17)
\]
\[
v_g = V_{T2} + V_F(D2) \quad (3-18)
\]

As \(V_g\) becomes negative and the forward voltage on \(T2\) becomes greater than its threshold voltage \((V_{T0(T2)})\), \(T2\) could be triggered.

**Transition Interval 1** : Resonance mode \((t_2 \rightarrow t_2^+)\)

As \(T2\) is triggered, the behavior of the circuit will be determined by the relation between source and load currents. Three conditions may exist:

a) \(i_{Lg} = I_{Out}\)

   Instantaneous commutation between \(T2\), \(D2\), and \(C\) will take place. We have:

\[
i_C = i_{D2} = 0 \quad (3-19)
\]
b) $i_{Lg} < I_{out}$

T2 and D2 will commutate until $I_{Lg}$ equals $I_{out}$. We have:

$$i_{Lg} = i_{T2}$$  \hspace{1cm} (3-21)

$$I_{out} = i_{T2} + i_{D2}$$  \hspace{1cm} (3-22)

c) $i_{Lg} > I_{out}$

Capacitor C will continue to conduct until $I_{Lg}$ equals $I_{out}$. We have:

$$i_c = i_{Lg} - I_{out}$$  \hspace{1cm} (3-23)

These conditions will be further discussed in section 3.6.2.

**Interval 3: Normal Conduction Mode ($t_2 \rightarrow t_3$)**

This interval is similar to Interval 0.

### 3.4.3 Analysis of the Resonant Circuit

During commutation, the characteristics of the source current will be decided by the resonant circuit formed by the source EMF voltage ($E_g$), source inductance ($L_g$), equivalent circuit resistance ($R$), and the series connected capacitor ($C$). That is why a detailed analysis of the resonant circuit will follow.

The equivalent circuit during commutation interval is shown in Fig.3-10. The Equivalent Series Inductance (ESI) of the capacitor will be negligible compared to $L_g$ and will not be considered. The equivalent resistance of the resonant circuit ($R$) could be written as follows:

$$R = R_g + R_c + r_{on} \text{ (BSW)}$$  \hspace{1cm} (3-24)
where $R_g$ is the equivalent resistance of the inductor, $R_C$ represents the ESR (Equivalent Series Resistance) of the capacitor and $r_{on}(BSW)$ is the slope resistance of the conducting switch (bidirectional switch). The threshold voltage of the conducting switch ($V_{T0}$) could be treated as a voltage source in series with the source voltage.

For the purpose of the current analysis, $L_g$ is treated as an unsaturated inductance. The effects of saturation on the resonant circuit will be presented later as the saturable inductor model is presented.

During the resonance interval, the resonant circuit is disconnected from the rectifier while the bidirectional switch stays connected. Using Kirchhoff's Voltage Law (KVL) we could write the following equation:

$$e_g = L_g \frac{di_g}{dt} + Ri_g + \frac{1}{C} \int i_g \, dt$$

(3-25)

By differentiating Eq. (3-25) we have:

$$C \frac{de_g}{dt} = L_g C \frac{d^2 i_g}{dt^2} + RC \frac{di_g}{dt} + i_g$$

(3-26)
For a rapid polarity reversal, low values of firing angle \( \theta_f = \frac{2\pi(t_1 - t_0)}{T} \) in Fig. 3-2 and 3.13) are desired. In such a case, \( E_g \) during resonance interval is small enough. By neglecting \( E_g \), the Eq. (3-26) will simplify and its homogenous solution could be found as:

\[
\begin{align*}
  i_{Lg} (t) &= e^{-\alpha t} (A_1 \cos \omega_r t + A_2 \sin \omega_r t) \\
  v_C (t) &= e^{-\alpha t} (B_1 \cos \omega_r t + B_2 \sin \omega_r t)
\end{align*}
\]  

(3-27)  

(3-28)

where

\[
\begin{align*}
  \alpha &= \frac{R}{2L_g} \\
  \omega_r &= \sqrt{\omega_0^2 - \alpha^2} \\
  \omega_0 &= \frac{1}{\sqrt{L_g C}}
\end{align*}
\]  

(3-29)  

(3-30)  

(3-31)

We assume the following initial conditions which conform to conditions in \( t_0 \) and \( t_3 \):

\[
\begin{align*}
  V_C (0) &= 0 \quad (3-32) \\
  i_{Lg} (0) &= I_{Out(0)} \quad (3-33) \\
  V_{CF} &= 0 \quad (3-34) \\
  I_{LgF} &= 0 \quad (3-35)
\end{align*}
\]

where \( V_{CF} \) and \( I_{LgF} \) are the forced response for the capacitor voltage and inductor current respectively.

In the state plane \((i_{Lg}, \sqrt{\frac{L_g}{C}}, v_C)\), the response of the circuit will be a spiral centered on \((0, 0)\). For a half period the response is shown in Fig. 3-11.
For a more precise solution, or for higher values of firing angle, the effect of source EMF voltage during the resonance phase should be considered. In that case, it will be represented as:

\[ e_g(t) = E \sin (\pi + \omega_g (t-t_1)) \]  

where \( \omega_g \) is the radian frequency of the source EMF voltage, \( E \) is its peak value and \( t_1 \) represents the zero crossing instant of the source EMF voltage (as shown in Fig. 3-2).

\[ e_g(t) = E \sin (\pi + \omega_g (t-t_1)) \approx E (\omega_g (t-t_1)) \]  

Inserting Eq. (3-37) into Eq. (3-26) gives:

\[-CE\omega_g = L_g C \frac{di_{lg}}{dt^2} + RC \frac{di_{lg}}{dt} + i_{lg}\]  

Assuming the particular solution is in the form of a constant \( K \) as suggested by the first term in the Eq. (3-38), the total solution to this equation will be:

Fig. 3-11 State Plane for the response of the resonant circuit

Supposing \( \omega_g << \omega_r \), during the resonance interval, we have:

\[ e_g(t) = E \sin (\pi + \omega_g (t-t_1)) \approx E (\omega_g (t+t_1)) \]  

Inserting Eq. (3-37) into Eq. (3-26) gives:

\[-CE\omega_g = L_g C \frac{di_{lg}}{dt^2} + RC \frac{di_{lg}}{dt} + i_{lg}\]  

Assuming the particular solution is in the form of a constant \( K \) as suggested by the first term in the Eq. (3-38), the total solution to this equation will be:
\[ i_{lg}(t) = K + e^{-\alpha t} \left( A_1 \cos \omega_r t + A_2 \sin \omega_r t \right) \]  \hspace{1cm} (3-39)

where \( K = -CE \omega_g \)

By applying the initial conditions to equations (3-39) and (3-25), we have:

\[ i_{lg}(0) = I_0 = K + A_1 \]  \hspace{1cm} (3-40)

and

\[ L_g \frac{di_{lg}(0+)}{dt} = E_{\omega g} \frac{\pi}{2\omega_r} - RI_0 \]  \hspace{1cm} (3-41)

\[ L_g \frac{di_{lg}(0+)}{dt} = -\alpha A_1 + \omega_r A_2 \]  \hspace{1cm} (3-42)

The coefficients \( A_1 \) and \( A_2 \) will be:

\[ A_1 = I_0 - K \]  \hspace{1cm} (3-43)

\[ A_2 = -\frac{\alpha}{\omega_r} I_0 - K \left( \frac{\alpha}{\omega_r} + \frac{\pi K \omega_0}{2\omega_r} \right) \]  \hspace{1cm} (3-44)

The damping ratio is defined as:

\[ \zeta = \frac{\alpha}{\omega_0} \]  \hspace{1cm} (3-45)

