



ESCUELA TÉCNICA SUPERIOR DE INGENIERÍA (ICAI)

GRADO EN INGENIERÍA ELECTROMECÁNICA

DISEÑO Y SIMULACIÓN DE UN CONVERTIDOR DC-DC PARA APLICACIONES FOTOVOLTAICAS

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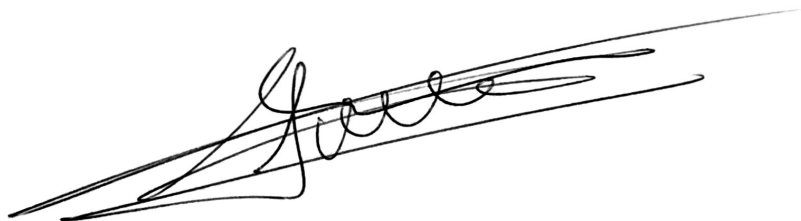
Julio 2018

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Resumen

Sinopsis

El objetivo de este proyecto de fin de grado consiste en el diseño y simulación de un convertidor DC-DC para aplicaciones fotovoltaicas. Se ha elegido la topología "Dual Active Bridge (DAB)" entre otros motivos por su carácter bidireccional y para satisfacer las necesidades de la aplicación deseada. Ésta es la conexión en paralelo de un elemento almacenador de energía como una batería, al panel fotovoltaico. Este paralelo irá conectado a la red a través de otro convertidor para transformar la corriente continua en alterna.

La topología elegida, el DAB, conectará la batería a un bus de continua de alta tensión, con la intención de, suministrar la cantidad correcta de energía a la red cuando exista un "mismatch" entre la energía que pueden proporcionar los paneles y la que necesita la red en ese momento. Esto podría suceder en caso de que, por ejemplo, apareciera una nube que hiciera que los niveles de potencia suministrada bajaran. Se realizó un estudio detallado de la topología para poder diseñar los componentes del convertidor teniendo en cuenta las estrategias de "Zero Voltage Switching (ZVS)" que mejorarán la eficiencia.

El proyecto fue formulado como una versión a escala de lo que podría ser un convertidor real por razones de seguridad en el laboratorio. Por esa razón, se han encontrado algunos problemas con el dimensionamiento de algunos componentes como el transformador. Parte de los próximos pasos que se deberían seguir son el diseño del control y la búsqueda de una solución para el diseño del transformador, para que al final el convertidor pueda llegar a ser construido.

Palabras clave: Dual Active bridge; DAB; fotovoltaico; bidireccional; soft-switching; ZVS; DC-DC.

Introducción

Actualmente, la demanda de fuentes de energía con menor impacto medioambiental se encuentra en alza. La idea de explotar los recursos naturales para la obtención de energía con la consecuente reducción de la contaminación siempre ha estado sobre la mesa, pero los problemas entorno a la viabilidad económica de dichos proyectos han sido también una constante. De hecho, este hecho ha sido el responsable del lento proceso de adopción e implementación de energías renovables.

Aunque existen diferentes enfoques para este problema, hay distintos países están apostando por diferentes tipos de energía, las plantas eólicas y fotovoltaicas (en adelante PV) son aquellas que están ganando mayor popularidad. En el caso de las instalaciones PV, se están desarrollando nuevas tecnologías que se enfocan principalmente en las áreas de paneles fotovoltaicos. Aun así, el problema de la intermitencia en la generación de electricidad sigue presente.

Una posible solución para las instalaciones PV consiste en la introducción de un set de baterías que sea capaz de administrar la energía necesaria en los momentos de poca irradiación solar como consecuencia del paso de nubes. El principio básico de operación será el siguiente: las baterías se cargan cuando la irradiación solar es tal que el panel se encuentra suministrando más energía de la que la red se encuentra demandando y se descargan cuando una nube impida al panel recibir la irradiación necesaria para cubrir la demanda de energía de la red. Por lo tanto, aunque la irradiación solar recibida por el panel fotovoltaico no sea constante por la aparición de nubes, las baterías serán capaces de suministrar la energía necesaria para que el flujo de la misma que llega a la red sea más constante y, como consecuencia, se reduzca el estrés y las necesidades de regulación de la red.

