ABSTRACT

A GAN BASED DUAL ACTIVE BRIDGE CONVERTER TO INTERFACE ENERGY STORAGE SYSTEMS WITH PHOTOVOLTAIC PANELS

by Hassan Athab Hassan

Renewable energy sources are intermittent in nature. Therefore, they are often used in conjunction with energy storage systems to ensure the continuous flow of power. Bidirectional DC/DC converters serve as an intermediary between the energy storage hardware and the inverter connected to the grid. Dual active bridge (DAB) converters are well suited for this application, especially when the power is greater than 1 kW. They offer bi-directional power flow and can achieve zero-voltage-switching (ZVS). Furthermore, gallium nitride (GaN) based transistors have many attributes that enable them to outperform silicon (Si) MOSFETs in this topology. This research compares the performance of a DAB converter designed with GaN transistors to one built with Si MOSFETs. A hardware prototype was built for 48 V to 400 V conversion with a rated power of 4 kW.
A GAN BASED DUAL ACTIVE BRIDGE CONVERTER TO INTERFACE ENERGY STORAGE SYSTEMS WITH PHOTOVOLTAIC PANELS

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Dedicated

to my inspirational parents
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Abstract

In photovoltaic systems, energy storage systems are portrayed as the quintessential solution for balancing fluctuating energy generation and managing the amount and the direction of the supplied power. Bidirectional converters (BDC) have been used for transferring electrical energy in two directions. The performance of the selected bidirectional converter is a vital factor that should be taken into consideration in advance of combining ESSs with renewables. Due to the merits of dual active bridge (DAB) converters, they are the more suitable converters to be exploited in such applications especially with power greater than 1 kW. A large number of the switches along with other drawbacks reduce the converter efficiency. The attributes of Gallium Nitride-based transistors have entitled them to be used in DAB converters in order to improve their performance since the silicon-based transistors attained their theoretical limits. Theoretically, GaN transistors such as high electron mobility transistors (HEMTs) can operate at higher frequencies and temperatures than Si. The size of the transformer, which sustains the galvanic isolation, and other passive elements (inductors and capacitors) shrink with increased frequency. The small on-resistance with small drain and gate to source capacitance in addition to the absence of the reverse recovery charge of HEMTs, contribute in minimizing the size of the heatsink. As a corollary, the compactness and high efficiency are achieved. Notwithstanding the advantages of GaN, the prevailing challenges are the parasitic inductance of the PCB in both the driving loop and the power loop and the Miller capacitance. The objective of this work is to evaluate the performance of DAB converter with GaN HEMTs and compare it with Si MOSFET based converter. Comparisons among the most accepted isolated converters are included. In addition, this work covers the principles of operation of HEMTs and the drawbacks of GaN-based transistors.
CHAPTER ONE

1. Introduction

1.1 Overview of Photovoltaic Systems

In 18th to the 19th centuries, the world witnessed the industrial revolution through the invention of the steam engine. The second revolution started with electrification, however, the development became rapidly. Nowadays, electricity is one of the most important necessities in modern life. It was a measure for development of countries and their progress. Most power plants use fossil fuels to generate electricity and, due to the many environmental and health risks of burning fossil fuels, the development of a country now is measured by its contribution in reducing greenhouse gases and its use of renewable resources. Improving electrical systems and reducing power loss in generation, transmission, and distribution lead to minimize of the emission of carbon dioxide (CO2) which is the main reason of climate change.

Electric and electronic apparatuses and equipment, whether they are sources or loads, operate with different voltages and currents - in term of frequencies and values. Some of them work on alternating voltages, some work on direct voltage, and other work on a combination of the two systems. Previously, several ways had been used for converting electric power such as motor-generator (M-G) set and rotary converter [1]. The most efficient way to convert power is by using semiconductor power devices. The converters are classified according to their functions into four types: rectifiers (AC to DC), inverters (DC to AC), converters or choppers (DC to DC), and cyclo-converters (AC to AC). Accordingly, power electronics, which deals with control of electrical energy and its conversion from one form to another, is the key for interconnecting different technologies such as solar power systems, wind power systems, and energy storage systems. Therefore, the development and improvement of power electronics contribute greatly in lowering the rate of harmful gas emissions.

The power range of the converters depends on the application for which they are used. This can be from milliwatts such as in mobile phones to megawatts such as in power transmission systems. In most cases, the power conversion is handled by using switch-mode power supplies (SMPS) conversion concept. SMPS converters provide many advantages such as high efficiency, high
power density, step up/down possibility, and multiple outputs. In SMPS, the transistor or transistors – according to the topology- work in switched mode. For isolated converters, usually a high frequency transformer is used to provide galvanic isolation between the input and output voltages. Additionally, the output voltage is regulated by changing the on-time duration of the pulse of the gate of the transistor (this duration is known as the duty cycle or the duty ratio). The efficiency of the converter is improved with switching mode because the switch is on and off for a specific time; therefore, it will minimize the loss. A lower loss is important because it results in a smaller size for the heat sinks and better utilization of the energy source. Also, the size of the passive components is inversely proportional to the switching frequency. By increasing the switching frequency, the size of the magnetic elements (transformers, inductors) as well as the required capacitors are reduced for storing the same amount of the power. The electric and magnetic energies are governed by [2], so that high power densities can be accomplished.

\[
E_L = \frac{1}{2} LI^2, \quad (1.1)
\]
\[
E_C = \frac{1}{2} CV^2, \quad (1.2)
\]
\[
P_L = f_s \Delta E_L, \quad (1.3)
\]
\[
P_C = f_s \Delta E_C, \quad (1.4)
\]

where \( E_L \) the energy stored in an inductor (\( L \)), \( E_C \) the energy stored in a capacitor (\( C \)), \( P_L \) is the inductor power, \( P_C \) is the power capacitor power, and \( f_s \) is the switching frequency.

Due to the limitations of the silicon (Si) based transistors, the switching frequencies cannot be increased by using these transistors [2]. Wide band-gap (WBG) based switches are replacing Si-based switches because their chemical and physical characteristics allow them to operate with higher frequencies, temperatures, and voltages [2].

SMPS are widely used in renewable applications. Depending on the application and the system architecture, they are used for different purposes such as stepping up or down the voltage, converting DC into AC or vice versa, isolating the source from the load, or managing the direction and quality of the power.
Renewable generating applications are divided into two main categories: standalone or off-grid applications, they are used to supply loads which are not connected to the electric grid, and grid-connected or on-grid where the sources and the loads are connected to the electric grid. Whether they are on or off grid, all renewable resources are unstable because of the fluctuating nature of such resources. Therefore, the energy storage systems are connected in parallel with renewable energy sources to compensate for these fluctuations, improve the efficiency, and enhance the reliability of the system [3],[4].

In photovoltaic (PV) systems, electricity generation significantly varies depending on the time of the day and the weather conditions [5], [6]. Therefore, energy storage systems are required in order to make full use of PV systems. In [7], different topologies for connecting solar modules are discussed. It demonstrates that the best applicable topologies for the power range 2 to 5 kW are string and multi-string topologies which are suitable for residential applications. DC/DC conversion is required to connect an energy storage system (ESS) as shown in Fig. 1.1 in order to improve the performance of the whole system by making it more stable and delivering the energy at the request.

![String and Multi-String Topologies with ESS](image)

Figure 1.1 String and Multi-String Topologies with ESS

In the previous case, the string output voltage is high. Moreover, PV modules can be connected in parallel to eliminate the power loss of partial shading and obtain better output at low light intensities [8], [9]. Accordingly, a possible connection of a high voltage energy storage system with the low voltage PV panels as shown in Fig. 1.2.
As a result, two types of energy storage systems can be used with PV systems, high voltage ESS which is above 100 V and low voltage ESS which is below 100 V. Different structures and topologies with various voltages are proposed for designing a PV system with battery packs. As stated in [10], the efficiency of the converter that connects the battery bank to the DC bus increases by increasing the number of the battery cells in series. Hence, the proposed system has a 333 V the battery voltage and 380 V the DC-link voltage linked by an H bridge buck/boost converter. Another system that uses a high voltage battery is presented in [11]. The voltage range of the battery bank is 168 V to 336 V and the voltage range of the DC-link is 600 V to 680 V. The battery bank and the DC-link are connected by a bidirectional buck/boost converter. A PV system with a low voltage ESS is demonstrated in [12], a half bridge converter operates in buck mood to charge the battery module with 54V normal voltage and in boost mood to feed the DC-link with 400V.

In [13], the system consists of three DC buses as illustrated in Fig. 1.3 (a), the high voltage 400V bus which is connected to the grid and the load by an inverter and to the intermediate bus by a dual half-bridge (DHB) converter, the intermediate or medium voltage 100V bus which connects the sources – PV arrays, and battery bank- to the high voltage bus via converters, and the low voltage 48V bus which connects the battery bank to the intermediate bus through a bidirectional converter. It is thought that reducing the system to two buses only – the high voltage and low voltage buses – and using two DC/DC converters with an inverter as shown in Fig. 1.3 (b) improve the system performance, simplify the system complexity, and lower the cost of the system.

Figure 1.2 Parallel Topology with ESS
From the above, it is concluded that an efficient bidirectional DC/DC converter is required to improve the performance for the whole PV systems. In term of a low power loss converter, wide band gap devices have lower losses comparing to the Si-based devices. In term of high power density, the fast switching property of wide band gap transistors enables the converters to achieve higher power densities than Si-based converters. Therefore, selecting the right topology for the bidirectional converter and improving its performance by using GaN-based transistors are discussed in this research in order to improve residential and home solar systems operation regarding to the cost and the size.

1.2 Objective and Outline

This research proposes a methodology for comparing the efficiency of GaN-based and Si-based single-phase dual active bridge converters. To make the comparison fair, the research focuses on the optimization of the design of both converters and using approximately the same PCB layout. The GaN and Si transistors used in this study have comparable voltages and currents ratings.
Chapter 2 explores GaN-based transistors. Their advantages and limitations in power electronics are discussed. Also, it compares the lateral and vertical GaN devices. Finally, it presents GaN-based dual active bridge converters that have been created.

In Chapter 3, a survey of bidirectional converters is provided along with the major differences between the isolated and non-isolated converters. In addition, it presents different isolated DC-DC bi-directional topologies and the advantages and disadvantages of each topology.

Chapter 4 introduces the principle of operation of a single-phase DAB and the states of operation. It also explains zero voltage switching and effect of the leakage inductance of the transformer on the range of the soft switching, amount of transferred power, and switching frequency.

In Chapter 5 theoretical analysis for a 4 kW (48V - 400V) DAB converter is presented. Simulation is provided to verify theoretical analysis results.

The experimental results for GaN-based DAB and Si-based DAB are discussed with the problems that have been faced in Chapter 6.

The last chapter concludes and provides a brief summary about the work done. In addition, it states the future work and the suggested ways for solving the problems.
CHAPTER TWO

2. Gallium Nitride-based Semiconductors

Existing semiconductor power devices are primarily based on silicon. Due to the development in many applications over the last couple of decades, Si devices have approached their theoretical limit and new materials have been utilized to meet the requirements such as high blocking voltage and switching frequencies [14]. In this chapter, WBG based devices and the differences among them and the Si-based devices are presented in addition to the principle of operation and characteristics of GaN-based transistors. Also, the constraints and difficulties of GaN material and lateral and vertical transistors are discussed. Commercially available GaN transistors and manufacturers are presented. The last section is a review of GaN-based dual active bridge converters.

2.1 Wide Bandgap Semiconductors and Power Applications

Power devices based on wide bandgap semiconductors could play an important role in improving a system’s performance. The characteristics of GaN and silicon carbide (SiC) enable the devices to operate at higher temperature, faster switching frequencies, and higher voltage stresses than silicon [14], [15]. Table 2.1 compares the important material properties of GaN and SiC with Si.

<table>
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<tr>
<th>Material</th>
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<tr>
<td></td>
<td>$E_G$ (eV)</td>
<td>$E_C$ (MV/cm)</td>
</tr>
<tr>
<td>Si</td>
<td>1.2</td>
<td>0.3</td>
</tr>
<tr>
<td>GaN</td>
<td>3.4</td>
<td>3.0</td>
</tr>
<tr>
<td>SiC</td>
<td>3.3</td>
<td>2.0</td>
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Table 2.1 Fundamental Material Properties of Si, GaN, and SiC [15][16]
The above properties of semiconductor materials according to [2], [15]–[18] as follows:

**The bandgap** ($E_G$) is the energy which is required to ionize atoms and create free electrons.

**The critical electric field strength** ($E_C$) is an important parameter for identifying the avalanche breakdown in power devices structures. Because the impact ionization coefficient depends on the electric field strength, the avalanche breakdown occurs when the electric field approaches the critical electric field.

**Breakdown Voltage** ($B_V$) is the voltage when the transistor breaks down and a leakage current starts to flow between the source and the drain while the gate is shorted to the source.