We could write:

\[ \zeta = \frac{\alpha}{\omega_0} = \frac{R}{X_L} \]  \hspace{1cm} (3-46)

In a circuit like that of a transverse flux machine (as source), we have:

\[ \zeta \ll 1 \]  \hspace{1cm} (3-47)

or

\[ \alpha \ll \omega_0 \]  \hspace{1cm} (3-48)
So we could write:

\[ \omega_r \approx \omega_0 \]  

(3-49)

and

\[ \alpha \ll \omega_r \]  

(3-50)

So the Eq.(3-39) will become:

\[
\begin{align*}
    i_{Lg}(t) &= K + e^{-\alpha t} ((I_0 - K) \cos \omega_r t + (-\frac{\alpha}{\omega_0} I_0 - \frac{\pi}{2} K) \sin \omega_r t) \\
\end{align*}
\]  

(3-51)

For a resonance half-cycle:

\[
e^{-\alpha t} \equiv 1 - \alpha t
\]  

(3-52)

As a result Eq. (3-51) could be further simplified and rewritten as:

\[
\begin{align*}
    i_{Lg}(t) &= (I_0 (1 - \alpha t) \cos \omega_r t) - \left( CE \omega_g (1 - \cos \omega_r t - \frac{\pi}{2} \sin \omega_r t) \right) \\
\end{align*}
\]  

(3-53)

The last term in the Eq. (3-53) represents the effect of source EMF voltage on the inductor current during the resonance phase.

The inductor current at the end of the resonance half period (I_1) will be:

\[
I_1 = -I_0 (1 - \frac{\alpha \pi}{\omega_r}) - 2CE\omega_g
\]  

(3.54)

since \( I_0 \) and \( I_1 \) always have the reverse polarities:

\[
\Delta I_{Lg} = |I_1| - |I_0| = -|I_0| \frac{\alpha \pi}{\omega_r} + 2CE\omega_g
\]  

(3-55)
The term \(-|\eta|\frac{\alpha_{\eta}}{\omega_r}\) in Eq. (3-55) shows the effect of damping on the initial inductor current after a half cycle, while the term \(2C\omega_0E_g\) represents the effect of the source EMF voltage. As expected, the source EMF voltage and its resistance have opposing effects on the final current (after a half-cycle). While a higher value for \(E_g\) increases the amplitude of the inductor current, higher \(R_g\) values will decrease it. It is also interesting to notice that the instantaneous value of source EMF voltage at the start of the resonance phase has no effect on \(\Delta I_{Lg}\). It is rather its peak value \((E)\) which appears in Eq. (3-55).

3.5 Rating Analysis

In this section a detailed rating analysis for all the components of the RPR converter will be presented. The results will be used later for numeric calculation and price analysis. For the purpose of this analysis, thyristors will be considered, however the analysis applies to any controlled switch. In the case of IGCTs the analysis applies to the combination of IGCT and its series reverse blocking diode.

3.5.1 Thyristor and Diode Half-Bridges (T1,T2,D1,D2)

**Peak repetitive off-state voltage**

During commutation, high voltages will be imposed on the thyristor half bridge.

We have:

\[ V_{T1} = V_g + V_{F(D1)} \approx V_g \]  \hspace{1cm} (3-56)

Since

\[ V_g = V_C + V_{T(T3)} \approx V_C \]  \hspace{1cm} (3-57)

so we have:

\[ V_{T1} \approx V_C \]  \hspace{1cm} (3-58)

so the Peak Repetitive off-state Forward Voltage \((V_{DRM})\) should be greater than \(V_{C_{max}}\).
During the same interval we have:

\[-V_{T2} = V_g - V_{F(D1)} = V_g \approx V_C\]  \hspace{1cm} (3-60)

So the Peak Repetitive off-state reverse Voltage \((V_{RRM})\) for T1 and T2 should be also greater than \(V_{C_{max}}\):

\[V_{RRM(T1,T2)} > V_{C_{max}}\]  \hspace{1cm} (3-61)

For the diode half-bridge, during the resonance interval, both diodes will conduct. So the \(V_{RRM(D1,D2)}\) will be decided in the normal conduction mode:

\[V_{RRM(D1,D2)} = \sqrt{2}E_{g(nom)}\]  \hspace{1cm} (3-62)

**Snubber**

Generally, snubbers are used along with thyristors to limit the \(dv/dt\) across the thyristors, however, in the RPR converter, no such snubbers are required since \(dv/dt\) will be determined by the voltage across the capacitor \(C\). We have:

\[I_C(t) = c \cdot \frac{dv_c}{dt}\]  \hspace{1cm} (3-63)

so from Eq. (3-58) we have:

\[\frac{dv_{T1}}{dt} \approx \frac{i_C(t)}{C}\]  \hspace{1cm} (3-64)

since the current \(i_C(t)\) will have its maxima at the start of the resonance interval \((t=t_0)\) where \(i_C(t_0) = I_{out}\), we have:

\[\frac{dv_{T1}}{dt} (\text{max}) \approx \frac{I_{out}}{C}\]  \hspace{1cm} (3-65)

The condition \(\frac{I_{out}}{C} < \frac{dv_{T1}}{dt} (\text{max})\) could be easily met considering the typical values of \(dv/dt\) for the thyristors and typical source parameters.
For the IGCTs, due to their high rates of $di/dt$, no $di/dt$ snubber or small values are required. Pulse thyristors also have very high $di/dt$ ratings. For the phase control thyristors and diodes the choice of $di/dt$ snubber will be based on their specifications.

**RMS and Average Currents**

Assuming constant load current, we have:

\[
I^2_{T2(\text{rms})} = \frac{1}{2\pi} \int_{\theta_f}^{\pi-\theta_f} I_{\text{out}}^2 d\theta = \frac{\pi - 2\theta_f}{2\pi} I_{\text{out}}^2
\]  

\[
I_{T2(\text{rms})} = \sqrt{\frac{1}{2} \frac{\theta_f}{\pi}} I_{\text{out}}
\]  

\[
I_{T2(\text{avg})} = \frac{1}{2\pi} \int_{0}^{\pi} I_{\text{out}} d\theta = \frac{\pi - 2\theta_f}{2\pi} I_{\text{out}}
\]  

\[
I_{T2(\text{avg})} = \frac{1}{2} \frac{\theta_f}{\pi} I_{\text{out}}
\]

For the diode half-bridge we have:

\[
I^2_{D2(\text{rms})} = \frac{1}{2\pi} \int_{0}^{\pi+2\theta_f} I_{\text{out}}^2 d\theta = \frac{\pi + 2\theta_f}{2\pi} I_{\text{out}}^2
\]  

\[
I_{D2(\text{rms})} = \sqrt{\frac{1}{2} \frac{\theta_f}{\pi}} I_{\text{out}}
\]  

\[
I_{D2(\text{avg})} = \frac{1}{2\pi} \int_{0}^{\pi+2\theta_f} I_{\text{out}} d\theta = \frac{\pi + 2\theta_f}{2\pi} I_{\text{out}}
\]  

\[
I_{D2(\text{avg})} = \frac{1}{2} \frac{\theta_f}{\pi} I_{\text{out}}
\]
3.5.2 Bi-directional Switch (T3, T4)

RMS and Average Currents

To simplify our calculation we will neglect the damping in the resonant circuit. This will not result in much inaccuracy in the case of sources with low damping factor as discussed in section 3.4.3. The current \( i_{T3} \) will be identical to \( i_c \) (in Fig. 3-12) in the time intervals \( t_0-t_1, t_5-t_4 \) and equal to zero for the rest of the period.