Con el objetivo de conectar una batería al resto de la planta PV, se va a diseñar un convertidor bidireccional DC-DC con topología “Dual Active Bridge”(en adelante DAB). Se trata de un diseño a escala para su construcción en un laboratorio de baja potencia por motivos de seguridad, por lo que el principal objetivo del diseño es demostrar la validez de los conceptos y corroborar el futuro funcionamiento del convertidor en caso de ser aplicado en una planta fotovoltaica. El convertidor DAB presenta una topología perfecta para esta aplicación puesto que no sólo puede emplearse en plantas PV, sino también en diseños PV residenciales. Esto se debe a la característica única que presentan estos convertidores de permitir el flujo bidireccional de energía, proveyendo de aislamiento galvánico al sistema y alcanzando altos valores de eficiencia con alta densidad de energía.

Estado del arte

Antecedentes de la tecnología fotovoltaica

La demanda de fuentes de energía con menor impacto medioambiental y coste del combustible prácticamente nulo se encuentra en alza. Por esta razón, la energía fotovoltaica se propone como alternativa para la satisfacción de dicha necesidad. Materiales como el Silicio (Si) se están estudiando para el desarrollo de células fotovoltaicas más eficientes. Sin embargo, este incremento en el uso de la tecnología fotovoltaica se ve contrarrestado por la incapacidad de provisión de energía cuando la irradiación solar recibida por las células fotovoltaicas se ve reducida a lo largo del día o es totalmente inexistente- durante la noche. De hecho, cuando las células reciben una irradiación reducida debido a la aparición de sombras- normalmente debidas al paso de las nubes-, el Maximum Power Point Tracker (MPPT) presenta diferentes picos y su control se convierte en una tarea mucho más ardua.

Estos retos suscitan un mayor interés si se destaca que el futuro de la producción de la electricidad estará muy probablemente influenciado por la generación distribuida. Es en este contexto donde el desarrollo de la energía fotovoltaica recoge sus frutos. De hecho, la implementación de baterías en la red cambiará la forma de regulación de la misma: generación y demanda deberán permanecer iguales pero las baterías podrán tomar parte en cualquiera de las dos según la situación que se de. Por esta razón, las empresas distribuidoras buscan la maximización de los beneficios que en un futuro serán capaces de extraer de la producción y distribución de energía en lo que es comúnmente conocido como Smart Grids (Redes inteligentes).

Convertidores DC-DC

Introducción

Históricamente, la conexión de los convertidores DC-DC ha sido unidireccional y su diseño se ha realizado para frecuencias de línea a 50 o 60 Hz. Si avanzamos algunos años, varias tipologías, algunas de ellas de altas frecuencias, comienzan a aparecer, así como algunas presentan transformadores provistos de aislamiento galvánico. Por lo tanto, resulta útil distinguir entre convertidores DC-DC aislados y no aislados.

No aislados

Normalmente, los convertidores más básicos como el buck o boost no están aislados galvánicamente y no presentan la capacidad de flujo bidireccional de potencia. Sin embargo, con pequeñas modificaciones y empleando conmutadores como MOSFETs en lugar de diodos, la potencia es capaz de circular en ambas direcciones. Uno de los mayores inconvenientes resulta ser que sin el transformador y su relación de transformación, estas topologías no resultan prácticas para diferencias grandes de voltaje.

Otros tipos de tipologías se han estudiado como resultado de la interconexión de las dos topologías simples. Esto resultó en un convertidor buck-boost y el Cuk, que pueden también emplearse en sus variantes bidireccionales.

Aislados

Entre los convertidores DC-DC aislados más populares se encuentran las topologías Flyback y Forward, que destacan por ser las más simples. Disponiendo únicamente de un interruptor y un diodo en su versión unidireccional y más básica, estos convertidores son derivados de los convertidores buck-boost y buck respectivamente. Algunas de las ventajas que presentan estas topologías son, en primer lugar y principalmente la simplicidad del circuito con un número muy reducido de elementos. Esto implica que su diseño resulta sencillo y presentan un control más simple ya que hay menos elementos que formen parte de la conversión al tener aislamiento galvánico.