**Electron saturation velocity** ($v_{sat}$) is defined as the maximum speed that can be reached by an electron at high electric fields in a specified medium.

**Drift velocity** ($u$) is the average velocity of an electron in the direction of the force imposed by an electric field.

**Electron Mobility** ($\mu_n$) is the constant of the proportionality between the drift velocity and the electric field.

**The dielectric constant** ($\varepsilon_s$) is the ratio of the permittivity of a semiconductor to the permittivity of the free space. The electric flux increases as the dielectric increases. Semiconductors with high dielectric constants breaks down more easily than the materials with low dielectric constants.

**Thermal conductivity** ($\chi$) refers to the amount and speed of transferring heat in a material from the region with higher temperature to the region with lower temperature.

From Table 2.1, it is noticed that the GaN and SiC have higher electron moving properties, larger saturation velocity, higher bandgap, and higher thermal conductivity than Si, which is clearly advantages for WBG materials. The larger value of $E_G$ indicates that the transportation of the electrons is less probable through the band when the temperature increases [17], [19]. The thermal conductivity of SiC is much larger than Si and GaN as well, which makes it convenient to be used in very high temperature environments. Although the thermal conductivity of the GaN and Si are almost the same, GaN’s higher bandgap allows its operation in relatively higher temperatures than Si. Therefore, using WBG-based devices enables the size of the cooling equipment to decrease. This can achieve higher power densities [2]. Due to each material’s characteristics, SiC-based
devices are used in high voltages applications while GaN-based devices voltage range is between 100 V and 700 V [2]. The high voltage devices reduce the need to connect the devices in series for high voltage applications, so they make high voltage converters easier and more efficient. However, GaN-based transistors are given emphasis more than other WBG materials because half of the power electronics devices are in the range 400 V to 700 V [15].

The cutoff frequency \( f_T \) of the semiconductors can be calculated by applying [16]:

\[
f_T = \frac{E_C \cdot V_{sat}}{2\pi \cdot B_V}. \tag{2.1}
\]

It is observed that the frequency and breakdown voltage are inversely proportional. The material that has higher \( B_V \) has a lower cutoff frequency. Because of a very high \( E_C \) and \( V_{sat} \), wide bandgap devices can accomplish high frequency operation.

According to [16], [17], GaN crystals can be found in different structures such as zincblende and wurtzite. Due to the higher bandgap in the wurtzite structure, it is preferred over the other structures. The wurtzite structure has a hexagonal unit cell and contains six atoms; therefore, it is stable and can resist high temperatures without decomposition. The formation of charged interfaces is a result of the crystal polarization in GaN and aluminum gallium nitride (AlGaN). However, electrons gather near the interfaces because of this polarization. Additional charges at the interface are induced because of the piezo-electric polarization. These two charge aggregation components result in high polarization based electron density in GaN-based hetero-structures close to the interface. So, a 2-dimensional electron gas (2DEG) can be formed without any doping as shown in Fig. 2.1. Accordingly, the increase in electron mobility is due to the purity and the absence of doping.

![Figure 2.1 Two-dimensional Electron Gas](image-url)
Since the electric field controls the current through the channel, high electron mobility transistors (HEMTs) belong to the family field effect transistors (FETs). In HEMTs, the majority carriers are electrons, thus; the current flows from the drain to the source through the channel which is formed in the 2DEG. If these electrons are confined in the 2DEG when the gate to source voltage $V_{gs}$ equals zero, the HEMT is a normally-on transistor and it is known as depletion mode device (D-mode). In order to control the flow of the current, a negative voltage $V_{gs}$ is required to deplete the channel and regulate the drain current. The voltage value at which the current equals zero and fully depletes the channel is called the threshold voltage ($V_{th}$). On the other hand, the enhancement mode (E-mode) are normally-off devices. The drain current is zero when the $V_{gs}$ is equal to zero or a negative value and it flows when the $V_{gs}$ is equal to or greater than the threshold voltage.

Because of the advantages of the enhancement mode HEMTs, they are more desirable than D-mode transistors. First, normally-off devices offer fail-safe operation which reduces the risk of failures in the system. Second, E-mode transistors reduce the power consumption during the stand-by state. Finally, they reduce the circuit complexity and system cost.

2.2 Electric Characteristics of GaN-based Transistors

The distance between the electrodes determines the breakdown voltage of the transistor. In addition, the on-resistance increases by increasing the space between the electrodes. Due to the higher breakdown voltages that WBG materials have, the distance between those electrodes can be decreased while and maintaining the blocking voltage as a Si device. In [16], the on-resistance ($R_{ds_{on}}$) is given as

$$R_{ds_{on}} = \frac{1}{W_g q \mu N_s} (L_g + L_d),$$

where $W_g$ and $L_g$ are the gate width and length respectively, $L_d$ is drift region length, $N_s$ is the sheet carrier density (electron concentration), $\mu$ is the 2DEG mobility, and $q$ is the electron charge ($-1.6 \times 10^{-19}$ coulombs).

So, the on-resistance is the sum of the channel resistance $R_{ch}$ and the drift region resistance $R_d$. In the low voltage range, [16] concludes, the $R_{ds_{on}}$ approaches the channel resistor while in the
high voltage range, it approaches to the drift region resistor. Therefore, to reduce conduction losses in low voltage devices, reducing $R_{ch}$ is important. However, reducing $R_d$ is the more important for reducing conduction losses in high voltage devices.

The drain current of a HEMT device is given as [18]:

$$I_d = \frac{W_g \mu C_g}{L_{ds}} \left[ (V_{gs} + V_{th})V_{ds} - \frac{V_{ds}^2}{2} \right],$$

(2.3)

where $L_{ds}$ is the drain to source distance, $V_{ds}$ is the drain to source voltage, $C_g$ is the gate capacitance, and $V_{th}$ is the threshold voltage.

Transconductance ($g_m$) indicates the incremental change of the drain current of a transistor versus the gate to source voltage $V_{gs}$ and it depends on the distance between the drain and the source, the thickness of AlGaN layer, and the resistivity of the ohmic contacts. However, the gain of a transistor is critically dependent on the transconductance. It is given by the following equation [18]:

$$g_m = \frac{\partial I_d}{\partial V_{gs}}.$$  

(2.4)

During transistor switching, the capacitors of the device determine the energy that will be lost in the transistor. These capacitors are the drain-source capacitor $C_{ds}$, gate-drain capacitor $C_{gd}$, and gate-source capacitor $C_{gs}$ as shown in Fig. 2.2. The speed of switching is determined by the amount of charge which is directly proportional to the transistor capacitors. These three capacitors form the total capacitances of the device: the input capacitance $C_{iss} = C_{gd} + C_{gs}$, the output capacitance $C_{oss} = C_{gd} + C_{ds}$, and the reverse capacitance $C_{rss} = C_{gd}$ which is known also as the Miller capacitance [2], [15]. The ratio $C_{gd}/C_{gs}$ refers to the Miller ratio. Due to the equation $C = Q/V$, these capacitances are functions of the voltages that are applied to the drain and the gate of the transistor. Therefore, the smaller value of the capacitance in WBG-based devices is one factor that enables these devices to switch faster than Si-based devices.
2.3 Lateral and Vertical GaN Power Transistors

GaN-based transistors are classified into two types of devices according to their structure, lateral HEMT and vertical GaN transistors. Currently, the vertical GaN transistors are not available commercially. However, lateral devices have been available since 2009 and, thus, they have been studied extensively. A vertical GaN device structure is the same as that of Si MOSFET. Moreover, this structure is very efficient for realizing high breakdown and low on-resistance although it requires a bulk GaN substrate which leads to an increasing in the cost of the device [2], [20]. Some fabrications methods to minimize the GaN substrate such as GaN-on-Si and GaN wafer fabrication in addition to the main fabrication method which is a GaN-on-GaN fabrication [20]. Fig. 2.3 (a) and (b), show the structures of the lateral and vertical GaN transistors [20].

![Figure 2.2 GaN Transistor Capacitances](image)

**Figure 2.2 GaN Transistor Capacitances**

![Figure 2.3 (a) GaN lateral HEMT Structure (b) GaN Vertical MOSFET Structure](image)

**Figure 2.3 (a) GaN lateral HEMT Structure (b) GaN Vertical MOSFET Structure**
The fabrication process of the vertical GaN transistors is more complicated than that of HEMT devices. Furthermore, the substrates used for lateral devices are usually Si, and are larger and cheaper than GaN substrates that are used in the vertical structure [21]. Unlike the vertical structure, since HEMT transistors are bidirectional switches, an anti-parallel diode is not needed to enable the current flow in the reverse direction that is known as third quadrant operation. In [2], the reverse current flow in HEMT is explained in detail. In contrast, the vertical GaN structures are expected to achieve higher power densities because the increased breakdown voltage for lateral structures tends to increase the drift region, which makes the transistor larger, while for a vertical structure an increase in the thickness of only the buffer layer is required.

2.4 GaN Challenges and Limitations

Although GaN transistors provide many advantages, they face two types of limitations, the limitation of the material itself and the limitation of the transistors that are based on GaN. In this section, two limitations are explored: the issue of self-heating and the dynamic on-resistance.

According to [16], [22], and [23], one of the challenges that affect the reliability and performance of GaN-based transistors is self-heating. Due to the high-power density of these transistors, a very high temperature can be reached when high current flows through the material. Increasing the channel temperature or self-heating of the device leads the performance of the device to degrade and may change its material properties such as the electron mobility and band-gap in addition to increasing in the gate leakage current. As the mobility decreases, the drain current decreases and the device voltage increases result in changing the I-V\(^1\) curve characteristics of the device. The main factor on which self-heating depends is the thermal conductivity of the material; therefore, the SiC-based devices are affected less than GaN devices. The aforementioned reason prompts the fabrication of GaN-based transistors on SiC substrates.

The second challenge is the dynamic on-resistance [20], [24]. It is a result of a current collapse phenomenon and it increases by increasing the switching frequency and the voltage. GaN-based devices are supposed to be switching faster than Si based devices; thus, it is a critical factor for those devices. Current collapse occurs when a high voltage transient is applied to the drain of the

\(^1\) Current vs. voltage characteristics.
transistor. It can be measured by determining the difference between the output current and its corresponding DC value. The conducting loss becomes higher when the dynamic on-resistance increases and can be mitigated by using field plates, connecting substrate and source of the transistor together, or adding a second p-doped drain contact [20], [24].

2.5 GaN-based Transistors

Several companies have been producing GaN-based power transistors. Different structures with different parameters transistors have been manufactured. Two main types of power devices are available commercially: the cascode devices and enhancement-mode devices. A summary of the manufacturers and available GaN devices with their parameters is presented [20]. Table 2.2 shows the available commercial cascode transistors and Table 2.3 shows the available (E-mode) transistors.

Table 2.2 Cascode GaN-based Transistors [20]

<table>
<thead>
<tr>
<th>Manufacturer</th>
<th>Rated Voltage (V)</th>
<th>Rated Current</th>
<th>R_{ds,on} (mΩ)</th>
<th>Q_g (nC)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transphorm</td>
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<td>9</td>
<td>290</td>
<td>6.2</td>
</tr>
<tr>
<td></td>
<td></td>
<td>17</td>
<td>150</td>
<td>6.2</td>
</tr>
<tr>
<td></td>
<td></td>
<td>34</td>
<td>52</td>
<td>19</td>
</tr>
<tr>
<td>Infineon</td>
<td>600</td>
<td>10</td>
<td>125</td>
<td>---</td>
</tr>
<tr>
<td>MicroGaN</td>
<td>600</td>
<td>---</td>
<td>320</td>
<td>---</td>
</tr>
<tr>
<td>RFMD/Qorvo</td>
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<td>25</td>
<td>85</td>
<td>16.2</td>
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<tr>
<td></td>
<td></td>
<td>30</td>
<td>45</td>
<td>15.7</td>
</tr>
<tr>
<td>VisIC</td>
<td>650</td>
<td>50</td>
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<td></td>
<td>1200</td>
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</tr>
<tr>
<td>Texas Instruments</td>
<td>600</td>
<td>12</td>
<td>70</td>
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</tr>
</tbody>
</table>
Table 2.3 E-mode GaN-based Transistor [20]

<table>
<thead>
<tr>
<th>Manufacturer</th>
<th>Rated Voltage (V)</th>
<th>Rated Current</th>
<th>$R_{ds_{on}}$ (m$\Omega$)</th>
<th>$Q_g$ (nC)</th>
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<td>---</td>
</tr>
<tr>
<td></td>
<td>1200</td>
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<td>---</td>
<td>---</td>
</tr>
<tr>
<td>Powdec</td>
<td>1200</td>
<td>16</td>
<td>---</td>
<td>---</td>
</tr>
</tbody>
</table>
2.6 Dual Active Bridge GaN-based Converter

Isolated converters provide many advantages such as safety, reliability, and wide conversion ratios. Extensive research has been carried out to improve the performance of isolated converters. On the other hand, limited research is available on GaN-based isolated converters, and even less is available on the topic of DAB converters based on GaN transistors. Isolated bidirectional converters will be discussed further in the next chapter.