We have:

\[
I_{T3(rms)}^2 = \frac{1}{2\pi} \int_0^{\theta_f} I_{out}^2 \cos^2 \frac{\pi}{2\alpha} \theta \, d\theta = \frac{\theta_f}{2\pi} I_{out}^2
\]  
(3-74)

\[
I_{T3(rms)} = I_{out} \sqrt{\frac{\theta_f}{2\pi}}
\]  
(3-75)

and

\[
I_{T3(avg)} = \frac{1}{2\pi} \int_0^{\theta_f} I_{out} \cos \frac{\pi}{2\theta_f} \, d\theta = \frac{2\alpha}{\pi} I_{out}
\]  
(3-76)

Peak repetitive off-state voltage

If T4 is triggered before current falls to zero in T3 (at the end of interval 1 as shown in section 3.4), the voltage on the thyristors during resonance in the bidirectional switch will be limited to the forward voltage of the thyristors. So the voltage rating of the switches will be rather imposed by the conduction period:

\[
V_{DRM(T3)} = V_{RRM(T3)} = \sqrt{2} * E_{g,nom}
\]  
(3-77)

3.5.3 Capacitor

Peak capacitor Voltage

Since \( V_{C_{\text{max}}} \) determines the \( V_{RRM} \) and \( V_{DRM} \) of the thyristors, the parameters affecting its value will be discussed in detail. Assuming no damping, all the energy stored in \( L_g \) will be transferred to C:
\[ L_g I_{L_g, \text{max}}^2 = CV_{c_{\text{max}}}^2 \] (3-78)

For a constant output current, \( I_{L_g, \text{max}} \) equals the output current (\( I_{\text{Out}} \)), giving:

\[ V_{c_{\text{max}}} = I_{\text{Out}} \sqrt{\frac{L_g}{C}} \] (3-79)

For a constant output current and trapezoidal input current, as shown in Fig. 3-2, we have:

\[ I_{L_g (\text{rms})}^2 = \frac{1}{2\pi} \int_0^{2\pi} I_{L_g (\omega)} \cdot d\omega \]

\[ = \frac{2\theta_f}{\pi} \frac{I_{\text{Out}}^2}{9} + \frac{2\pi - 4\theta_f}{\pi} I_{\text{Out}}^2 \] (3-79)

\[ I_{L_g (\text{rms})} = I_{\text{Out}} \sqrt{1 - \frac{4\theta_f}{3\pi}} \] (3-80)

so for the nominal current we could write:

\[ I_{\text{Out}, \text{nom}} = \frac{I_{L_g, \text{nom}}}{\sqrt{1 - \frac{4\theta_f}{3\pi}}} \] (3-81)

We define \( X_{g (pu)} \) as:

\[ X_{g (pu)} = X_g \frac{I_{L_g, \text{nom}}}{E_g, \text{nom}} \] (3-82)

so we could write:

\[ I_{L_g, \text{nom}} = \frac{E_g, \text{nom} \cdot X_{g (pu)}}{L_g \omega_g} \] (3-83)

Substituting equations (3-81) and (3-82) into Eq. (3-79) will give:

\[ V_{c_{\text{max}}} = E_g, \text{nom} \cdot X_{g (pu)} \frac{\pi}{2\theta_f} \frac{1}{\sqrt{1 - \frac{4\theta_f}{3\pi}}} \] (3-84)
Since $\theta_f = \frac{\pi \theta_g}{2 \theta_0}$, we may alternatively write:

$$V_{C_{\text{max}}} = E_{g, \text{nom}} X_{g, \text{(pu)}} \frac{\omega_0}{\omega_g} \left( \frac{1}{2 \omega_g} \right) \sqrt{1 - \frac{2 \omega_g}{3 \omega_0}}$$  \hspace{1cm}(3.85)

As Eq.(3.84) shows, for a given generator power and reactance, the peak capacitor voltage will be dictated by the choice of nominal voltage and the firing angle $\theta_f$. Fig. 3-12 shows $\frac{V_{C_{\text{max}}}}{E_{g, \text{nom}}}$ as a function of $\frac{\omega_0}{\omega_g}$.

![Graph showing $\frac{V_{C_{\text{max}}}}{E_{g, \text{nom}}}$ as a function of $\frac{\omega_0}{\omega_g}$](image)

Fig. 3-12: $\frac{V_{C_{\text{max}}}}{E_{g, \text{nom}}}$ as a function of $\frac{\omega_0}{\omega_g}$ from 1 to 20 equivalent to $\theta_f$ from 90° to 4.5°

As could be seen, reducing $\frac{\omega_0}{\omega_g}$ will lower the rating of the converter, however, as will be shown later in section 3.7.1, it will also affect the capability of the converter to extract maximal power from the source.
Capacitor RMS Current

We could write:

\[ I_{C(rms)}^2 = I_{T3(rms)}^2 + I_{T4(rms)}^2 \]  \hspace{1cm} (3-86)

Using the Eq. (3-75) we obtain:

\[ I_{C(rms)} = I_{out} \sqrt{\frac{\theta_f}{\pi}} \]  \hspace{1cm} (3-87)

3.6 Analysis of the Control Strategy for the RPR-Converter

In this section, we will discuss some of the issues concerning the trigger timings for the proper operation of the converter. The appropriate control strategy depends on the type of switches used. Since the purpose of this analysis is not to replace a detailed design process, only some of the main issues will be mentioned.

3.6.1 Gate Control circuit for the Bidirectional Switch

The resonance interval always begins with the firing of one thyristor in the bidirectional switch (T3 for interval 1). In order to maximize the power output and minimize \( V_{C_{\text{max}}} \), the exact charging time of the capacitor could be calculated so that \( V_{C_{\text{max}}} \) coincides with the zero crossing of the EMF voltage and source current. For the unsaturated inductance, this could be calculated from Eq. (3-30) and Eq. (3-31).

\[ t_1 - t_0 = \frac{\pi}{2 \sqrt{\omega_0^2 - \alpha^2}} = \frac{\pi}{2} \sqrt{\frac{L_g}{C}} \]  \hspace{1cm} (3-88)

For a saturable inductance, the calculation could be done through a look-up table, as will be explained later in section 3.10.2 where the simulation model is explained.
The next thyristor in the bidirectional switch (T4 at \( t_1 \) or T3 at \( t_4 \)) could be triggered just before the zero crossing of the inductor current when the first thyristor turns off. This will coincide with the zero crossing of the source EMF voltage if the timing for the trigger of the first thyristor is well calculated. In the resonance interval the second thyristor should always be triggered before the turning-off of the first thyristor, otherwise the thyristors might be exposed to \( V_{C_{\text{max}}} \). This will be later discussed in section 3.5.2.

Fig. 3-13: Source EMF voltage \( (e_g) \), Source Current \( (i_{Lg}) \), Capacitor Voltage \( (v_C) \) and Capacitor current \( (i_C) \) for the suggested converter
Table 3.2 The trigger timing for the suggested converter

<table>
<thead>
<tr>
<th>Time</th>
<th>( t_0 )</th>
<th>( t_1 )</th>
<th>( t_2 )</th>
<th>( t_3 )</th>
<th>( t_4 )</th>
<th>( t_5 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thyristor</td>
<td>T3</td>
<td>T4</td>
<td>T2</td>
<td>T4</td>
<td>T3</td>
<td>T1</td>
</tr>
</tbody>
</table>

3.6.2 Gate Control for the controlled Half-Bridge

For the analysis in this section, only the trigger events in the first commutation interval (\( t_2-t_0 \)) in Fig. 3-13 will be considered. The same principal applies to the trigger events in the second interval (\( t_5-t_3 \)).