Estos conversores han ido evolucionando hacia otros con características más adecuadas para las aplicaciones en las que se emplean según han pasado los años. El flujo bidireccional de potencia es la característica principal que se desea conseguir con estos conversores.

De los conversores descritos previamente, la topología escogida para el diseño de este proyecto será DAB. Algunas de las razones para esta elección incluyen las siguientes:

- Flujo bidireccional de potencia
- Aislamiento galvánico
- Alta eficiencia
- Alta densidad de energía

Estas características son esenciales para la aplicación deseada: la implementación de baterías en sistemas PV. El flujo bidireccional de potencia puede ser una de las características más importantes, ya que la batería debe ser capaz de cargarse y descargarse para su correcto funcionamiento.

Soft-switching fundamentals

Las técnicas de Soft-switching siguen en proceso de investigación, ya que tienen como principal objetivo la minimización de las pérdidas debidas a la conmutación de los elementos que cambian de estado interrumpiendo la corriente y el voltaje. Las pérdidas de conmutación han sido tradicionalmente un factor limitante en el diseño de convertidores, ya que la evolución hacia frecuencias de conmutación mayores presenta ventajas que convierten los diseños en proyectos muy atractivos. Una característica importante, si no la más, de emplear frecuencias altas es la reducción del tamaño del conversor y de sus componentes. Sin embargo, debe buscarse un equilibrio entre tamaño y eficiencia, puesto que a mayores frecuencias, las pérdidas por conmutación aumenta.

Existe además otra desventaja: las corrientes y voltajes que deben soportar los elementos son mayores, siendo por tanto las pérdidas por conducción mayores.

Topología Dual Active Bridge (DAB)

La topología básica de un DAB se define a partir de dos puentes activos conectados a través de un transformador diseñado para altas frecuencias. Dado que los dos puentes son puentes activos, como su nombre indica, el flujo de potencia puede ser bidireccional. Aunque hay múltiples variaciones para la inversión y rectificación de ondas como las técnicas de Pulse Width Modulation (PWM), la forma más simple es la de generación de ondas cuadradas con un duty cycle del 50%. De esta forma no se añade complejidad al control de la conmutación de los MOSFET. Además, siendo la razón por la que este proyecto consiste en una simulación para la demostración de la validez de los conceptos para sistemas mayores de alta potencia, se ha descubierto que las ondas cuadradas son más adecuadas para altas potencias. La imposición de ondas cuadradas en los devanados de los transformadores permite que el tiempo “muerto” o fases en las que la tensión es nula se reducen y, aunque esto puede reducir las pérdidas de conmutación, reduce también la cantidad total de potencia transmitida. Además, la implementación de técnicas de Zero Voltage Switching (ZVS) reducen este problema.

Al mismo tiempo, el control de potencia se mantiene esencialmente igual: el ángulo relativo entre las dos ondas cuadradas define la cantidad y dirección del flujo de energía en cada instante de tiempo. En este proceso, la inductancia de dispersión de flujo (y cualquier otra inductancia colocada en serie con la misma) del transformador juega un papel importante ya que es la encargada de la transmisión de potencia. Es importante destacar que la potencia fluye desde las ondas cuadradas adelantadas a las retrasadas, esto es, si la potencia necesita fluir de un nivel bajo de tensión (la batería) a otro de tensión mayor (el bus DC), la onda cuadrada obtenida en el lado de baja tensión deberá estar adelantada con respecto al lado de alta del puente. La topología del circuito puede observarse en la Figura 3.8.

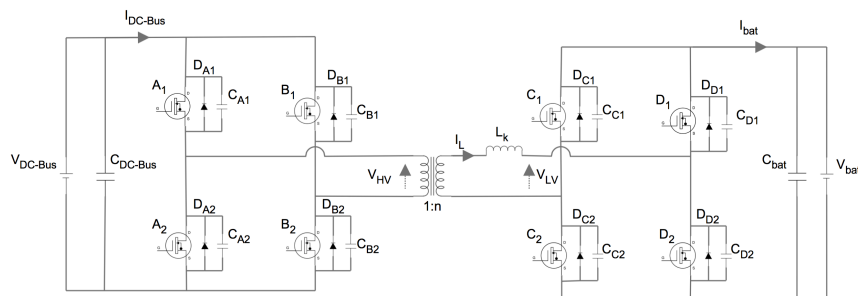


Figure 0.1.: Topología Dual Active Bridge

Motivación

La razón principal para el desarrollo de este proyecto es la profundización en el mundo de la electrónica de potencia no solo con la implementación de una topología extremadamente avanzada para principiantes, sino además para el diseño de un proyecto de principio a fin. Además, como el soft switching es una técnica todavía en proceso de investigación, su empleo y por lo tanto la atención prestada sobre la misma resulta importante.