In [25], [26], a 1.7 kW bidirectional converter has been presented. H-bridges are used on the input and the output of the converter to form a DAB. Enhancement mode GaN-based transistors are used from Efficient Power Conversion (EPC). The high side input voltage is 130 V and the low side input voltage is 52 V. The converter operates at a switching frequency of 50 kHz. The two modes of operation are discussed: buck mode and boost mode. The maximum measured efficiency is 98.8 % so it reduces approximately 40 % of the power loss compared to the existing state-of-art.

A bidirectional plug-in hybrid electric vehicle (PHEV) battery charger based on GaN-on-Si is discussed in [27], and 1 kW dual active bridge is presented. The DAB is connected to batteries on one side and to an AC/DC converter on the other side. Normally-off GaN devices from HRL Laboratories are used to build the battery charger. The converter was tested at 250V and 500 kHz with phase-shift modulation. The measured inductor current is 4.2 A at 1 kW while the measured output current is 4 A, thus; the converter has a small circulating energy. Although the total charger efficiency is 94.2 %, the DAB efficiency is 97.2 %.

In [28], a high step-down unregulated fixed ratio DAB is described. It is a resonant converter that operates at 1 MHz switching frequency. It steps down the voltage from 150 V to 12 V and is supposed to be used in data centers. EPC transistors are used on the primary bridge when the secondary bridge transistors are either GaN or Si. A 94.1 % efficiency at 120 V, 60 W input was achieved when all transistors of the two sides are GaN. When the secondary side devices are Si, a peak efficiency of 93.1 % at 120 V, 165 W input was achieved.

A 1 kW GaN-based isolated bidirectional converter for DC microgrid energy storage systems is proposed in [29]. The converter consists of a half-bridge and center tap with active clamp circuit. Due to low voltage, high current, and large conversion ratio, the active clamp circuit is used on the low side. It is a hybrid converter, the switches on the high side are GaN-based and Si-based on the
low side. The specifications of the converter are 400 V/12 V high/low side voltages, and 100 kHz switching frequency. The achieved power density and efficiency are 30 W/in$^3$, and 98.3 % respectively.

Device loss comparison of GaN-based LLC, Dual active bridge and phase shift quasi switched capacitor converter is presented in [30]. In this work, cascode transistors from Transphorm are selected for the primary side and E-mode transistors from EPC are selected for the secondary side. The three converters were designed for 1 kW, 400-48 V application case. The results show that the DAB converter maintains lower device loss than the other two converters from 300 W to 1 kW at different switching frequencies from 200 kHz to 2 MHz.

### 2.7 Summary

Wide bandgap materials based devices are being used in order to reduce the power loss and increase the efficiency and power density of converters. Their ability to operate at higher temperature, higher voltages, and higher switching frequencies make them more suitable for many applications. The main characteristics of semiconductors have been reviewed in addition to the differences among the materials. The structure of GaN devices was presented in addition to the principle of operation of HEMT. The two types of HEMTs: enhancement mode and depletion mode and the limitations and challenges of GaN based the transistors were discussed. The manufacturers and commercial available GaN transistors were listed in a table. In the last section, a literature review on dual active bridge converters based on GaN devices was presented.
CHAPTER THREE

3. Bidirectional Converters

Bidirectional converters (BDC) are used in many applications where transferring power in two directions is required. Some examples include electric and hybrid vehicles (EV& HEV), fuel cells systems, energy storage systems (ESS), renewable energy, and so on [31]. As mentioned above, the bidirectional converters act as the interface between different types of energy sources with energy storage devices and manage the flow of the power as shown in Fig. 3.1.

In photovoltaic (PV) systems, the bidirectional converters are used to transfer the energy to the energy storage systems when the generated power is more than the demanded power, and to deliver the energy from the storage systems when the generated power is less than the required [32]. Thus, the bidirectional converters keep a stable bus voltage and make full usage of the photovoltaic cells and the storage devices under all possible conditions and requirements.

In addition to reducing the cost, bidirectional converters improve the performance of the whole system in term of power density and efficiency. BDCs are divided into two types: non-isolated converters, and isolated converters. Depending on the application, non-isolated bidirectional
(NBDC) type is preferable where the size, weight, and cost are the main concern and where a wide voltage conversion range is not required [33], [34].

Most isolated bidirectional converters (IBDC) have the same structure. Typically, they consist of two switching DC-AC converters connected through their AC side via high-frequency transformer as shown in Fig. 3.2 [3], [35].

![Isolated Bidirectional Converters Structure](image)

Figure 3.2 Isolated Bidirectional Converters Structure

Depending on the power flow direction, one of the converters transforms the DC voltage to an AC waveform, which is then applied to the transformer. The other converter rectifies this AC signal and creates a DC voltage. Isolated bidirectional converters are used when a wide conversion ratio is required. Unlike the non-isolated converters, they can be more complicated and costly, but they enhance the safety and reliability of the system in addition to achieving the soft switching feature by applying the phase-shift control [36], [37].
3.1 Converter Selection

Selection of the topology for the primary and the secondary for the bidirectional converters depends on the converter ratings such as the required voltage and the power. For each topology, the limitations such as the switches stress and the peak and ripple current determine the suitable voltage and power ranges.

3.1.1 Fly-back Converters

The fly-back converter is one of most popular isolated DC-DC converters. Because of its simplicity, it is used in many applications and it has many configurations. Fig. 3.3 shows the basic topology of a fly-back converter.

![Figure 3.3 Fly-back Converter](image)

The magnetizing inductance \( L_m \) of the transformer is utilized as a functional inductance. During the on-time period, when the switch \( S_1 \) is closed, the diode \( D_2 \) is reversed biased and the voltage across the primary side of the transformer equals to the input voltage \( V_1 \). The primary current ramps up with a slop \( V_1/L_m \) and charges the magnetizing inductor. During this period, the energy is transferred from the input source to the magnetizing inductor. When the switch \( S_1 \) is turned off, the magnetizing current induces a positive current into the secondary side of the transformer which turns the diode \( D_2 \) on, supplying the load [38], [39].

The bidirectional fly-back converter is derived from the unidirectional converter by replacing the unidirectional switches with bidirectional ones as shown in Figure 3.4. During the forward conduction, the switch \( S_1 \) and the diode \( D_2 \) operates, the power transfers from \( V_1 \) to \( V_2 \). During the backward conduction, the switch \( Q_1 \) and the diode \( D_1 \) operate, the power transfers from \( V_2 \) to \( V_1 \) [40], [41].
Although the many advantages of fly-back converters such as reducing the number of components, step up and down feature, and multiple outputs capability, they suffer from different limitations. In addition to the double of voltage ($V_1$ or $V_2$) on the switches, the fly-back converter can have voltage spikes that will stress other components and this spike increases with increasing leakage inductance in the transformer. In addition, a large magnetizing inductor is required, because it storages the whole energy during operation; therefore, it affects the size of the transformer. Furthermore, a high peak current because of the discontinuous mode [38], [42][43].

### 3.1.2 Forward Converters

Fig. 3.5 shows a forward converter circuit. It is also well-known for its isolation feature which is provided by the transformer. Unlike the fly-back converter, the input and output circuits conduct simultaneously. When the switch $S_1$ is turned on, the input voltage is applied to the primary winding and a scaled voltage appears across the secondary winding. Then, the voltage of the secondary side of the transformer is applied to $LC$ circuit through the diode $D_2$ which becomes forward biased during this interval. When the switch $S_1$ is turned off, the currents of the primary and the secondary windings of the transformer reduce to zero. However, the load current continues flowing the output inductor. The diode $D_3$ provides the freewheeling path for the inductor current. During freewheeling, the inductor current ramps down but the existence of the capacitor keeps the output voltage with the required range. The energy in the core of the transformer is reset by the tertiary winding and the diode $D_1$ which becomes forward biased after the switch $S_1$ is turned off. The rest winding method has a considerable drawback which is the high discharging spike because
of the leakage inductance of the transformer [44]. Different methods and schemes for resetting the transformer are stated in [45].

![Diagram of Forward Converter](image1)

**Figure 3.5 Forward Converter**

The main advantages of the forward converter are the capability for providing multiple outputs by means of an isolation transformer and the output current has low ripples. In addition, forward converters are easy to control therefore they are used for isolated offline applications [44]. However, the drawbacks of the forward converter limit its operation. The transformer should be designed with a low leakage inductance to reduce the voltage stress of the switch.

For bidirectional power flow applications, the basic topology of the forward converter is not used. Therefore, it is believed that using an alternative topology such as two-switch forward converter [46] can achieve the bidirectional power flow as shown in Fig. 3.6. In [47], a forward converter is combined with a full bridge converter through a transformer in order to manage the two directions power flow for charging and discharging the ESS.

![Diagram of Bidirectional Forward Converter](image2)

**Figure 3.6 Bidirectional Forward Converter**
3.1.3 Push-Pull Converters

The basic configuration of a push-pull converter is shown in Fig. 3.7. It also provides galvanic isolation by the high-frequency transformer. It can be used for step-up or step-down the voltage by changing the turn ratio of the transformer. The advantages of this topology are simplicity and the capability to be used for higher power applications than fly-back and forward converters. This leads to increased use in industrial DC power applications [48]. In addition, the switches have a common ground, which made them easy to drive.

![Figure 3.7 Push-Pull Converter](image)

The two transistors cannot be switched on simultaneously because of the opposite directions of the primary windings, generate reverse flux which cancels the mutual inductance of the primary side in addition to causing a high current that may damage or even destroy the transistors.

When the switch $S_1$ is on $D_3$ is forward biased and $D_4$ is reversed biased. The voltage on the corresponding winding equals to the input voltage. The current ramps up and its magnitude depends on the turn ratio of the transformer and the output inductor $L_{out}$. In the second interval, when $S_2$ is turned on, $D_4$ conducts and $D_3$ is reversed biased. In between the two intervals, the inductor current splits equally between the secondary windings and free-wheels through the output diodes.

The aforementioned converter has two major limitations, the switches must handle double of the voltage $2*V_1$, which make them more expensive especially in high voltage applications and the difficulty in designing a center tap transformer.
To achieve bidirectional power flow, another converter is connected to the secondary side of the transformer. Most of the bidirectional converters that use push-pull topology are current-fed converters. Current-fed topology has many advantages over its voltage-fed counterpart, lower input current ripple, smaller transformer because of the lower turn ratio, limited diode ringing and easier current control ability [49]. The main reason behind using a bidirectional current-fed converter instead of voltage-fed is that it is suitable for applications which have a wide variation in source voltage and require higher voltage conversion ratio, therefore; the voltage stress, in this case, is less, in addition to it has a wide soft-switching range [50]. Fig. 3.8 shows a bidirectional voltage-fed push-pull converter.

![Bidirectional Push-Pull Converter](image)

Figure 3.8 Bidirectional Push-Pull Converter
3.1.4 Half-bridge Converters

Half-bridge converter also utilizes a power transformer like a push-pull converter. It consists of two switches in parallel with a capacitor leg as it shown in Fig. 3.9. The two capacitors \((C_1 \& C_2)\) work as a voltage divider, and each capacitor provides a dc voltage \(V_{1/2}\) at its terminals [51].

![Half-bridge Converter](image1)

By connecting another converter on the secondary side of the transformer, bidirectional power transfer is achieved as it shown in Fig. 3.10. In forward mode, when the power is desired to be transferred from \(V_1\) to \(V_2\), the primary side functions as an inverter while the secondary side functions as a rectifier. In backward mode, the two converters switch their roles, the primary side works as a rectifier and the secondary side works as an inverter. The leakage inductor \(L_g\) limits the current between the two bridges.

![Bidirectional Half-bridge Converter](image2)
Half-bridge converters are used for low and medium power applications where the number of switches needs to be minimized [31]. Unlike push-pull converter, the half-bridge converter’s switch voltage stress does not exceed the input voltage. Moreover, only one primary winding and one secondary winding are required, then; the transformer core’s window can be better utilized. Although the abovementioned advantages, the drawbacks of the half-bridge converter are the applied voltage to the windings is half of the input voltage because of the two capacitors, therefore; the turns ratio must be doubled to achieve the desired output voltage and the average current flowing through windings is two times to supply the desired power increases winding and switches conduction losses. Furthermore, the capacitor should sustain a large ripple current, so these capacitors are bulky and their size increases with increasing the transferred power. To achieve high power efficiency, low-ESR capacitors are required which increases the converter cost. As a result, the capacitors become bulky and expensive when increasing the power and the voltage [52]–[54].