If IGCTs are used as switches for S1 and S2, they must always be turned off after the triggering of the bidirectional switch (at \( t_0 \)) to avoid any disruption in the source current. If thyristors are used for S1 and S2, a time interval equal to the turn-off time (\( t_q \)) should elapse before the resonance could begin and a current path should always exist for the source current.

As the capacitor C discharges and just before the forward biasing of the switch (T2 at \( t_2 \)), one thyristor of the half-bridge may be triggered. As shown in Eq (3-55), the inductor current at the end of a half cycle of resonance (\( I_L \)) will be different from its initial value (\( I_0 \)) by \( \Delta I_L \). If the output current (\( I_{Out} \)) remains constant during this interval, the commutation between the two thyristors (T2 and T4 for the transition interval 1) will depend on \( \Delta I_L \) only. In the presence of ripple in the output current, the analysis of commutation time (between T2 and T4 for transition interval 1) will rather be a function of both \( \Delta I_L \) and the ripple in the output current (\( \Delta I_{Out} \)).

3.7 Performance of the RPR-Converter

In chapter two, the limitations of the diode rectifier to extract power from a source with high inductance were shown. In this section, the capability of the RPR converter topology to extract power will be explored.
3.7.1 Power Extraction

Assuming constant load current \( (I_{\text{Out}}) \) and no losses, we can approximate the output power of the converter as follows:

\[
P_{\text{out}} = \frac{T}{2\pi} \int_0^T \left( v_{\text{out}}(t) - I_{\text{out}}(t) \right) dt = \frac{1}{2\pi} I_{\text{out}} \int_0^{2\pi} E_{\text{out}} \sin(\omega t) d(\omega t)
\]

\[
= I_{\text{out}} \int_{\theta_f}^{\pi-\theta_f} \sin(\omega t) d(\omega t) = \frac{2}{\pi} I_{\text{out}} E \cos(\theta_f)
\]

Substituting \( I_{\text{Out}} \) from Eq. (3-81) we have:

\[
P_{\text{out}} = \frac{2}{\pi} E I_{Lg,\text{nom}} \cos(\theta_f)
\]

(3-91)

Increasing \( \theta_f \) will lower the distortion factor (DISF) of the source current (represented by the term \( \sqrt{1 - \left( \frac{4}{3} \frac{\theta_f}{\pi} \right)} \)) and decrease the output voltage (represented by the term \( \frac{2}{\pi} E \cos(\theta_f) \)).

For an ideal diode rectifier with constant load current and ideal source, the average output power extracted would be expressed by:

\[
P_{\text{out(Diode-Rectifier)}} = \frac{2}{\pi} E^* I_N
\]

(3-92)

If we normalize the output power of the RPR converter to that of the ideal diode rectifier we have:
\[
P_{\text{Out (Norm)}} = \frac{\cos(\theta_f)}{\sqrt{1 - \left(\frac{4 \theta_f}{3 \pi}\right)}}
\]  \hspace{1cm} (3-93)

Fig. 3-14 shows the normalized output power (Eq. 3-93) versus firing angle.

As could be seen, for a firing angle smaller or equal to 26°, the capability of the new rectifier to extract power will be superior to that of an ideal diode rectifier. With a firing angle equal to 13.5°, the generator could be 2.5% underrated compared to a diode rectifier (with no source inductance). Higher firing angles may also be applied to reduce \( V_{c_{\text{max}}} \) and the rating of the converter, however, in that case, Eq. (3-93) will not be a good approximation for the output power since the effect of damping, and the resulting increase in commutation time (as discussed in section 3.6.2), should be also considered.

### 3.7.2 Total Current Harmonic Distortion (TCHD)

Assuming unity displacement factor (DPF), we could write:
where $I_1$ is the rms value of the fundamental component of the input current and $P_g$ is the active power output from the source. Assuming no losses in the source and rectifier, we have:

$$P_{\text{out}} = P_g$$

(3-95)

So using Eq. (3-91), we could write:

$$\frac{I_1}{I_{Lg,\text{nom}}} = \frac{2\sqrt{2}}{\pi} \frac{\cos(\theta_f)}{\sqrt{1 - \left(\frac{4\theta_f}{3\pi}\right)}}$$

(3-96)

The TCHD is defined as:

$$TCHD = \sqrt{1 - \left(\frac{8\cos^2(\theta_f)}{\pi^2 - \frac{4\theta_f}{3\pi}}\right)} \times 100\%$$

(3-97)

So we could write:

$$TCHD = \sqrt{1 - \left(\frac{8\cos^2(\theta_f)}{\pi^2 - \frac{4\theta_f}{3\pi}}\right)} \times 100\%$$

(3-98)

### 3.8 Loss analysis

In the section, analytical formulas for calculating the losses in the proposed topology will be developed. They will be used to predict the losses of the rectifier and later will be compared to those of a PWM rectifier.

#### 3.8.1 Loss Analysis of the RPR-Converter

Unlike the PWM converter, loss calculation in the new topology is a rather simple task, since the losses will be limited to conduction losses. The losses could be classified as:
1. Rectifier losses ($P_R$) including the losses in the thyristor half-bridge ($P_{Thy}$) and diode half-bridge ($P_D$)
2. Bidirectional switch losses ($P_{BSW}$)
3. Capacitor losses ($P_C$)
4. Losses in the forced commutation circuit in the case where thyristors are used for S1 and S2 ($P_{FCC}$)

So the total losses could be written as:

$$P_{Loss} = P_{Thy} + P_D + P_{BSW} + P_C + P_{FCC}$$ \hfill (3-99)

The latter three are small compared to rectifier losses (for small values of $\Theta_t$), however all will be analyzed in detail in the next section ($P_{FCC} = 0$ if IGCTs are used for S1 and S2).

### 3.8.1.1 Rectifier Losses

For low operating frequencies, thyristor and diode conduction losses will dominate all other rectifier losses and switching losses will be ignored. Conduction losses in a switch could be expressed as follows:

$$P_{SW} = \frac{1}{T} \int_0^T v_F(t) i_{on}(t) \, dt$$ \hfill (3-100)

where $v_F$ and $i_{on}$ are the instantaneous forward voltage and current in the switch. Some manufacturers like ABB provide an “on-state characteristic model” for their thyristors and diodes which could be used for a precise loss calculation, otherwise the linear model for the thyristors and diodes, will give a very reasonable accuracy:

$$V_T = V_{T0} + i_{on}(t) r_{on}$$ \hfill (3-101)

$$V_F = V_{F0} + i_{on}(t) r_{on}$$ \hfill (3-102)

where $V_{T0}$ is the threshold voltage and $r_{on}$ is the on-state resistance of the switch.
For a constant load current, the losses in the thyristor half bridge could be calculated as follows:

\[
P_{\text{Thy}} = 2 \frac{1}{2\pi} \int_{\theta_f}^{\pi-\theta_f} (V_{T0,\text{Thy}} + I_{\text{out}\,\text{on,\text{Thy}}} I_{\text{out}}) d\theta
\]

\[
= \frac{\pi - 2\theta_f}{\pi} (I_{\text{out}} V_{T0,\text{Thy}} + I_{\text{out}}^2 r_{\text{on,\text{Thy}}} )
\]  

(3-103)  

(3-104)

where the subscript \text{Thy} represents the thyristors in the controlled half-bridge.