Finalmente, al ser el principal objetivo el diseño de este convertidor en específico para aplicaciones fotovoltaicas, resulta entrelazado de forma natural con la implementación de baterías. Por lo tanto, y debido a la existencia de baterías en diferentes áreas como pueden ser aplicaciones aeroespaciales, en el interior de vehículos eléctricos o incluso en la carga y descarga de las baterías de los mismos, el estudio de la topología del convertidor resulta muy interesante. Debido a la demanda de baterías con mayor densidad de energía y mayor potencia, un conversor DAB sería ideal debido a las características especiales que presenta y que se han enunciado con anterioridad.

Objetivos

Los principales objetivos de este proyecto incluyen:

- La elección de la topología.
- El estudio de la topología DAB: No solo sus ventajas y desventajas, sino sus modos de operación y las ecuaciones que definen su comportamiento.
- El diseño de todos los componentes para los parámetros específicos del proyecto: Esto incluye, entre otros, el dimensionamiento de los condensadores e inductancias y el cálculo de los límites de potencia.
- Los resultados de la simulación y su análisis.

Sin embargo, más allá de las competencias de este proyecto, pero no menos interesantes, se encuentran los siguientes:

- El control del sistema.
- La regulación de los casos con condiciones de baja carga.
- La construcción del convertidor.

Abstract

Summary

This thesis explores the dual active bridge (DAB) topology and implements it on a bi-directional DC-DC converter. The application it is intended for is the connection in parallel of an energy storage device, such as a battery, with photovoltaic (PV) cells, and then in series to the grid through another converter, but this time a DC-AC.

The DAB topology will be implemented in the interconnection of a battery to the high voltage DC bus, with the intention of, through a control scheme that may be designed afterwards, supplying the correct amount of energy when the PV cells cannot provide a steady power output, like for example in the case of cloud appearance and, therefore, a decrease in the insolation levels. A detailed study of the topology was made in order to choose the components for the converter. In addition, zero voltage switching (ZVS) techniques have been implemented to improve the efficiency.

The project was designed as a proof of concept, thereby limiting the power output to safe values for the laboratory. For that reason, some issues have arisen in terms of the components choices, leaving us with a transformer that is hard to build for the desired specifications. Hence, future work includes the design of the control scheme as well as the realization of the transformer for a later construction.

Subject Keywords: Dual active bridge; DAB; photovoltaic; bi-directional; soft-switching; ZVS; DC-DC.

Introduction

Nowadays, the demand of lower environmental impact energy is on the rise. The idea behind the exploitation of natural resources to obtain energy while contaminating less has always been on the table, but there was always a problem: economic viability. In fact, this factor alone has made renewable energy adoption in the world a slow process.

While there are different approaches to this problem and various nations are betting on different types of energy, wind and photovoltaic (PV) plants are becoming increasingly popular. In the case of PV installations, new technology is being developed, especially in the area of photovoltaic panels. However, the problem of the intermittency of power generation still exists.

A solution for PV installations would be to introduce a set of batteries that would be able to supply the energy needed when shading became present. The basic principle of operation is the following: batteries will be charged when the solar panel is able to supply an excess power, but the grid does not have the demand for it, and will be discharged when a cloud appears, as the panel is not able to produce the same amount of energy with the sunlight available at that moment. Therefore, if insolation is not constant due to the appearance of a cloud, the batteries can supply the energy needed to provide a more constant power flow and therefore, reduce the stress and regulation requirements of the grid.