3.1.5 Full-Bridge Converters

A single phase full bridge converter consists of two legs and each leg consists of two switches. Depending on the secondary winding of the transformer, the secondary rectifier is either full bridge diode or voltage doubler rectifiers as shown in Fig. 3.11. In hard switching method, the switches $S_1$ and $S_4$ are turned on simultaneously as a pair and consecutively with the switches $S_2$ and $S_3$ [51]. Hence, the primary winding of the transformer is connected to the DC supply $V_1$ through the leakage inductance $L_g$ which is usually very small comparing with the output filter $L_{out}$.

![Figure 3.11 Full Bridge Converter](image)
During the dead-time, when all switches are off, the antiparallel diodes provide a path for the current in the leakage inductor. In the phase shift method, soft switching can be achieved without requiring any additional resonant components. The voltage of the primary winding is controlled by shifting the phase of the switching in the two converter legs. The current flows from the source to the windings and its maximum value is determined by the output filter. However, the soft switching feature might be lost under light load conditions [55].

Unlike the half-bridge converter where the inductor current passes through DC capacitors, a full bridge converter does not suffer from high capacitor ripple currents because the inductor current flows through active switches. The main drawback of this topology is the total switching loss is significant due to the large number of switches comparing to the other topologies.

### 3.2 Dual Active Bridge Converters

Whether it is a single-phase or a three-phase converter, the dual active bridge converter consists of two H full bridge converters, one operating in inversion mode and the other operating in rectification mode, interfaced through a high-frequency transformer on their AC side. Each bridge can be controlled to generate a high-frequency AC voltage at its transformer terminals ($+V_{dc}$, 0, $-V_{dc}$) and to determine real and reactive power flow between the two bridges by changing the magnitude and the phase relationship between the two voltages. Fig. 3.12 (a) (b) shows three and single-phase dual active bridge converters.
Selecting between the two topologies, single-phase, and three-phase, depends upon the advantages and drawbacks of each topology. In [56]–[58], single and three-phase DAB DC-DC converters and their transformers are compared. First, the ripple of the input current of the single-phase converter is higher than its alternative three-phase, therefore; it requires a large DC link capacitor to absorb the oscillations in the current. Second, due an extra set of legs in three-phase DAB converter, the switches VA ratings are reduced because the single-phase DAB’s current stress is higher. Third, in the transformer core of the three phase DAB converter, the flux components are shifted by 120 degrees so they cancel each other and minimize the required size of the transformer. On the other hand, the benefits of the potential reduction in the size of the three-phase DAB transformer are totally negated for thermal reasons, since the surface area provided by the smaller core is not sufficient to dissipate the heat generated by the magnetic and ohmic losses. Moreover, the high number of the semiconductor switches, 12 transistors in the three-phase DAB, negates the loss benefit of the lower peak current. Finally, achieving identical leakage inductances in each phase is relatively difficult.
### 3.2.1 Dual Active Bridge Converter Applications

Due to their many advantages, DAB converters are used in different applications. They reduce the number of components since the output inductor filter is not required. The zero-voltage switching (ZVS) is possible over a wide operating range and does not need extra components. They have the capability for stepping the voltage up or down by changing the turn ratio of the transformer. Galvanic isolation is provided by either a transformer or capacitors. Moreover, they minimize the voltage stress across the switches. In addition, high power densities with high efficiency are achieved. Conversely, the drawback of DAB is the presence of high circulating current in the transformer which increases losses and leads the converter to lose zero voltage switching at light load [55], [59], [60].

1- **Automotive Applications:**

Nowadays, the trends in the automotive industry are to reduce the fuel consumption. Hybrid electric vehicles (HEV), electric vehicles (EV), and fuel cell vehicles are being commercialized. For their high efficiency, high power density, and low weight, DAB converters are being used for these vehicle technologies such as on-board battery charger module (OBCM) for electric vehicles and plug-in hybrid electric vehicles (PHEV) [61], [62].

2- **Renewable Energy Applications:**

DAB converters can be used to manage the output voltage of the renewable power sources such as photovoltaic cells, wind turbine, and fuel cell. They also connect the low sources’ bus with the high voltage buses. Therefore; they are considered the core circuit for interfacing the renewable energy sources with the AC systems [63], [64].

3- **Uninterruptable power supply (UPS):**

During the normal operation, DAB operates in charging mode and supplies the electric power from the main source to the battery. On the contrary, during the main power failure, the load is provided with power from the battery by the DAB converter which operates in discharge mode. Depending on the batteries that are used, the converter is either step the voltage of the battery up or down or provide isolation only in case the two voltages of the system are equal.
4- **Multi-port Systems:**

Sometimes, variable sources and loads with different voltages are needed to be combined together. By using multi-port DAB, the sources and loads are matched to the system with electric isolation, in addition to decreasing the size and weight of the system because of using a single iron core transformer. In [65]–[68], different structures of multi-port active bridge DC/DC converters are proposed.

5- **Smart Grid Application:**

Because of the difference in voltages and currents in frequency and magnitude, in addition to controlling the power direction especially during the fluctuation, connecting variable sources and storage systems with the Smart Grid is a difficult task [69]. The Smart Grid uses Alternating Current (AC) while the sources and storage system are Direct Current (DC). Therefore, converters are used to interface these two forms of voltages and currents and to regulate the power flow during steady-state and transient situations. Usually, the power rating for those converters are in the kilowatts range, therefore; DAB is implied in these systems.

### 3.3 Summary

Isolated bidirectional converters are used where the voltage range is large and galvanic isolation is required for safety purposes. Each topology has its own limitations and advantages therefore, knowing these properties facilities to choose among them. The popular isolated converters and their advantages and disadvantages have been presented. In spite of the drawbacks of dual active bridge converter, it is very appropriate for high power applications. Consequently, it is used in different applications such as automations and renewables. DAB is a very promising converter to be used for connecting PV and grid with energy storage systems.
CHAPTER FOUR

4. Single Phase Dual Active Bridge Operation Principle

The principle of operation of a DAB, power flow analysis by using phase shift switching, and the modes of operations are presented in this chapter. In addition, types of gate drive circuits and the losses in the gate circuit for the selected switches are discussed.

The DAB converter consists of two full bridge converters connected by a high-frequency transformer. Two AC voltages are provided to the primary and the secondary of the transformer in order to transfer power. Fig. 4.1 shows the circuit of a single-phase dual active bridge converter. The structure of single phase DAB can be represented by two controlled square wave voltage sources connected by an inductor as shown in Fig. 4.2. The generated voltages are shifted from each other by a specific angle $\delta$. For simplicity, the supply voltages are considered constant; all losses, magnetizing inductance, and parasitic capacitances of the transformer are neglected. Depending on the states of the switches, the two voltages have three levels ($+V_{dc}$, $0$, $-V_{dc}$).

![Figure 4.1 Single-Phase DAB Converter](image)

![Figure 4.2 DAB Equivalent Circuit](image)
Fig. 4.3 shows the leakage inductance of the transformer and the voltages across it. The voltage across the inductor is given by the following equation:

\[ v_{Lg}(t) = v_1(t) + v_2(t). \]  

(4.1)

This voltage causes the current to flow in the inductor from one bridge to another in four states. In the first state, the current increases until the angle \( \delta \) is reached. Then, in the second state, it increases according to the difference between the two voltages. Afterwards, the current will decreases as the phase shift angle is attained. Finally, it decreases smoothly depending, also, on the difference between the voltages. The four states are determined as follow:\(^2\):

\[ i_{Lg}(t_1) = \frac{1}{L_g} \int_{t_0}^{t_1} (v_1(t) + v_2(t)) \, dt, \]  

(4.2)

\[ i_{Lg}(t_2) = \frac{1}{L_g} \int_{t_1}^{t_2} (v_1(t) - v_2(t)) \, dt, \]  

(4.3)

\[ i_{Lg}(t_3) = \frac{1}{L_g} \int_{t_2}^{t_3} (-v_1(t) - v_2(t)) \, dt, \]  

(4.4)

\[ i_{Lg}(t_4) = \frac{1}{L_g} \int_{t_3}^{t_4} (-v_1(t) + v_2(t)) \, dt. \]  

(4.5)

The instantaneous power, whether it is received or transmitted by one of the sources, it can be calculated by,

\[ p(t) = v(t).i_{Lg}(t). \]  

(4.6)

\(^2\) In Eqs. (4.2 – 4.5), \( i_{Lg} \) is the current in the inductor as seen in Fig. 4.3.
In one switching cycle, the average power can be calculated as,

\[ P = \frac{1}{T} \int_{t_0}^{T+t_0} p(t) \, dt. \]  

(4.7)

However, the power magnitude and direction can be regulated by adjusting one or more of the following parameters:

1) The phase shift angle, \( \delta \); its range is \((-\pi \text{ to } \pi)\).
2) The duty cycle of the two voltages \( D_1 \) and \( D_2 \); their range is \((0 \text{ to } 0.5)\).
3) The switching frequency, \( f_s \).

4.1 Power Flow Analysis

In phase shift switching as shown in Fig. 4.5, the DAB converter operates with a constant switching frequency. The switches are turned on and off diagonally with maximum duty cycles, \( D_1 = D_2 = 50\% \). The phase shift angle is the only parameter that is adjusted in order to control the direction and the amount of the transferred power. As it introduced in Eqs. (4.2 – 4.5), the voltages \( V_1(t) \) and \( V_2(t) \), and the inductor current, repeat every half-cycle with reversed sign. During steady state, by assuming the current is flowing through the devices \( D_1, D_4, Q_1, \) and \( Q_4 \) as shown in Fig. 4.4 (a), the soft switching feature occurs. Accordingly, the voltages across the transistors \( S_1 \) and \( S_4 \), is zero; thus, the transistors \( S_1 \) and \( S_4 \) are turned on at zero voltage and then \( Q_1 \) and \( Q_4 \) are turned off. Eventually, the inductor current reverses and starts flowing through \( S_1 \) and \( S_4 \), as shown in Fig. 4.4 (b). The transistors \( S_1 \) and \( S_4 \) are turned off, the energy stored in inductor is transferred to the snubber capacitors, which are placed across the transistors, and the turn-off occurs under a zero voltage condition; when the transistors carry a certain minimum current, as illustrated in Fig. 4.4 (c). At the same time \( C_1 \) and \( C_4 \) are charging, the capacitors \( C_2 \) and \( C_3 \) will be discharged and force the diodes \( D_2, D_3 \) to be forward-biased. So, these diodes start commutating and conducting the inductor current as in Fig. 4.4 (d). In the same way, the transistors \( Q_1 \) and \( Q_4 \) are turned off under a zero voltage switching condition. Similarly, the other four transistors \( S_2, S_3, Q_2, \) and \( Q_3 \) are switched on in the same sequence.
Figure 4.4 (a, b, c, d) Steady State Operation Modes.
In Fig. 4.5, the operation waveforms and switching waveform of the DAB converter are presented.

Figure 4.5 DAB Converter Waveforms
From Eqs. (4.2 – 4.5) and Fig. 4.4, it is noted that the DAB converter operates in two modes. The first mode (Mode 1) is when the voltage sources have different polarity and the second mode (Mode 2) is when they have the same polarity. During Mode 1, the inductor current can be determined as,

\[ i_{Lg}(t) = \left( \frac{V_1 + V_2}{L_g} \right) t + i_{Lg}(0). \] (4.8)

In Mode 2, the inductor current can be expressed as,

\[ i_{Lg}(t) = \left( \frac{V_1 - V_2}{L_g} \right) \left( t - \frac{\delta}{2\pi f_s} \right) + i_{Lg}(\delta), \] (4.9)

where \( V_1 \) is the voltage produced by the first bridge and \( V_2 \) is the referred voltage produced by the second bridge, given as,

\[ V_2 = nV_H. \] (4.10)

In Eq. (4.10), \( n \) is the transformer turns ratio, and \( i_{Lg}(0) \) and \( i_{Lg}(\delta) \) are the instantaneous values of \( i_{Lg} \) when \( V_1 \) and \( V_2 \) change their polarities.