The conduction losses for both diodes during resonance will be:

\[
P_{D(\text{res})} = 4 \frac{1}{2\pi} \int_{-\theta_f}^{\theta_f} (V_{F0,\text{Diode}} + I_{\text{out}\,r_{\text{on,\text{Diode}}}} I_{\text{out}}) d\theta
\]

\[
= \frac{4\theta_f}{\pi} (I_{\text{out}} V_{F0,\text{Diode}} + I_{\text{out}}^2 r_{\text{on,Diode}} )
\]  

(3-105)  

(3-106)

where (\text{Diode}) represents the diodes in the diode half-bridge.

During normal conduction mode, the loss equation for the diodes will be similar to Eq. (3-104), so the total diode losses will be:

\[
P_D = \frac{\pi + 2\theta_f}{\pi} (I_{\text{out}} V_{F0,\text{Diode}} + I_{\text{out}}^2 r_{\text{on,Diode}} )
\]  

(3-107)

and the conduction losses in the rectifier is:

\[
P_R = (\frac{\pi + 2\theta_f}{\pi}) (I_{\text{out}} V_{F0,\text{Diode}} + I_{\text{out}}^2 r_{\text{on,Diode}} )
\]

\[
+ (\frac{\pi - 2\theta_f}{\pi}) (I_{\text{out}} V_{T0,\text{Thy}} + I_{\text{out}}^2 r_{\text{on,\text{Thy}}} )
\]  

(3-108)
3.8.1.2 Losses in the bidirectional switch

The conduction losses in the bidirectional switch could be expressed as:

\[ P_{BSW} = 2(V_{T_0,BThy}I_{(avg)_{BThy}} + r_{on,BThy}I_{(rms)_{BThy}}^2) \]  (3-109)

where the subscript BThy represents the thyristors in the bidirectional switch.

The average and RMS currents in the bidirectional switch have already been calculated in equations (3-76) and (3-75), so we have:

\[ P_{BSW} = \frac{2\theta_f}{\pi} I_{out}^2 \left( \frac{V_{T_0,BThy}}{\pi I_{out}} + \frac{1}{2} r_{on,BThy} \right) \]  (3-110)

3.8.1.3 Capacitor Losses

Losses in the capacitor will be:

\[ P_C = I_{C_{(rms)}}^2 ESR_C \]  (3-111)

so using Eq. (3-87), we have:

\[ P_C = I_{out}^2 \frac{\theta_f}{\pi} ESR_C \]  (3-112)

where ESR_C is the Equivalent Series Resistance of capacitor C.

3.8.1.4 Losses in the forced commutation circuit

The losses in the forced commutation circuit of Fig. 3-7 will be the total of losses in the IGBTs (P_{Tr}) and in the diodes (P_{Da}) and could be calculated as follows:

\[ P_{Tr} = 2 \frac{1}{T} \int_0^{t_s} I_{out}(V_{CE0,IGBT} + I_{out}r_{on,IGBT}) dt \]
= \frac{2}{T} \left( I_{\text{out}} V_{\text{CE0,IGBT}} + I_{\text{out}}^2 r_{\text{on,IGBT}} \right) \tag{3-113}

, and

P_{\text{FCC}} = \frac{2}{T} \left( I_{\text{out}}^2 (V_{\text{CE0,IGBT}} + V_{\text{T0,DA}}) + (I_{\text{out}}^2 (r_{\text{on,IGBT}} + r_{\text{on,DA}})) \right) \tag{3-115}

So the total losses in the forced commutation circuit will be:

\begin{align*}
P_{\text{FCC}} &= \frac{2}{T} \left( I_{\text{out}}^2 (V_{\text{CE0,IGBT}} + V_{\text{T0,DA}}) + (I_{\text{out}}^2 (r_{\text{on,IGBT}} + r_{\text{on,DA}})) \right) \\
\text{where } V_{\text{CE0}} \text{ is the threshold collector-emitter voltage of the IGBT.}
\end{align*}

As expected, as \( \theta_t \) increases, the losses in the thyristor half-bridge will decrease, while losses in the diode half-bridge, bi-directional switch and capacitor will increase. Increasing output current will have a more pronounced effect on the capacitor loss. The share of each component in the total losses will be illustrated in the design example in section 3.8.2.

3.8.2 Loss calculation in the MW range

The Small wind turbines (below 100kW) could not compete with other sources of electricity generation, and comprise only a very small share of the installed wind power capacity in the world. The MW range wind turbines, on the other hand, are a very competitive source of electricity and have a growing share of the market. That is why the loss calculation for the proposed topology in this section will be based on data from a 3 MW turbine design. The losses will be analytically calculated and then will be compared to those of a PWM rectifier.

3.8.2.1 Loss calculation for the RPR-Converter topology

For the purpose of this analysis, the loss formulas of section 3.8 have been used to calculate the semiconductor losses for a generator-rectifier in the MW range. The 3MW turbine design discussed in chapter two with a rotor diameter of 90 m and rated
mechanical rotor power equal to 3.3 MW was considered for the analysis [8]. The power curve along with other turbine characteristics were already shown in Fig. 2-2.

The mechanical and electrical losses in the generator were not taken into account. It was assumed that the turbine power will be equally divided between three transverse flux machines sharing the same shaft. The following specifications for each machine were assumed:

\[ E_g = \frac{690}{\sqrt{3}} \] @ 16 rpm

\[ X_g = 2 \text{ p.u.} \]

\[ p = 188 \] (number of pole pairs)

ESR of the capacitor was assumed to be 1 mΩ which is a reasonable value for this type and size of capacitor (MKP type capacitors from Vishay were examined [41]).

Knowing the desired power curve, the current curve of the machine versus wind speed was established (Fig. A-1 (b), Appendix A). The wind speed interval considered was from 3 to 12.5 m/s (Cut-in speed - rated wind speed). Output power versus wind speed is shown in (Fig. A-1 (c), Appendix A).

IGCTs were used as switches for the controlled half bridge to simplify the design. Appropriate diodes (as reverse blocking and half-bridge diodes), and thyristors (for the bidirectional switches) were also selected whose specifications are given in the table A-1 in Appendix A. The IGCTs, their reverse blocking diodes and the two freewheeling diodes were overrated mainly due to the very limited choice of switches at these ratings particularly for the IGCTs.

The following characteristics are shown in figures 3-15 (a), (b), and (c) in the following sequence:

(a) Rectifier losses versus wind speed
(b) Capacitor peak voltage versus wind speed (unsaturated inductance)
(c) Firing angle versus rotor speed
Fig. 3-15 (a) Rectifier losses versus wind speed, (b) Capacitor peak voltage versus wind speed (unsaturated inductance), (c) Firing angle versus rotor speed for the 1 MW IGCT version of the proposed topology as discussed in section 3.8.1
Fig. 3-16 shows the efficiency of the rectifier versus wind speed.

As could be seen, the efficiency will be high for the whole range of wind speeds in contrast to PWM rectifiers where efficiency falls to very low values as the wind speed decreases. This will be further discussed in the next section as PWM rectifier losses are considered.

The efficiency reaches its peak value when generator speed reaches its nominal value (16 rpm) at rated wind speed (10 m/s). Fig. 3-17 shows the distribution of losses in the rectifier at this point.