In order to connect the battery to the rest of the plant, a bidirectional DC-DC converter will be designed with a "Dual Active Bridge" (DAB) topology. As this design is to be simulated in a laboratory of low power -due to safety requirements- the design will be proof-of-concept to demonstrate the capability of the idea in a real photovoltaic plant. The DAB converter is a perfect topology for this application because not only can it be used in big solar plants, but also in residential PV designs. This is due to its unique characteristics of enabling bidirectional power flow, providing galvanic isolation and achieving high efficiency with a high energy density.

Literature Review

Photovoltaic background

Demand for renewable energy with low environmental impact and virtually zero fuel cost has been on the rise. For that reason, photovoltaic energy has been proposed as a solution to this problem. Different materials are being studied at the moment, such as Silicon, in order to create more efficient photovoltaic cells. However, this increase in performance is offset by the inability to provision energy during nighttime or in when the cell receives low levels of sunlight in instances like in partially shaded conditions. In fact, when the cells have their received insolation reduced by the effect of a shade -usually as a result of clouds- the Maximum Power Point Trackers (MPPT) have different peaks and their controls becomes a much more complex task.

These challenges become more apparent when the most likely future of electric production is heavily influenced by distributed generation. In this context, is where photovoltaic energy harvesting shines. Moreover, battery implementation in the grid will change the way regulation works: supply and demand must remain equal, but batteries can take part in either of them depending on the situation. For this reason, utilities are trying to maximize the benefits they will be able to obtain from the future of energy production and distribution through what is commonly known as Smart Grids.

A major disadvantage of PV energy harvesting is its power fluctuation, which can contribute to grid stability issues, specifically if we are referring to voltage and frequency stability. In order to solve this problem, energy storage systems have been used to design new and modified power plants with the aim to provide a less fluctuating power output.

DC-DC Converters

Introduction

Historically, the connection of DC-DC converters has been unidirectional and have been designed at line frequencies at 50 Hz or 60 Hz. If we forward some years other topologies, various at higher frequencies, start appearing, a few of them including transformers that provided galvanic isolation. Therefore, it is useful to distinguish between isolated and non-isolated DC-DC converters.

Non-Isolated:

Usually, the most basic converters such as the buck or the boost are not isolated and do not have bidirectional power flow capabilities, however, with slight modification and by using switches such as MOSFETs instead of diodes, the power is enabled to circulate in both directions. One of its most important drawbacks is that, without the transformer and its turns ratio, for big voltage differences this topologies becomes impractical.

As a next step, some other topologies were studied as a result of the interconnection of both of the simple topologies. This ended un in a buck-boost and in the Cuk, which can also be used in their bidirectional variants.

Isolated:

Among some of the more popular isolated DC-DC converters the Flyback and the Forward topologies are the most simple ones. By only having one switch and one diode in its unidirectional and most basic version, they are derived from the buck-boost and the buck converter respectively. Some of the advantages these topologies have are, first and foremost the simplicity of the circuit with a very low part count. This means that they are designed easily and it is simpler to control them as less elements take part in the conversion while having galvanic isolation.

These converters have been evolving to ones with more suitable characteristics to the applications they are catered to as time advanced. Bidirectional power flow is one of the key features several of these converters look for.

For the previously described below, the chosen topology for the design and simulation this paper is going to focus on will be the Dual Active Bridge converter (DAB). Some of the reasons for the selection of this type of converter include the following:

- Bidirectional power flow
- Galvanic isolation
- High efficiency
- High energy density

These characteristics are essential for the desired application: the implementation of batteries in PV systems. Bidirectional power flow may be one the most important one, as the battery should be able to charge and discharge in order to function correctly.

Soft-switching fundamentals

Soft switching techniques are still being researched as their main objective is to minimize the losses related to the switching of devices that change states interrupting current and voltage. Commutation losses have traditionally been a limiting factor for the design of converters, as usually going to higher switching frequencies have advantages that make them very attractive. One, if not the most, important characteristic of going to higher frequencies is the reduction of size of the converter and its individual parts. However, there has to be a trade off between size and efficiency, as higher switching frequency means that commutation losses rapidly increase.

This also comes with a sacrifice: the currents and voltages that the elements sustain are higher, making conduction losses also higher.