The instantaneous current values are given as,

\[ i_{Lg}(0) = \left( \frac{V_1 - V_2}{L_g} \right) \left( t - \frac{\delta}{2\pi f_s} \right) + i_{Lg}(\delta), \] (4.11)

and

\[ i_{Lg}(\delta) = \left( \frac{V_1 - V_2}{2L_g} \right) \left( t - \frac{\delta}{2\pi f_s} \right) + \left( \frac{V_1 + V_2}{2L_g} \right) \frac{\delta}{2\pi f_s}. \] (4.12)

Therefore, the average current can be calculated as,

\[ i_{Lg} = \frac{1}{2\pi} \int_0^{2\pi} i_{Lg}(t) \, dt, \] (4.13)

or

\[ i_{Lg} = \frac{1}{\pi} \left( \int_0^\delta \left( \frac{V_1 + V_2}{L_g} \right) t + i_{Lg}(0) \right) dt + \int_0^\delta \left( \frac{V_1 - V_2}{L_g} \right) \left( t - \frac{\delta}{2\pi f_s} \right) + i_{Lg}(\delta) \, dt. \] (4.14)
By simplifying Eq. (4.14) and substituting into Eq. (4.6) and as stated in [55], [59], the out power of the DAB converter can be calculated as,

\[ P_{out} = \left( \frac{V_1 V_2}{2 \pi f_s L_g} \right) \left( \delta - \frac{\delta^2}{\pi} \right). \]  

(4.15)

By plotting Eq. (4.15) for different values of \( \delta \), as shown in Fig. 4.6, it is noticed that the maximum power is transferred from one source to another in the forward mode at \( \delta = \pi/2 \) and in the reversed mode at \( \delta = -\pi/2 \).

![Figure 4.6 Transferred Power vs. Phase-Shift Angle](image)

### 4.2 Power Loss Analysis

Efficiency is an important concern with any converter because it allows for a better increase in power density and reduces the cost. Because of the high number of switches in the DAB, the power loss of the switches is high. As other converters, DAB power switches losses are divided into three types: switching losses, conducting losses, and gate drive losses. The switching loss is the loss of the energy that occurs during transition time of a transistor. During the turn-on transition, the loss is measured when the drain current starts rising to a specific value and the drain to source voltage drops to zero for an ideal transistor. In the turn-off period, the loss is measured when the drain to source voltage starts increasing and the drain current falls to zero. Since the DAB has a zero voltage
switching feature for reasonable transferred power, the switching losses can be neglected. Fig. 4.8 shows an ideal transistor switching loss. During one complete switching cycle the loss is given by Eq. (4.16), [70]:

\[ P_{sw} = \frac{1}{2} V_{ds} I_{d} (t_{on} + t_{off}) f_{s}, \]  

(4.16)

where \( P_{sw} \) is the switching loss and \( t_{on} \) and \( t_{off} \) are the time periods during transition.

![Figure 4.7 Ideal Transistor Switching waveforms](image)

In the conduction state, the power loss depends on the current that passes through the device. For an H bridge, the RMS current for each crossed pair of transistors is equal to the RMS current that flows through the transformer and the leakage inductor for a half-cycle. Referring to Eqs. (4.8) to (4.14) and the Fig. 4.5, the RMS current for each transistor can be calculated as,

\[ I_{d\_rms} = \sqrt{\frac{1}{2\pi} \left( \int_0^\delta \left( \frac{V_1 + V_2}{L_g} t + i_{Lg}(0) \right)^2 dt + \int_0^{\pi/2} \left( \frac{V_1 - V_2}{L_g} \right) \left( t - \frac{\delta}{2\pi f_s} \right) + i_{Lg}(\delta) \right)^2 dt}. \]  

(4.17)

As a result, the power conducting losses is obtained by,

\[ P_{con} = I_{d\_rms}^2 R_{ds\_on}. \]  

(4.18)

The third type of losses is the gate-drive loss. It the dissipated energy in the gate of the transistor at each cycle during the time that the gate voltage increases or decreases. Therefore, the dissipated power is directly proportional to the energy and the switching frequency as given in Eq. (4.19).

\[ P_{g} = E_{g} f_{s}. \]  

(4.19)
4.3 Gate Drive Circuit

Although power transistors have different types and structures, the principles for driving them are the same. The output current of the transistors is proportional to the charge that is initiated by the control electrode; therefore, they are charge-controlled devices [71]. As mentioned in chapter 2, the switching speed is determined by the parasitic capacitors \( C_{gs} \) and \( C_{gd} \) of the transistor. The parasitic capacitors are shown in Fig. 4.8 (a). During the transistor turn-on time, the gate to source voltage has three stages, as shown in Fig. 4.8 (b) [71]. In the first stages, \( V_{gs} \) charges the \( C_{gs} \) capacitor and it is divided into two levels: the first level is from 0 to \( V_{th} \); the transistor stays off and no drain current flows, the time period of this level is known as the turn-on delay time \( (t_{d(on)}) \), and the second level is from \( V_{th} \) to the Miller plateau voltage \( (V_{gs,Mp}) \) when the transistor starts conducting and the drain current starts to flow. \( V_{ds} \) begins declining by the beginning of the second stage and the gate current \( (I_g) \) passes through \( C_{gd} \) only. The slope of \( V_{gs} \) in this stage is zero because the value of \( C_{gd} \times V_{gd} \) changes very quickly and all the gate current is used to discharge \( C_{gd} \). The last stage starts when \( V_{ds} \) is equal to \( I_d \times R_{ds,on} \) or to the lowest value and both gate capacitors start charging. At the end of the third stage, \( C_{gd} \) and \( C_{gs} \) are fully charged and \( V_{gs} \) reaches its maximum value \( V_g \). The turn off procedure of a transistor is back tracking of the turning on stages. It begins when both gate capacitors start discharging and ends when \( V_{gs} < V_{th} \). As a result, the stages of turning a transistor on and off can be controlled depending on the gate drive topology, its efficiency, and the type of switching, whether it is hard or soft switching.

![Figure 4.8 (a) Transistors Capacitances (b) Transistor Turn-on Waveforms](image)
According to [72], gate drive circuits are classified into three main topologies: voltage source circuits, current source circuits, and resonant circuits. Fig. 4.9 shows the simplified representations for the three topologies.

![Gate Drive Topologies](image)

**Figure 4.9 Gate Drive Topologies (a) Voltage-Source Circuit (b) Current-Source Circuit (c) Resonant Circuit**

Because the power loss increases with switching frequency in a gate drive circuit, the current source and resonant circuits can be used for high-frequency applications [72]–[75]. They can recover the energy stored in the capacitor and increase the efficiency of the circuit; therefore, they have higher efficiencies than the voltage source topology. Furthermore, ringing and overshoot can be minimized by using these topologies. In contrast, the power density of the current source and resonant circuits is lower than the voltage source topology because additional inductors, capacitors, and semiconductors components are required.

Despite the low efficiencies of the voltage source gate driver circuits, they are the most common topologies. The capacitors $C_{gs}$ and $C_{gd}$ are charged through a series resistor in voltage source circuit [72]. The charging and discharging time for gate capacitors depend on the series resistance which is sum of the internal resistor of the gate of the transistor, an external resistor, and the gate driver resistor and the voltage of the gate driver during turn on and off, as shown in Fig. 4.10. Usually, the external resistor of the drive circuit and the gate driver are higher than the internal resistors of the transistor ($R_g + R_{driver} \gg R_{t,\text{internal}}$), so only $R_g$ and $R_{driver}$ are taken into account for determining the rise and fall time for the gate voltage. When $R_g$ is small and the switching frequency is high, the gate voltage will have high overshoot and undershoot because of the parasitic inductance in the charging and discharging loops which cause the ringing as a result of $L \frac{di}{dt}$.
4.3.1 Power Losses in Gate Drive Circuit

The elements of power dissipation in gate drive circuits are the charging and discharging dissipation [76], quiescent current draw dissipation, and cross-conduction (shoot-through) dissipation. The last two dissipation elements are important for designing gate drive for low power converters. In this research, only the first element will be considered because the converter is designed for high power.

The gate capacitance of a transistor determines gate switching losses. The gate capacitance value can be obtained by dividing the total gate charge \( Q_g \) of the transistors by \( V_{gs} \). However, the charging and discharging losses are given as,

\[
P_{g,sw} = C_g \ast V_{gs}^2 \ast f_s, \tag{4.20}
\]

\[
Q_g = C_g \ast V_{gs}. \tag{4.21}
\]

By substituting Eq. (2) into Eq. (1),

\[
P_{g,sw} = Q_g \ast V_{gs} \ast f_s. \tag{4.22}
\]

For the transistors that have been selected for building full bridge converters for DAB, the maximum power losses are given in Table 4.1. In Fig. 4.11, power losses of the gate for each transistor are plotted when all transistors have the same driving voltage at switching frequency equals 200 kHz. From the graph, it can be seen that the gate power losses for GaN-based devices
are higher for the transistors that have lower rated drain to source voltages. On the other hand, the Si-based transistor, which has the lowest rated $V_{ds}$, has the minimum gate power dissipation.

Table 4.1 Gate Circuit Losses

<table>
<thead>
<tr>
<th>Transistor</th>
<th>GaN</th>
<th>Si</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>GS66508T</td>
<td>GS61008T</td>
</tr>
<tr>
<td>$P_{g,sw}$ (mW)</td>
<td>6.96</td>
<td>14.4</td>
</tr>
</tbody>
</table>

Figure 4.11 Gate Drive Loss vs. Driving Voltage.

4.3.2 Gate Driver Integrated Circuit

The charging and discharging losses of the transistors discussed in the previous section and the required peak driving current, are the main parameters for selecting the gate drive IC. Furthermore, other parameters should be taken into account, these are: temperature range, input and output delay time, temperature monitor, miller clamp, negative power supply, and separate output. Because the miller phenomenon has a high effect when GaN transistors are used, the gate driver should have a miller clamp function in order to avoid shoot-through.

Based on the datasheet for the gate drive IC (Si827x) [77], the power dissipation in the gate driver IC is the sum of power dissipated by the bias supply current, the internal parasitic switching losses, and power dissipated by the series gate resistor and load. Approximately, the power loss of the gate driver IC is given by,
\[ P_d = f_s Q_g V_{drive} \left( \frac{R_p}{R_p + R_{g, on}} + \frac{R_n}{R_n + R_{g, off}} \right) \]  

(4.23)

where \( R_p \) is the \( R_{ds(on)} \) of the driver pull-up switch, \( R_n \) is the \( R_{ds(on)} \) of the driver pull-down switch, \( R_{g, on} \) and \( R_{g, off} \) are the gate external resistors during turning on and off respectively, and \( V_{drive} \) is the driver side supply voltage.

The pull-up and down resistors for the selected gate driver ICs are given as follows [77]–[79]. The gate driver IC for all the high voltage bridges is (Si8271GB-IS) \( (R_p = 2.7 \, \Omega, \, R_n = 1 \, \Omega) \). For driving EPC 2022, the gate drive IC (UCC27611) is used and the resistors are \( (R_p = 1 \, \Omega, \, R_n = 0.35 \, \Omega) \). The IC (LM5114 MF) with \( (R_p = 2.2 \, \Omega, \, R_n = 2.28 \, \Omega) \) is exploited to the low voltage Si transistor.

Fig. 4.12 shows the power losses of the gate drivers with switching frequency from 100 kHz to 500 kHz for the selected transistors. Assuming all other parameters are fixed, the power dissipation increases by increasing the switching frequency. The results for (Si8271GB-IS) IC at 200 kHz are given in Table 4.2, the power dissipation for (UCC27611) IC is 3.02 mW, and for (LM5114 MF) equals 91.72 mW.

Table 4.2 Gate Driver Losses

<table>
<thead>
<tr>
<th>Transistor</th>
<th>GS66508T</th>
<th>GS61008T</th>
<th>STL57N65M5</th>
<th>STL45N65M5</th>
<th>TK31V60W</th>
</tr>
</thead>
<tbody>
<tr>
<td>( P_d ) (Si8271GB-IS)</td>
<td>3.8 mW</td>
<td>7.86 mW</td>
<td>125.8 mW</td>
<td>107.44 mW</td>
<td>112.68 mW</td>
</tr>
</tbody>
</table>

Figure 4.12 Gate Drivers Loss vs. Switching Frequency.
It is obvious that the gate drivers have higher power dissipation when they are used to switch Si-based transistors because the total gate charge and the driving voltage are higher. By comparing the gate circuit losses from the last section with the gate driver losses, it is seen power dissipation in the ICs are higher than the power losses of the gate circuit except for the EPC transistor because the gate driver IC has very low pull-up and pull-down resistors.

4.4 Dual Active Bridge with Soft-Switching

Since the switching loss occurs because of the overlap of the voltage of the switch and the current flowing through the switch, soft switching techniques are used for reducing the dynamic dissipation for a transistor. Ideally, in the soft switching, the device changes its state from on to off or vice versa at zero voltage or zero current or both. Practically, the switching transitions happen at low voltage or current and lead to very low dissipations. The soft switching converters are divided into three major categories: resonant converters, resonant switch converters, and phase shifted converters [80]–[82]. In resonant converters, additional components are required and the switching frequency is adjusted for controlling the transferred power and for achieving soft switching. On the contrary, in phase shifted converters, the parasitic elements usually are utilized for soft switching purposes and the phase shift is adjusted for changing the transferred power as stated previously.

Accordingly, in phase shifted bridge converters, the control system enables the use of the parasitic elements of the main components (transistors and the transformer) during the soft transition. Likewise the conventional converters, the phase shifted converters operate at a constant frequency and lower voltage and current stresses. Therefore, the phase shifted converters behave like resonant converters during transitions modes and behave like conventional converters during the other modes.