The share of bidirectional switch and capacitor losses will drastically decrease at lower output powers, where the resonance interval will be smaller compared to normal conduction interval. Fig. 3-18 shows the loss distribution for a rotor speed half its rated value.
Capcitor, 287 W  ■ BSW, 482 W, □ Rectifier, 5498 W

Loss Distribution

Fig. 3-17 Loss distribution at rated wind Speed (10m/s), $P_{out}=597.671$ kW

Capcitor, 8 W  ■ BSW, 37 W, □ Rectifier, 1169 W

Loss Distribution

Fig. 3-18 Loss distribution at 5 m/s, $P_{out}=74.821$ kW
3.8.2.2 Losses in a PWM converter

For the purpose of comparison, it was assumed that the same power from the turbine-generator is equally distributed among three conventional 3-phased two-level PWM VSI converters. The 3-phased generators have the same line-neutral EMF voltage versus rotor speed, and the same reactance. Rotor characteristics are also assumed to be similar to that of section 3.8.2.1.

Converter loss calculation in PWM converters has been discussed in many papers (among others in [37], [38], and [39]). Such calculation methods use the datasheet specifications for energy loss per pulse in the IGBT ($E_{on}$, $E_{off}$) and diode turn-off energy dissipation at nominal ratings to calculate the switching losses at all operating points. For the purpose of this analysis, however, PCIM simulator [32] was employed which uses a polynomial function of the IGBT and diode switching energy. With this simulation method, it is also possible to compute static and transient losses and temperature rises in IGBT modules (with user's choice of heat-sinks) in a two-level voltage source inverter topology. This will result in a very realistic loss calculation method.

For the simulation tool, the appropriate IGBTs and 9K/kW water-cooled heat-sinks were selected so that for the operating point with highest dissipation, the junction temperature will be limited to 122° C. The switching frequency was chosen to be a rather low value (1000Hz) to lower the switching losses. Only semiconductor losses were considered.

Knowing the generator-turbine characteristics, the required operating point parameters for the PCIM simulator were determined and the switching and conduction losses for each element (Diode-IGBT) were calculated at each wind speed. Fig. 3-19 shows the efficiency of the PWM-rectifier versus wind speed. As could be seen, efficiency remains rather constant if the rotor speed is kept constant (equivalent to wind speeds from 10 to 12.5 m/s), however, as rotor speed decreases (equivalent to wind speeds below 10 m/s), the efficiency will decrease rapidly.
Fig. 3-19: Efficiency of the PWM rectifier versus wind speed

Fig. 3-20 (a) and (b) show the share of IGBT and diode switching and conduction losses in the total losses for two operating points (highest and lowest efficiencies).

As could be seen, while both conduction and switching losses will decrease at low power, the share of switching losses will increases dramatically. This clearly explains the low efficiencies at low wind or rotor speeds.

In the case of thyristors with external commutation circuits (for the controlled half-bridge), higher efficiencies could be expected. The IGCT RB-Diode combination was replaced with ABB phase control thyristors with similar ratings. The losses in the forced commutation circuit were not considered since the circuit already presented in Fig. 3-8 will not be practical at these ratings. This is because, unlike thyristors, IGBTs do not have high pulse current ratings (see section 3.4.1).
Fig. 3-20: PWM converter loss distribution for (a) cut-in speed (3 m/s), $P_{Out}=20.28$ kW, (b) rated wind speed (10 m/s), $P_{Out}=620.69$ kW. Losses indicated are for one switch (IGBT+Diode).
The results for different calculations are compared in Fig. 3-21. The consequences on the cost of the energy produced will be clearer in the next section when the total annual energy production of the two topologies (PWM and the RPR rectifiers) are compared.

![Diagram of efficiency comparison]

Fig. 3-21 The efficiency of the PWM rectifier compared with the IGCT and Thyristor versions of the Proposed Topology

### 3.9 Economical Aspects

In this section, the annual energy production of the same turbine with RPR and PWM converters will be calculated and compared to the extra investment necessary for the RPR converter to see if the latter, despite its high switch ratings, is economically competitive compared to a PWM-rectifier.
3.9.1 Energy Savings

The same power curve and the two sets of efficiency characteristics for the RPR and PWM converters (for the 3.3 MW turbine of section 3.8.2) were used to calculate the annual energy production of the turbine.

We will first need the Rayleigh distribution to approximate the wind speed probability [34]:

\[ p(v) = \frac{v}{2u_a^2} \exp\left(-\frac{v^2}{4u_a^2}\right) \]  

(3-116)

where \( p(v) \) is the Rayleigh probability density function, \( v \) is the wind speed in m/s, \( u_a \) is the long term average wind speed in m/s.

An average annual wind speed equal to 7 m/s was used. Fig. 3-22 (a) shows the Rayleigh distribution for wind speeds up to 18 m/s.

The results of the calculation at each wind speed is shown in Fig. 3-22 (b). The amount of energy saved with the new topology at each wind speed is shown in Fig. 3-22 (c). The total energy savings was found to be 102252 kWh/ year.

3.9.2 Semiconductor Prices

As shown in section 3.5.1, the switches in the controlled half-bridge are exposed to high voltages during the resonance interval, so they will have higher ratings than those of a PWM converter and their prices will be higher (at least for the IGCT version). For the purpose of evaluation, three 1MW rectifier units (whose losses were already calculated) were used.
Fig. 3-22: (a) Rayleigh distribution for an average wind speed of 7 m/s, (b) annual energy production of the turbine at each wind speed, (c) energy saved with the new topology at each wind speed per year.
For the proposed topology, the IGCT version was used which is less efficient and costs more than the thyristor version, however it does not need a forced commutation circuit and due to its simplicity, it can be easily realized and evaluated. The semiconductor prices for this topology were obtained directly from ABB. The price for the capacitor is obtained from [36] for a similar type of capacitor with similar KJoule rating. For the PWM converter the IGBT prices are calculated on a 170$ per MVA basis (per IGBT) and were obtained from [35]. The evaluations are very rough estimates and are intended to give a very general idea of the extra investment required. The filtering and heat-sink requirements were not compared though they will much different in the two topologies. For a precise evaluation, the whole energy conversion system should be considered.

Table 3.3 compares the semiconductor prices for the two topologies. As could be seen the extra investment required is less than 9 percent of the economies made.

<table>
<thead>
<tr>
<th>Semiconductor prices (PWM), US$</th>
<th>15147</th>
</tr>
</thead>
<tbody>
<tr>
<td>Semiconductor (IGCTs+Diodes)+ Capacitor prices (the RPR converter Topology), US$</td>
<td>26751</td>
</tr>
<tr>
<td>Extra Investment Required, US$</td>
<td>11603</td>
</tr>
<tr>
<td>Energy Saving in 20 Years, kWh</td>
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</tr>
<tr>
<td>Energy Saving in US$ @0.07$/kWh</td>
<td>123269</td>
</tr>
</tbody>
</table>

3.10 Simulation of the RPR-Converter

The numerical calculation presented in section 3.8.1 is based on simplifications which will be useful for a rapid approximation in an Excel worksheet. For a more precise analysis, the source-converter circuit was simulated using Matlab Simulink. Fig. 3-23 shows the simulation model.
3.10.1 Source Simulation

In order to simulate the source, a saturable inductance block was built in Simulink as shown in Fig. 3-24.

Fig. 3-23: simulation model of the RPR-Converter

The forced commutation circuit (FCC) block shown in the model, is that of Fig. 3-8 already explained in section 3.4.1. The trigger control block generates trigger pulses with the proper timings (as explained in section 3.6), whose trigger waveforms will be presented later in Fig. 3-27.

As mentioned in section 2.2, under nominal conditions, core saturation will occur in a TFPM machine. That is why the source reactance has been modeled with a non-linear inductance.
As could be seen, it is essentially a controlled current source whose current is a function of flux linkage($\psi$) as implemented by the *Look-Up table* $i=f(\psi)$. The integrator block uses the measured voltage across the inductor terminal as input and outputs the flux. Since the initial conditions can be defined in the Integrator block of Simulink, the changes in flux linkage (due to permanent magnet only) could also be considered in the model. In this way the variation in inductance due to rotor position may also be included (though it was not considered in the present simulation).

To build the look-up table for the saturable inductor model in Simulink, the experimental results from the DC test method on a TFPM machine, which is not discussed here, were used (similar to that mentioned in [25]). The Current-Flux characteristic of the saturable inductance model is shown in Fig. 3-25.

![Current-Flux characteristic](image)

Fig. 3-25 The Current-Flux characteristic of the saturable inductance model
3.10.2 Simulation Results

To simulate the generator-converter, the following parameters (similar to those of an experimental machine), were used:

\[ E_g = 15 \text{Vrms} @ 60\text{Hz} \]
\[ I_{\text{nom}} = 6.8 \text{ A} \]
\[ X_{Lg} = 9.8 \Omega @ 60\text{Hz} = 4.44 \text{ pu (unsaturated)} \]
\[ R = 0.05 \Omega = 0.023 \text{ pu} \]
\[ C = 2.587 \mu \text{F} \]
\[ \text{ESR}_C = 14 \text{ m}\Omega \]
\[ \text{Firing angle } (\theta_f) = 8.8^\circ \]

Diodes (D1, D2), datasheet values for a 600V, 15A diode:
\[ V_{F0 (D1, D2)} = 0.5 \text{ V} \]
\[ r_{\text{on (D1, D2)}} = 18 \text{ m}\Omega \]

Thyristors (T1, T2), datasheet values for a 800V, 16A thyristor:
\[ V_{T0 (T1-T4)} = 0.8 \text{ V} \]
\[ r_{\text{on(T1-T4)}} = 45 \text{ m}\Omega \]
\[ t_q (\text{FCC}) = 100 \mu \text{Sec.} \]

The parameters used were defined in section 3.4. No snubber circuit was included.

An interesting figure showing the capability of the suggested converter and the limitations of a diode rectifier to extract the maximal power out of source with high inductance, is that of the Fig. 3-26. The base values chosen for \( P_{\text{out}} \) and \( I_{\text{out}} \) are the output current and voltage of an ideal diode rectifier with constant load current and no source inductance, i.e.:

\[ I_{\text{out}, \text{base}} = I_{Lg, \text{nom}} \]
\[ P_{\text{out}, \text{base}} = \frac{2}{\pi} E I_{Lg, \text{nom}} \]
Fig. 3-26: Output power vs. output dc current. The RPR-converter (line), diode rectifier (dotted), \( X_{lg} = 4.47 \) pu

The simulation waveforms for the RPR-converter at nominal power are shown in figures 3.27 and 3.28. The waveform for output voltage in Fig. 3-27(b) is almost that of a diode rectifier with no source inductance. The small voltage distortion could be explained by the fact that \( i_{lg} > i_{out} \) during the transition interval 1 (as discussed in section 3.4.2), which means that capacitor C will continue to conduct until \( i_{lg} \) equals \( i_{out} \). Fig. 3-27 (c) and (d) show the output and source currents respectively. The capacitor waveform and the gate trigger signals are shown in Fig. 3-28 (b), (c) and (d). T3 and T4 could be triggered simultaneously, since at each trigger point only one has the right conditions for conduction.

The waveforms in figures 3-27 and 3-28 represent a condition where the inductor is still unsaturated. To show the effect of the saturation, the output current was increased so that the inductor enters well into the saturation region (the inductance varies between 26mH to 8mH) and the reactance of the inductor will be the very exaggerated value of 10.81 pu (for \( I_{nom (rms)} = 16.55 \) A). Fig. 3-29 shows the source currents, source EMF voltage and capacitor voltage for such a condition.
Fig. 3-27: (a) source EMF voltage ($e_g$), (b) output voltage ($v_{out}$), (c) output current ($I_{out}$), and (d) source current ($i_{Lg}$), for the RPR converter.
Fig. 3-28: (a) Source EMF voltage ($e_g$), (b) Source current ($i_{Lg}$), (c) Capacitor voltage ($v_c$), and (d) Gate trigger signals for the RPR converter
Fig. 3-29: (a) EMF voltage ($e_g$), source current($i_{Lg}$), (b) capacitor voltage($v_C$) for the RPR converter, $X_g=10.81$ pu, $I_{nom\,(rms)}=16.55$ A

To show the effect of the saturation on resonance period and $V_{Cmax}$, the inductance model was used in a RLC circuit (as shown in Fig. 3-30) with the following parameters:

- $R_g=.05$ mΩ
- $C=2.587 \mu F$
- $ESR_C=14$ mΩ

The circuit was simulated in Simulink and $T_0/2$ and $V_{Cmax}$ were measured for various initial inductor currents. The results are shown in Fig. 3-31.
Fig. 3-30: Test circuit

The Fig. 3-31 (a) could be used to predict the VRRM and VDRM of the half-bridge thyristors and Fig. 3-31 (b) could be used as a look-up table to predict the resonance period.
As could be seen from the waveforms of Fig. 3-29, saturation of the inductor not only does not result in any distortion in the source current (as in the case of passive compensation), on the contrary, due to saturation, $V_{C_{\text{max}}}$ will be lower than its value from Eq. (3-79) (for the unsaturated inductance). This could be considered an important advantage of the RPR-converter.

3.11 Experimental Results

In order to validate the simulation results in section 3.10, an experimental model of the RPR-converter was built which is shown in Fig. 3-32.

![Fig. 3-32: The experimental circuit](image)

**Source**

An autotransformer in series with an inductance was used to model the transverse flux machine. One phase of a 3-phase autotransformer with a power rating of 7 kVA was used. The source inductance was made of two 13.8 mH parallel laminated core inductances.
Converter
Thyristors were used in the controlled half-bridge (for T1 and T2), and the forced commutation circuit was identical to that of Fig. 3-8.

The eZdsp F2812 stand-alone DSP (Digital Signal Processor) board was used to generate the trigger pulses for the controlled switches. The synchronization with source frequency was implemented using an simple optocoupler based zero-crossing detector. To improve noise immunity, the input pulses to gate drivers were optically isolated. All gate drivers were also isolated from their respective switches either through pulse transformers or opto-couplers.

The following parameters were measured:

\[ E_{g\ (no\ load)} = 15.56 \text{ Vrms, } 60 \text{ Hz} \]
\[ L_g = 13.8 \text{ mH (equal to } 2.36 \text{ pu for } I_{\text{nom}} = 7.07 \text{ A}) \]

The equivalent impedance of the autotransformer was measured by the short circuit test and was found to be less than 0.03 pu and was ignored.

The capacitor used was a 4.7 \( \mu \)F, 630V metallized polypropylene capacitor (MKP type). Its ESR at the resonance frequency was measured to be 14.7 m\( \Omega \). The firing angle was programmed to \( \theta_f = 8.4^\circ \).

Fig. 3-33 shows the waveforms for Source voltage \( (e_g) \) and source current \( (i_{Lg}) \). The output voltage \( (v_{Ou}) \) along with source current is shown in Fig. 3-34. The same waveforms are shown along with converter input voltage \( (v_g) \) in Fig. 3-35. The input current is 7 A and the source reactance is equal to 2.36 pu.

The efficiency of the source was estimated to be around 70%. Due to high core losses in the laminated core inductance \( (L_g) \), and low input voltage, \( \Delta I_{Lg} \) in Eq. (3-55) will not be
negligible. To avoid additional commutation time, smaller output inductor was used. That is why current waveform is not a totally rectangular form. In the case of a TFPM machine the efficiency will be much higher and the damping effect will be much lower.