The DAB converter is well known for its soft switching feature that is inherent in the converter and does not need to have any auxiliary components. The parasitic capacitors of the transistor and the body diode, in the case of using MOSFET or the transistor itself in case of using HEMT, are used for achieving ZVS along with the leakage inductance of the transformer. The influence of the dominant system parasitics, such the magnetizing inductance of the transformer and the device
capacitance on soft-switching converter of DAB converter is discussed in [59]. Performance analysis of the DAB converter under ZVS is presented in [83]. In [84], the ZVS boundaries for DAB is identified by using frequency domain analysis. The boundaries of soft switching of a DAB converter can be changed by changing the control method. A single phase-shifting strategy is applied in DAB in order to extend the soft switching region and decrease the inductor peak current comparing to the traditional phase shift strategy [85]. A voltage doubler circuit to extend soft switching range of DAB converters is proposed in [86]. In addition to use a dc blocking capacitor in the voltage doubler circuit, the gate signal sequence of the low side voltage bridge is rearranged for generating a high voltage in the low side bridge for reducing the circulating power and increasing the soft switching boundaries.

However, to achieve ZVS for all switches in a DAB, two conditions need to be fulfilled. First, the energy stored in the leakage inductance should be sufficient to charge and discharge the transistor capacitors. Second, the dead-time, which prevents the simultaneous switching of the same leg transistors, should not be more than the resonant process and not less than the time that is needed to discharge the capacitor before turning on the transistor.

The phase shift control leads to soft switching in the DAB converter as described along with ZVS process in the prior section 4.1. Fig. 4.13 shows the equivalent circuit of an H bridge of a DAB converter. During the dead-time, the inductor current charges the capacitor of the turning off transistor and discharges the capacitor of the other transistor of the same leg. When the transistors

![Figure 4.13 Equivalent Circuit of an H Bridge Converter](image)
(S1 and S4) are switched off, the capacitors (C1 and C4) are charged and the capacitors (C2 and C3) are discharged so the transistors (S2 and S3) are turned on under zero voltage.

The minimum inductor current that is required for achieving ZVS can be obtained from the required energy to charge and discharge the capacitors. By assuming the two voltages (V1 and V2) across the leakage inductance, as shown in Fig. 4.3, are equal, the minimum required current can be derived as follows,

\[ E_{Lg} = E_{c1} + E_{c2}, \]  

where \( E_{Lg} \) is the stored energy in the leakage inductance and \( E_{c1} \) and \( E_{c2} \) are the required energy to charge/discharge the capacitors (C1 and C2) respectively. Because the capacitors of the transistors are equal, Eq. (4.24) can be written as:

\[ E_{Lg} = 2E_c, \]  

\[ \frac{1}{2} L_g I_{Lg \_min}^2 = 2CV^2, \]  

\[ I_{Lg \_min} = \frac{2V}{Z}, \]

where \( I_{Lg \_min} \) is minimum required current, \( C = C_1 = C_2, V = V_1 = V_2 \), and \( Z = \sqrt{\frac{L_g}{C}} \).

Furthermore, the required dead-time can be determined from the resonant time of the leakage inductance and the capacitors of the transistors. So, the dead-time \( (t_{dead}) \) is given as:

\[ t_{dead} = \frac{\pi}{2} \sqrt{2L_g C}. \]

For the selected transistors and by using the output capacitances from the appropriate datasheets, the minimum required current and the needed dead-time for attaining turn on ZVS are calculated as given in Table 4.3.

<table>
<thead>
<tr>
<th>Transistor</th>
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<th>Si</th>
</tr>
</thead>
<tbody>
<tr>
<td>GS66508T</td>
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<tr>
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<tr>
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<td></td>
</tr>
<tr>
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<td>2.3</td>
</tr>
<tr>
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<td>2.12</td>
<td>2.4</td>
</tr>
<tr>
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<td></td>
</tr>
<tr>
<td>TK31V60 W</td>
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<td></td>
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<tr>
<td>BSC035N1</td>
<td></td>
<td></td>
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<tr>
<td>0NS5ATM1</td>
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<table>
<thead>
<tr>
<th>( t_{dead} ) (ns)</th>
<th>analytical</th>
<th>simulation</th>
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<tr>
<td>GS66508T</td>
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</tr>
<tr>
<td>GS61008T</td>
<td>42</td>
<td>34.7</td>
</tr>
<tr>
<td>EPC2022</td>
<td>72</td>
<td>68</td>
</tr>
<tr>
<td>STL57N65 M5</td>
<td>70</td>
<td>-</td>
</tr>
<tr>
<td>STSL45N65 M5</td>
<td>64</td>
<td>-</td>
</tr>
<tr>
<td>TK31V60 W</td>
<td>59</td>
<td>-</td>
</tr>
<tr>
<td>BSC035N1 0NS5ATM1</td>
<td>69</td>
<td>-</td>
</tr>
</tbody>
</table>

Table 4.3 ZVS Current and Dead-time
Fig. 4.14 shows the simulated values for the dead-time for the GaN-based transistors by using the software LTspice. The LTspice files for the selected Si-based transistors are not provided; therefore, the simulated values for the dead-time are not included.

Figure 4.14 Simulation Values for the Dead-time for the GaN-based Transistors
4.5 Summary

The operation principles of a DAB converter depend upon the difference between the voltages across the leakage inductance of the transformer. As a result, the amount of the transferred power can be adjusted by controlling the phase shift angle (δ), the duty cycles of the two bridges, and the switching frequency. On the other hand, the direction of the power flow can only be controlled by changing the phase shift. In the phase shift switching method, the voltage of the leakage inductance of the transformer has four states that determine the average current through it. Power flow analysis according the phase shift switching was presented in this chapter in addition to the zero-voltage switching feature in a DAB. Moreover, the switching and conducting losses were discussed.
CHAPTER FIVE

5. A 4 kW Dual Active Bridge Converter

In this chapter, two DAB converters are presented, namely, a GaN-based and a Si-based converters. The selection of the transistors for the GaN and Si boards is discussed. The power losses for each converter are calculated theoretically and verified by the simulation with either two DC sources or a DC source on one side and a pure resistor on the other side.

The converter is to be designed according to the following specifications: low side voltage \( V_{LV} = 48 \) V, high side voltage \( V_{HV} = 400 \) V, and the maximum transferable power is supposed to be \( P_{out} = 5 \) kW. According to Eq. (4.15), the power is inversely proportional to the inductance of the transformer and the switching frequency. Therefore, the value of the parameters \( (2\pi f_s L_g) \) is calculated to make the transferred power to the load whether it is a resistor or another source is around 4 kW for simulation purposes. In the experimental result, the maximum transferrable power is determined by changing the switching frequency only because the transformer has a fixed leakage inductance.

To select components for the converter, the current needs to be calculated. The maximum transferrable power can be achieved when the phase shift is equal to 90°. To obtain 4 kW at \( V= 48 \) V and by using Eq. (4.15), the switching frequency with the leakage inductance should equal \( f_s L_g = 0.072 \). By applying Eq. (4.17), the RMS current of any transistor of the low voltage bridge is equal to \( I_{d_{\text{rms}}} \approx 96.23 \) A and the RMS current of a transistor of the high voltage bridge is equal to \( I_{d_{\text{rms}}} \approx 11.55 \) A. It is worth mentioning here, the RMS current of the transformer is calculated and it is equal to 136 A for the low voltage side and 16.33 A for the high voltage side.

5.1 Switches Selection

The first step for designing a SMPS is the selection of the transistor. In this section, the transistors’ rating requirements are specified. The main parameters taken into account are the rated and maximum voltages, current, maximum operating frequency, and driving voltage. Seven groups of transistors are selected for the two converters. Exploiting datasheets and applying the equations
support to find the operating area for each transistor. For the GaN-based converter, the transistors of the high voltage bridge are from GaN Systems and the low voltage side bridges transistors are from GaN Systems and EPC. For the Si-based converter, the MOSFETs from STMicroelectronics and Toshiba are selected for the high voltage bridges and low voltage side transistors are from Infineon. Tables 5.1 [87]–[90] shows the important parameters for each transistor.

Table 5.1 Selected Transistors Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>symbol</th>
<th>GaN</th>
<th>Si</th>
</tr>
</thead>
<tbody>
<tr>
<td>Drain-source voltage</td>
<td>Vds (V)</td>
<td>GS66 508T GS610 008T</td>
<td>STL57N65 M5 STL45 N65M5 TK31V6 0W BSC035N10 NS5ATM1</td>
</tr>
<tr>
<td>Drain current</td>
<td>I_d (A)</td>
<td>30 90 90</td>
<td>22.5 22.5 30.8 100</td>
</tr>
<tr>
<td>Drain-source on-resistance</td>
<td>R_{ds,on} (mΩ)</td>
<td>50 7 3.2</td>
<td>61 75 78 3.5</td>
</tr>
<tr>
<td>Gate-source voltage</td>
<td>Vgs (V)</td>
<td>-10 to 7 -10 to 7</td>
<td>±25 ±25 ±30 ±20</td>
</tr>
<tr>
<td>Gate-source threshold voltage</td>
<td>V_{gs_th} (V)</td>
<td>1.7 1.3 1.4</td>
<td>4 4 2.7-3.7 3</td>
</tr>
<tr>
<td>Input capacitance</td>
<td>C_{iss} (pF)</td>
<td>260 590 1400</td>
<td>4200 3470 3000 5000</td>
</tr>
<tr>
<td>Output capacitance</td>
<td>C_{oss} (pF)</td>
<td>65 280 840</td>
<td>100 82 70 770</td>
</tr>
<tr>
<td>Reverse capacitance</td>
<td>C_{rss} (pF)</td>
<td>2 12.4 7</td>
<td>6 7 9.5 34</td>
</tr>
<tr>
<td>Total gate charge</td>
<td>Q_g (nC)</td>
<td>5.8 12 13</td>
<td>96 82 86 70</td>
</tr>
<tr>
<td>Reverse recovery charge</td>
<td>Q_r (nC)</td>
<td>0 0 0</td>
<td>7000 6000 3500 122</td>
</tr>
</tbody>
</table>

From Table 5.1, it is obvious the high voltage GaN based transistors have lower on-resistance, capacitances, and gate charge, although the drain to source voltages and the drain current are almost the same except for the Toshiba transistor. Thus, the losses calculated theoretically for the GaN-based converter are less than the Si-based converter. Alternatively, in the experimental testing, many factors affect GaN-based devices, such as parasitic inductances of the printed circuit board and the packages. Therefore, the power loss of the GaN-based converter might be higher than the Si-based converter.
5.2 Power Loss

The theoretical power losses and the efficiency for the DAB converter are presented in this section. The maximum RMS current occurs at a 90° phase shift. Consequently, the efficiency of the converter decreases when an increase in the output power occurs because of the increase in the switches’ current, which leads to an increase in the conducting losses. However, depending on the converter itself and the parameters of the transistors that are used in the bridges, sometimes the output power decrease may contribute in power loss increase because of switching loss, which is a result of losing the zero-voltage switching feature.

By applying Eq. (4.18), the conducting power loss for each transistor at the maximum transferred power is given as follows: the conducting loss for the high voltage GaN transistor is equal to 6.67 W, the conducting loss for the low voltage GaN and EPC transistors are approximately equal to 64.8 W and 29.63 W, respectively, the high voltage Si transistors STL57N65M5, STL45N65M5, and TK31V60W have conducting losses equal to 8.133 W, 10 W, and 10.4 W respectively, and the loss for the low voltage Si transistor is 32.41 W. Fig. 5.1(a) shows the curves of the conducting power loss for the high voltage GaN and Si transistors and Fig. 5.1(b) shows the conducting power loss for low voltage transistors for a different phase shift. The power losses of the output parasitic capacitors ($P_{Coss}$) of the transistors are given as:

$$P_{Coss} = 0.5 \ C_{oss}V^2$$

(5.1)
As a result, the total power loss at 100 kHz for the GaN-EPC-based converter for all transistors is about 147.7 W, for GaN-GaN-based converter is 288 W, and for the Si-based converter is about 173 W. Fig 5.2 illustrates the total power loss for the converters vs. the transferred power.