As could be seen the waveforms are quite similar to those of simulation.

### 3.11.1 Efficiency

The efficiency of the converter was measured using two power analyzers: PA4400 from AVPower and Norma D6000 from LEM.

The following parameters were used for the test:

\[ E_{g(no\ load)} = 15.56 \text{ Vrms}, 60 \text{ Hz} \]
\[ L_g = 13.8 \text{ mH} \]
\[ \Theta_f = 8.4^\circ \]
\[ L_{out} = 10 \text{ mH} \]

The following values were measured:

\[ I_{L_g} = 6.005 \text{ A} \]
\[ I_{out} = 6.123 \text{ A} \]
\[ P_{in} = 57.90 \text{ W} \]
\[ P_{out} = 44.81 \text{ W} \]
\[ E_i = 5.512 \text{ V}_{(dc)} \]
\[ P_{EI} = 0.257 \text{ W} \]

where \( P_{in} \) and \( P_{out} \) are the power measured at the input and output terminals of the converter respectively, \( E_i \) is the auxiliary supply in the forced commutation circuit as shown in Fig. 3-8. The values for input and output power are the results of 6 minute integration. Results for both instruments are included in Appendix B.
Fig. 3-33 Waveforms for Source voltage ($e_g$, CH1), source current ($i_{Lg}$, CH2)

Fig. 3-34 Waveforms for converter source current ($i_{Lg}$, CH2), and converter output voltage ($v_{out}$, CH3), $X_g=2.36$ pu
The total input power to the converter will be the sum of $P_{in}$ and $2P_{EI}$, so the efficiency could be calculated as:

$$\eta = \frac{P_{out}}{P_{in} + 2P_{EI}} = 76.71\%$$

As could be seen the auxiliary supply required in the forced commutation circuit is very small compared to input power ($\frac{P_{EI}}{P_{in}} = 0.0088$).

Table 3.4 presents the results for numerical calculation for the same output power and output current (with formulas developed in section 3.8), showing the loss distribution in
different components of the RPR-converter. The deviation from the measured value for efficiency could be explained by the high ripple in the output current and the fact that thyristor and diode model parameters are only approximate values due to unknown junction temperature.

Table 3.4

<table>
<thead>
<tr>
<th>D-HB</th>
<th>Thy-HB</th>
<th>BSW</th>
<th>Capacitor</th>
<th>FCC</th>
<th>η</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.84 W</td>
<td>6.67 W</td>
<td>0.39 W</td>
<td>0.02 W</td>
<td>0.26 W</td>
<td>78.62%</td>
</tr>
</tbody>
</table>

In order to compare the results with those of a diode rectifier with no (or rather negligible) source inductance (the inductance of autotransformer ≈ 0.1 mH), the thyristor half-bridge was replaced with a diode half-bridge similar to the one already used. The input voltage was adjusted to achieve the same output power. The following values were measured:

\[ P_{in} = 56.18 \text{ W} \]
\[ P_{out} = 44.14 \text{ W} \]

Efficiency (η) = 78.57%

As could be seen the difference was found to be less than 2 percent.

**Discussion**

The simulation and the experimental results introduced in this chapter were quite similar and confirm that the proposed topology could be used with sources with high values of saturable source reactance like TFPM machine. It was shown that the capability of the new topology to extract active power could be better than a diode rectifier with no source inductance. It was also shown that the proposed topology is far superior to a PWM two-level rectifier at low loads and considerably more efficient at all load and frequency conditions (with a maximum efficiency close to 99% for the 3MW turbine considered).
Despite the lower switch utilization, it was made clear that the economy in energy due to low losses is obviously more important than the extra initial investment required which means that its use in MW wind energy conversion systems is quite justified.
Conclusion

An uncontrolled bridge-rectifier for constant voltage and current loads will seriously limit the power output of a TFPM machine. On the contrary, the uncontrolled rectifier with shunt compensation was found to be an interesting rectifier choice for a wind turbine application due to fact that it will be highly efficient and could totally compensate the large unsaturated source inductance at nominal conditions.

For a variable speed machine, the power delivery capability of the shunt compensated rectifier will decay as the speed falls below the nominal value for which the shunt capacitor was calculated. However, in a wind turbine application it will be still capable of delivering the required (optimal) output power in the whole speed range.

In the case of saturation, the power delivery capability of a shunt compensated rectifier will be only slightly affected. Despite saturation, high efficiencies could be attained and the source current will have low distortion during the whole speed range even at nominal conditions where saturation has a more pronounced effect.

The series compensation is not recommended due to lower efficiencies, lower power delivery capability and the larger value for the capacitor compared to shunt compensation.
Though, shunt compensation is highly recommended for a wind turbine application for which power requirements are lower at lower speeds, it will not be the proper choice where output power should be proportional to machine speed.

For a Diode-bridge rectifier with constant load current, passive compensation is not a practical solution and a shunt compensated VSR with DC-link inductance could be a better substitute, though the addition of an inductor will imply additional losses and a higher price.

The PWM rectifier could be used as a second choice for a TFPM machine, however, for a load with constant current, the PWM-CSR has high conduction losses. This means that none of the existing solutions could provide us with a high efficiency current source rectifier. That is why in chapter 3, a new CSR topology was introduced which, compared to shunt compensated rectifier, will improve the power transfer at all speeds and its performance is not affected by saturation.

A detailed analysis of the suggested RPR-converter was conducted and a design example in MW range was used to compare the losses in a RPR-converter with those of a PWM-VSR (which has smaller losses than a PWM-CSR). The results showed that the RPR rectifier will be more efficient than a PWM rectifier. Besides, the analysis of loss distribution in the RPR-converter showed that the losses will be mainly in the rectifier half-bridges, which means that the total losses will be close to those of a diode-rectifier if thyristors are used for the controlled half-bridge.

Compared to a PWM-rectifier, the RPR-converter has higher switch ratings (in the controlled half-bridge), however, the price estimation for the semiconductors showed that the choice of RPR-converter for high-current sources will be quite justified if energy savings due to higher efficiencies are considered.
A simulation model was used to study the functionality of the RPR-converter with saturable source inductance. The results were found to be in good agreement with those of the experimental model.

**Suggestions for further work**

1. The use of the proposed topology for multi-phase sources should be investigated.
2. A low loss, low priced, thyristor based FCC for a MW range RPR-converter should be designed.
3. A more comprehensive investigation of the TFPM machine with saliency and for various EMF voltage waveforms (other than sinusoidal) should be conducted (for both RPR-converter and shunt compensated bridge-rectifier).
4. The price analysis should become more elaborated.
References


[32] P.J. Van Duijsen, P. Bauer, U. Killat, " Realistic benchmarking of IGBT-modules with the help of a fast and easy to use simulation-tool ", PCIM 04, Nurnberg, revised 01.07.05.


Fig. A-1 The assumed EMF voltage, current, and power curves of the machine versus wind speed (Cut-in speed to rated wind speed) for a 1 MW rectifier as discussed in section 3.8.1.
<table>
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<tr>
<th>Description</th>
<th>VRRM (V)</th>
<th>VDRM (V)</th>
<th>IFAVM (A)</th>
<th>VT0, VF0 (V)</th>
<th>$r_{on}$ (mΩ)</th>
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<td>IGCT for S1, S2</td>
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<td>1.19</td>
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<td></td>
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</tbody>
</table>
Appendix B

(a)

(b)

Fig. B-1 Efficiency measurements for:
(a) 6 minute integration with PA4400
(b) 5 minute integration with Norma D6000.