![Graph showing total power loss vs. transferred power for different converters.](image)

**Figure 5.2 Total Power Loss vs. Transferred Power**

For the assumed conditions, switching frequency and leakage inductance, the efficiency of the converter whether it is GaN or Si based is very low. Although the conducting losses 3.62% for the GaN-EPC and 4% for the Si of the maximum transferred power, the high circulating current in the transformer and the high ripple current in the input or the output of the DAB, increase losses in the other components of the converter. Therefore, two types of efficiency are considered. The first type is obtained by dividing the transferred power which can be calculated from the Eq. (4.15) by the input power which can be calculated from Eq. (5.1).

\[ P_{in} = V I_{rms}, \]  

(5.1)

where \( V \) is any side voltage (the high side voltage = 400 V and the low side voltage = 48 V). So, the efficiency of the converter is very low. According to Eq. (5.2), the efficiency of the converter is 61.24% at 4 kW. Eq. (5.2) is plotted in Fig 5.3.

\[ Eff_1 (%) = \frac{P_{out}}{P_{in}} * 100\%. \]  

(5.2)
The second type of efficiency can be determined by adding the total conducting losses and $P_{\text{Coss}}$ to the transferred power (4.15) and by considering it as the input power for the converter. In this type of efficiency, the circulating current and ripple are not taken into account. Type 2 efficiency is given by Eq. (5.3). At the maximum transferred power, the GaN-EPC-based converter efficiency is 96.5 %, GaN-GaN-based converter efficiency is 93.3 %, and the Si-based converter efficiency is 96.1 %, as shown in Fig. 5.4. The curves start when the output power is around 150 W.

$$Eff_2 = \frac{P_{\text{out}}}{P_{\text{out}} + P_{\text{con}} + P_{\text{Coss}}} * 100\%.$$  \hspace{2cm} (5.3)

Figure 5.4 Type 1 Efficiency

Figure 5.3 Type 2 Efficiency
The measured leakage inductance of one of the transformers that have been manufactured in the laboratory is 10 µH. Thus, a reasonable phase shift and reasonable power transfer cannot be achieved at the same time. Assuming that the lowest switching frequency that can be used is 200 kHz and the power is 2000 W without saturating the transformer, the required phase shift is about 10.4°, when \( V_1 \) and \( V_2 \) across the leakage inductance are equal. In addition, the conducting loss will be very low according to the low RMS current. Fig 5.5 (a) shows the losses for the converter and Fig. 5.5 (b) shows the efficiency for the DAB converter excluding the loss of the abovementioned transformer. The curves start when the output power is about 100 W.

![Diagram](image_url)

Figure 5.5 DAB converters with the Lab Transformer (a) Power losses (b) efficiencies
5.3 Simulation

This section presents several simulations for various scenarios according to the different applications of the DAB. The simulation was done by using the simulation package PowerSim (PSIM). Fig. 5.6 illustrates the DAB basic circuit that used in PSIM. Because the circuit is an open control loop, the transistors are driven by square wave sources for simplicity; so, it is easy to change the switching frequency and duty cycle for each leg separately.

![DAB PSIM Basic Circuit](image)

**Figure 5.6 DAB PSIM Basic Circuit**

Two small resistors are connected in series with the sources to eliminate the DC offset in the primary and secondary voltages of the transformer and the voltage of the leakage inductance as well. The voltage source that drives the transistors, $S_1$ and $S_4$, is considered as the reference for the other three sources. The second source, which controls the $S_2$ and $S_3$, has a fixed phase difference equal to $180^\circ$ with the reference source. For the high voltage H bridge, the control voltage of $Q_1$ and $Q_4$ is shifted by $\delta$ according to the reference source and the control voltage of the transistors $Q_2$ and $Q_3$ has phase difference equal to $180^\circ + \delta$.

To verify the conducting loss (calculated in the previous section) the switching frequency is changed to meet the requirements for transferring 4 kW. The direct method for comparing the analytical conducting losses with the simulation value is through comparing the calculated RMS current with the simulated value. From the previous section, the calculated RMS currents for the transformer at 4 kW are 136 A and 16.33 A. The simulated values are 138 A and 17.2 A for the low voltage side and high side of the transformer respectively, as shown in Fig. 5.7.
Fig. 5.8 shows the high ripple currents in the input and output because of the high circulating current in the leakage inductance. Thus, a large capacitor is required to absorb the high ripple in the current and the value of the capacitor increases by increasing the transferred power. The simulation was done without connecting a capacitor to either side of the converter, in order to observe the case that produces the highest ripple and makes the RMS current of the primary and secondary of the transformer equal to the RMS current of the sources. Many control methods have been presented to minimize the circulating current [91]–[94] but they will not be discussed in this research.
For simulation the DAB by using the parameters of the laboratory transformer, the maximum transferrable power is at 200 kHz. The first scenario simulates the converter with two DC sources 48 V and 400 V connected on each bridge. The simulated RMS current of the low side of the transformer is 44.12 A, which is very close to the calculated value 43.36 A. Fig. 5.9 presents the simulated waveforms.

![Figure 5.9 DAB with the Two Sources](image)

The second scenario replaces the high voltage source with a pure resistor. The value of the resistor is calculated by dividing the voltage of the high side by the RMS current from the previous scenario. Thus, the resistor value is about 72.72 Ω. From Fig. 5.10, it noticed that the currents of the components are less than the previous case.

![Figure 5.10 DAB with a Resistor on the High Side](image)
The third scenario replaces the low voltage source with a resistor where the value obtained in the same way as the last case. The resistor value is 1.09 Ω. Fig. 5.11 shows the current of the low side of the transformer and the input and the resistor currents.

![Graph showing current of transformer and input and resistor currents](image)

Figure 5.11 DAB with a Resistor on the Low Side

### 5.4 Summary

In the previous chapter, the selection of the transistors for the GaN-based and Si-based were discussed. The converters were assumed to operate under zero voltage condition, therefore, the losses and efficiencies were calculated by neglecting the switching losses. Furthermore, the other components such as the transformer and the DC-link capacitors were considered ideal when the three converters were compared. The bridge with transistors that have lower on-resistance has better performance in term of the efficiency. The analytical values of the efficiencies of the converters were verified by the simulation. The converter was simulated with different scenarios. It is found that the scenario that uses the DAB converter for stepping up the voltage with a resistor connected to the high voltage bridge has the lowest conduction losses although transferring the same amount of power in the other scenarios.
CHAPTER SIX

6. Experimental Results

After verifing the analytical results by the simulation for the DAB and calculating the minimum requirements for achieving ZVS, the GaN-based DAB will be compared with its Si-based counterpart when they operate with the same switching frequency. In this chapter, different PCBs are presented for the high and low voltage H bridges for constructing a DAB. The experimental results for the GaN-based and Si-based DAB converters are discussed in this chapter.

The first iteration of the hardware prototype of the GaN-based high voltage bridge is shown in Fig. 6.1. In this iteration, the integrated circuits (IXDN609SI) from IXYS are used for driving the transistors. An adjustable driving voltage is obtained from the low dropout regulator (LDO) (XC6216C202PR-G) which is from Torex Semiconductor. Three capacitors (0.22 µF, 4.7 µF and 0.1 µF) are placed close to the gate driver as decoupling capacitors for suppressing the noise in the driving voltage. A diode and resistor are connected in antiparallel with the gate drive resistor to provide a low turn off resistance. The isolators (PES1-S5-S12-M-TR) and (SI8410BB-D-IS) provide the power and signal isolation respectively. The DC-link consists of four film capacitors (30 µF each) and twenty ceramic capacitors (2.2 µF each), all capacitors are connected in parallel in order to minimize the equivalent series inductance (ESL) and equivalent series resistance (ESR) compared to the using one large capacitor.

Figure 6.1 GaN First Iteration Prototype
The four transistors are located on the bottom layer of the board, as shown in Fig. 6.2. On the top layer, the gate drivers, isolators, and voltage regulators are placed. The heatsink is attached directly to the top of the all transistors for dissipating heat.

![Figure 6.2 GaN First Iteration Transistors Layout](image)

The problem with the above layout is the negative overshoots in the gate to source voltage and drain to source voltage during switch transition states, as seen in Fig. 6.3 (a) and (b). The overshoots occur when the input voltage equals 40 V and they enlarge by increasing the voltage and can cause damage to the transistors. The reasonable explanation for the spikes is the discharging of the Miller capacitors through the large parasitic inductance of the PCB. The inductances of the power loops are expected to be large because of the relatively long distances among the transistors especially the distance between the transistors of one leg (Q1 & Q3 and Q2 & Q4), it can be seen in Fig. 6.2.

![Figure 6.3 First Iteration Vgs and Vds Spikes](image)

(a) turn on (b) turn off
The second iteration prototype is shown in Fig. 6.4. Many changes have been made in order to minimize the parasitic inductance in the power loop and reduce the ringing in the gate drive voltage as much as possible. The distance between the transistors of one leg has been reduced to 3 mm. Furthermore, ceramic capacitors (2.2 µF) are located very close to each transistor in addition to two film capacitors (95 µF each), and those are connected between the positive and the negative terminals of the input voltage. Transistors and gate drivers are placed on the same layer. Isolated gate drive IC (Si8271GB.IS) from Silicon Labs is used because it has many advantage such as the isolation and under-voltage lockout (UNLO). Moreover, turn on and off circuits are separate; therefore, an antiparallel diode is not required. The turn on resistor is (10 Ω) and the turn off resistor is (2 Ω). Two capacitors (10 µF and 47 µF) are connected across the input supply voltage and the ground. Two low ESL (10 µF) capacitors from TDK Corporation and (0.1 µF) capacitor are connected across the driver supply voltage and the driver ground for decoupling purposes. Additional (47 µF) capacitors are attached to the input for each isolated power supply.

![Figure 6.4 GaN Second Iteration Prototype](image)

Fig. 6.5 shows the gate voltage and drain to source to voltage of the second iteration. It is noticed that both power and gate loops have improved and the parasitic inductances have been minimized. In this layout, the negative spike in the gate voltage starts at input voltage around 70 V.
Two Si-based high voltage boards were created by using different transistors (STL45N65M5) and (TK31V60W) as shown in Fig. 6.6. The gate driver ICs which are used in these PCBs are the same as the ICs that used for high voltage GaN-based board. In those two boards, the gate drive voltages (12 V) are obtained directly from the isolated power supply; therefore, LDOs are not required. In addition to the gate drivers’ decoupling capacitors, two capacitors (1 µF and 0.22 µF) are connected between the positive and ground of each gate drive voltage supply in order to suppress the noise. The gate input signal return for each transistor is separate from the main current flow as it seen in appendix B.
The gate to source voltage and the drain to source voltage (during transition at 90 V input voltage of one of the Si transistors of both boards) are shown in Fig. 6.7 (a) and (b). It can be seen that the noise in the waveforms is very low. The proper operation frequency of these transistors is 20 kHz; therefore, the switching losses increase dramatically as the switching frequency increases. Furthermore, the temperature of the transistors exceeds the rated value the switching frequency increases above 50 kHz. Therefore, soft switching is required for wide ranges of the transferable power and input and output voltages when they are used to construct a DAB with high switching frequency.

For the low voltage bridge, two prototypes were created, EPC devices are used in a prototype and Si devices are used in the other. Each board consists of three PCBs: the mother board, which contains the power supplies and the DC-link capacitors, and the daughter boards, which contain the transistors, the digital isolators, and ceramic capacitors. Fig. 6.8 shows the two low voltage
prototypes. The DC-link of each board consists two film capacitors (220 µF each) and twenty ceramic capacitors (4.7 µF each) placed on the mother board and the bottom layers of the daughter boards.

The main challenge of utilizing EPC transistors on a PCB is the assembling. The Land Grid Array (LGA) packaging requires special tools for soldering. A heat gun was used only for soldering. Fig. 6.10 shows two EPC devices on the board. Furthermore, the expected high current (100 A) that flows through the devices tends to heat the traces and increase the temperature of the device. So, all pads need to be soldered on the board properly. The excessive high temperature, because of bad connection, damages the transistors and the board as well, as shown in Fig. 6.10.
Moreover, a GaN-based low voltage prototype was created, as shown in Fig. 6.11. The transistors (GS61008T) form GaN systems are used. The gate drive circuit and the voltage for each transistor is the same as the 2nd iteration of the GaN-based high voltage board. In addition to the ceramic capacitors (2.2 µF), two film capacitors (220 µF each) are used to build the DC-link. A heatsink is attached to the board because the transistors have relatively high on resistance which leads to raise the temperature of the transistors.

![GaN-based Low Voltage Board](image)

Figure 6.11 GaN-based Low Voltage Board

The waveforms of the voltages $V_{gs}$ and $V_{ds}$ for the GaN-based low voltage board are shown in Fig. 6.12 when the input voltage equals 48 V. It is seen that the gate voltage has an overshoot around 1 V during the turning off transition because of the Miller capacitor effect especially with low $V_{ds}$.

![Low Voltage GaN Transistor Transition Waveforms](image)

Figure 6.12 Low Voltage GaN Transistor Transition Waveforms
The full bridge boards are used to construct three DAB converters as follows: GaN-EPC, GaN-GaN, and Si-Si converters. A DAB test setup is shown in Fig. 6.12. For the GaN-EPC DAB converter, the maximum output power reached was 723.65 W with efficiency 96% at input voltage about 300 V. Because the current was around 20 A and because of soldering issues, the EPC transistors did not survive with that power. The maximum power reached for with the GaN-GaN converter was the 860 W at input voltage around 330 V, but the losses were very high; therefore, it is not used to build a DAB for comparison purposes.

![DAB Test Setup](image)

Figure 6.13 DAB Test Setup

However, the more success was achieved with GaN-Si and Si-Si converters. Therefore, they are used to compare the performance DAB converters. The first test was to use the DAB as a step-down converter by connecting the power source to the high voltage board and the load (resistor) to the low voltage board. The experimental results for two different values of the phase shift are given in table 6.1 for an input voltage ~ 400 V. Fig. 6.13 (a) shows the low and high voltage and the current of the transformer for the Si-based converter and the GaN-based waveforms are shown in Fig. 6.13 (b).

![Waveforms](image)

Table 6.1 Step-down Experimental

<table>
<thead>
<tr>
<th>DAB</th>
<th>$I_{in}$ (A)</th>
<th>$V_{out}$ (V)</th>
<th>$I_{out}$ (A)</th>
<th>$\eta$ (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Si-Si</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>TK31V60W</td>
<td>2.85</td>
<td>55.91</td>
<td>19</td>
<td>92.3</td>
</tr>
<tr>
<td></td>
<td>3.67</td>
<td>57.8</td>
<td>23.6</td>
<td>92.2</td>
</tr>
<tr>
<td>STL45N65M5</td>
<td>2.83</td>
<td>54.55</td>
<td>19.2</td>
<td>92.4</td>
</tr>
<tr>
<td></td>
<td>3.64</td>
<td>57</td>
<td>23.3</td>
<td>91</td>
</tr>
<tr>
<td>GaN-Si</td>
<td>2.8</td>
<td>58</td>
<td>17.7</td>
<td>92</td>
</tr>
<tr>
<td></td>
<td>3.3</td>
<td>57.6</td>
<td>21.3</td>
<td>93</td>
</tr>
</tbody>
</table>
From the transformer current waveforms, it is noticed that the voltage of the secondary side (low voltage) of the transformer is higher when the GaN board is used because the dead-time of Si transistors is longer. In spite of the difference in the output voltages for the two converters the efficiencies are the same.

The second test uses the DAB as a step up converter. The source was connected to the low voltage board and the load was connected to the high voltage bridge. Table 6.2 shows the experimental results of the converters with output voltage ~ 400 V. The high voltage winding has higher voltage when GaN bridge is used. In other word, the difference between the voltages across the leakage inductance of the transformer is more in case of the Si-Si converter, as shown in Fig. 6.15 (a) and (b).

Figure 6.14 Step-down Experimental Transformer Waveforms (a) Si-Si (b) GaN-Si
From the experimental tests, the temperature of transistors of the board is higher when it is connected to the source than the load. For instance, the temperature of the low voltage transistors in step-up test was higher than when the DAB is used as step-down converter. Moreover, the temperature of the high voltage Si transistors exceeds the rated value when the transferred power is less than 300 W because of the loss of the ZVS feature. In contrast, the temperature of the GaN transistors does not increase even when the transferred power is very low.

In summary, the performance of the above-mentioned Si-based and GaN-based converters are the same in terms of the efficiency and power density unless the switching frequency increases more than 200 kHz.
CHAPTER SEVEN

7. Conclusion and Future Work

In this study, the objective was to compare a GaN-based DAB converter with its Si-based counterpart for interfacing energy storage systems to PV applications. The principles of operation of the DAB converter were discussed in addition to the criteria for selecting the suitable bidirectional converter.

Generally, renewable energy sources, the especially PV module, whether they are on or off grid, are connected with energy storage systems because of the fluctuating nature of these sources and because of the difference between the demanded power and generated power. Isolated bidirectional converters (IBDC) are the most applicable converters for such applications for safety purposes and interfacing the wide voltage ranges. IBDCs increase the system efficiency and power density in addition to the reducing the system complexity. The DAB converter has been chosen among the other topologies because it is more suitable for high power applications. For improving its performance, GaN-based transistors are used and compared with Si-based converter in order to validate the improvement. The biggest challenge in designing DAB is the circulating current that results from the leakage inductance of the transformer. The efficiency of the DAB decreases with increasing circulating current. In addition to DAB challenges, use of GaN transistors complicates the converter and increases the obstacles for achieving a high power transfer.

It is noted from the transistors parameters and the results from the last chapters (5 and 6) that, if the two converters operate with the same switching frequency and both of them achieve ZVS, the converters have the same performance. The 0.5 % difference in the efficiency can be neglected because of the limitations of using GaN transistors. Therefore, for improving DAB performance by using GaN transistor, achieving high switching frequency is the main point in order to increase the power density of the converter.

For future work, the two converters can be compared in terms of electromagnetic interference (EMI). In addition, two inductors can be connected in series with the two sources to create a current-fed DAB to minimize the pulsating currents and study the effects of these inductors on the transferred power and the performance of the converter. Enhance the design the EPC board and
improve the soldering method to the PCB. Because the low voltage side bridge experiences a very high current around 100 A at full load, the H bridge can be replaced by two buck converters by using two inductors as shown in Fig. 7.1 (a). In this case, the current stress of each transistor will be around half of the output current of the converter. Moreover, it is thought that the two inductors can be replaced by one inductor as shown in Fig. 7.1 (b).

Figure 7.1 Double-Buck DAB Converter (a) Two Inductors (b) One Inductor
References


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Appendix A. Derivation of the RMS Current and Average Power of a DAB

From the figure below and by assuming the voltages $V_1$ and $V_2$ are not equal, the RMS current of the leakage inductance of the transformer is given as:

![Leakage Inductor Waveform](image)

$$I_{LG,RMS} = \sqrt{\frac{1}{\pi} \left[ \int_0^\delta \left( \frac{V_1 + V_2}{L_g} t - I_0 \right)^2 dt + \int_\delta^{\pi} \left( \frac{V_1 - V_2}{L_g} t - I_\delta \right)^2 dt \right]}$$

$$= \frac{1}{\pi} \sqrt{\left[ \left( \int_0^\delta \left( \frac{V_1 + V_2}{L_g} t^2 \right)^2 dt \right) - \left( \int_0^\delta \left( \frac{V_1 + V_2}{L_g} t \right)^2 dt \right) + \left( \int_\delta^{\pi} \left( \frac{V_1 - V_2}{L_g} t^2 \right)^2 dt \right) + \left( \int_\delta^{\pi} \left( \frac{V_1 - V_2}{L_g} t \right)^2 dt \right) \right]}$$

$$= \frac{1}{\pi} \sqrt{\left[ \left( \frac{1}{3} \left( \frac{V_1 + V_2}{L_g} \right)^2 \right)^{\delta} - \left( \frac{1}{3} \left( \frac{V_1 + V_2}{L_g} \right)^2 \right)^{\pi} + \left( \frac{1}{3} \left( \frac{V_1 - V_2}{L_g} \right)^2 \right)^{\delta} + \left( \frac{1}{3} \left( \frac{V_1 - V_2}{L_g} \right)^2 \right)^{\pi} \right]}$$
\[
\sqrt{\left[ \frac{1}{3} \left( \frac{V_1 + V_2}{L_g} \right)^2 \delta^3 \right] - \left( \frac{V_1 + V_2}{L_g} I_0 \right) \delta^2 + \left( I_0^2 \delta \right) + \left( \frac{V_1 - V_2}{L_g} I_\delta \right) (\pi - \delta)^2 + \left( I_\delta^2 (\pi - \delta) \right)} \tag{A.1}
\]

The instantaneous currents \(I_0\) and \(I_\delta\) can be obtained as:

\[
I_\delta + I_0 = \frac{V_1 + V_2}{L_g} \delta
\]

\[
I_0 = \frac{V_1 + V_2}{L_g} (\pi - \delta) + I_\delta \tag{A.2}
\]

\[
I_\delta + \frac{V_1 + V_2}{L_g} (\pi - \delta) + I_\delta = \frac{V_1 + V_2}{L_g} \delta
\]

\[
I_\delta = \frac{V_1 - V_2}{2L_g} (\pi - \delta) + \frac{V_1 + V_2}{2L_g} \delta \tag{A.3}
\]

The average transferred power is obtained from the average current and the voltage.

\[
P_{out} = V \cdot I \tag{A.4}
\]

The leakage inductance current has two modes:

Mode 1:

\[
i(\theta) = \frac{V_1 - V_2}{X_{L_g}} \theta + i(0)
\]

Mode 2:

\[
i(\theta) = \frac{V_1 - V_2}{X_{L_g}} (\theta - \delta) + i(\delta)
\]

The average current is given as:

\[
I = \frac{1}{2\pi} \int_0^{2\pi} i(\theta) \, d\theta
\]
\[
\frac{1}{\pi} \left( \int_{\delta}^{\pi} \left( \frac{V_1 V_2}{X_{lg}} \theta + i(0) \right) d\theta \right) \]

\[
= \frac{1}{\pi} \left( \left( \frac{V_1 V_2}{X_{lg}} \frac{\theta^2}{2} + i(0) \theta \right)_{0}^{\pi} \right) + \left( \frac{V_1 V_2}{X_{lg}} \left( \frac{\theta^2}{2} - \delta \theta \right) + i(\delta) \theta \right)_{\delta}^{\pi} \]

\[
= \frac{1}{\pi} \left( \left( \frac{V_1 V_2}{X_{lg}} \frac{\theta^2}{2} + i(0) \theta \right)_{0}^{\pi} \right) + \left( \frac{V_1 V_2}{X_{lg}} \left( \frac{\theta^2}{2} - \delta \theta \right) + \left( \frac{V_1 V_2}{X_{lg}} \delta + i(0) \theta \right) \right)_{\delta}^{\pi} \]

\[
= \frac{1}{\pi} \left( \frac{V_1 V_2}{X_{lg}} \frac{\delta^2}{2} + i(0) \delta + \frac{V_1 V_2}{X_{lg}} \left( \pi^2 - \delta^2 \right) + \frac{V_1 V_2}{X_{lg}} (\delta^2 - \delta \pi) \right) + \left( \frac{V_1 V_2}{X_{lg}} \delta + i(0) \theta \right) - \frac{V_1 V_2}{X_{lg}} \delta^2 + i(0)(\pi - \delta) \right) \]

\[
= \frac{1}{\pi} \left( \frac{V_1 V_2}{X_{lg}} \left( \delta^2 - \frac{\delta^2}{2} \right) + \frac{V_1 V_2}{X_{lg}} \left( \pi^2 - 2\delta \pi + \frac{\delta^2}{2} \right) + \frac{V_1 V_2}{X_{lg}} \delta \pi + i(0) \left( \delta \pi - \frac{\delta^2}{2} \right) \right) \]

\[
= \frac{1}{\pi} \left( \frac{V_1 V_2}{X_{lg}} \left( \delta - \frac{\delta^2}{2} \right) + \frac{V_1 V_2}{X_{lg}} \left( \pi - \delta \right) + \frac{V_1 V_2}{X_{lg}} \left( \pi - \delta \right) + \frac{V_1 V_2}{X_{lg}} \left( \frac{\delta^2}{2} - \frac{\delta^2}{2} \right) + i(0) \right) \]

\[
= \frac{1}{\pi} \left( \frac{V_1 V_2}{X_{lg}} \left( \delta - \frac{\delta^2}{2\pi} \right) + \frac{V_1 V_2}{X_{lg}} \left( \frac{\delta}{2} - \frac{\delta^2}{2\pi} \right) + \frac{V_1 V_2}{X_{lg}} \left( \frac{\delta}{2} - \frac{\delta^2}{2\pi} \right) + i(0) \right) \]

\[
= \frac{1}{\pi} \left( \frac{V_1 V_2}{X_{lg}} \left( \delta - \frac{\delta^2}{2\pi} \right) + \frac{V_1 V_2}{X_{lg}} \left( \frac{\delta}{2} - \frac{\delta^2}{2\pi} \right) + \frac{V_2}{X_{lg}} \left( \frac{\delta}{2} - \frac{\delta^2}{2\pi} \right) \right) \]

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From the equations (A.4) and (A.5), the average transferred power is given as:

\[ P_{\text{out}} = \frac{V_1 V_2}{X_{Lg}} \left( \frac{\delta - \delta^2}{\pi} \right) \]  

(A.4)

\[ l = \frac{V_2}{X_{Lg}} \left( \delta - \frac{\delta^2}{\pi} \right) \]  

(A.5)

**Appendix B. PCB Layouts for the H bridges converters**

In this section, the layout of the layer that has the power transistors for each board and the silkscreen for that layer are shown. From the two images for each layout, it can be seen the positions of the transistors and the gate driver ICs in addition to other components.

Figure B.1 GaN Top Layer Layout
Figure B.2 Si Top Layer Layout

Figure B.3 EPC Top Layer Layout