Dual Active Bridge Topology (DAB)

The basic topology of a Dual Active Bridge (DAB) topology is defined by two active bridges linked through a transformer that is usually fit for high frequencies. As both of the bridges are active, hence the name, the power-flow can be bidirectional. Although there are multiple variation in the way of wave inversion and rectification such as Pulse Width Modulation (PWM) techniques, the simpler version of generating square waves with a 50 percent duty cycle limits the added complexity of MOSFET commutation control. In addition, and due to the reason that this project is a proof of concept for a big, high power facility, it has been discovered that square waves are more beneficial in case of high power applications. By imposing square waves on the transformers windings, the "dead" time or phases where the voltage is zero is reduced, and although this can reduce switching losses, it also reduces the total amount of power transmitted. Moreover, since ZVS techniques will be implemented, this is not a big problem.

At the same time, the control of the power remains essentially the same: the relative angle between both of the square waves define the amount and the direction of the energy flow at that instant of time. In this process, the leakage inductance (and any other inductances placed in series with it) of the transformer plays an important role, as it is in charge of transmitting the power. It is worth noting that the power flows from the leading to the lagging square wave, that is, if power needs to flow from the low voltage (the battery) to the high voltage side (the DC bus), the square wave created on the LV side should be leading the one from the HV bridge. The circuit topology can be seen in Figure 3.8.

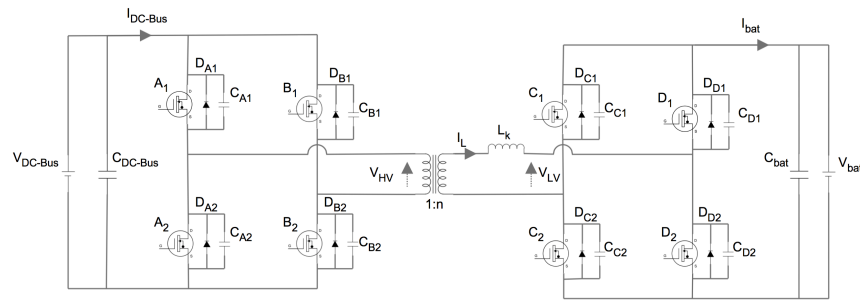


Figure 0.2.: Dual Active Bridge topology

Motivation

The main reason for this project is to get more in depth into the world of power electronic not only through an extremely advanced topology for beginners, but to design a project from beginning to end, including its simulation and the problems that may occur in the process.

On top of that, since soft switching is an area still being researched, providing some insight to the process and bringing it to attention is also important.

Finally, while the main focus of this specific converter design is in its photovoltaic application, it is almost naturally intertwined with battery implementation. Therefore, and due to the existence of batteries in several areas such as aerospace applications, in the inside of electric vehicles, or even being involved in the charging and discharging of the batteries of these electric cars, the converter topology is very interesting to study. Due to the demand for more energy dense batteries and more powerful batteries, a Dual Active bridge converter would be ideal for the special characteristics listed above.

Objectives

The main **objectives of the project include:**

- The choice of the topology.
- The study of the Dual Active Bridge topology: Not only its advantages and drawbacks, but more importantly, its modes of operation and the equations that shape its behavior.
- The design of all the components for the specific parameters of the project: This includes but is not limited to the sizing of capacitors and inductances and the calculation of switch stress.
- The simulation of the results.

However, **outside of the scope of the project** are the following:

- The control of the system.
- The regulation for cases in low load conditions.
- The construction of the converter.

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1. Introduction

Nowadays, the demand for lower environmental impact energy is on the rise. The idea behind the exploitation of natural resources to obtain energy while polluting less has always been on the table, but there was always a problem: economic viability. In fact, this factor alone has made renewable energy adoption in the world a slow process. Nevertheless, times and technology are changing; generating and consuming renewable energy has become an option. Looking at energy consumption and sustainability from a future-oriented perspective has led us to realize that, with this rate of consumption of fossil fuels, in a few years their reserves will be near depletion. This fact could lead to economic uncertainty, due to the forecast of price increases, and as the reserves of this type of energy start to diminish, energy will have to be extracted from different sources. This change of traditional energy sources to new ones has to be gradual for a sustainable and seamless transition to occur. Different nations are starting to incorporate more renewable sources into their energy mix in order to meet the deadlines they are self-imposing not only to contaminate less but also to save money in the long run, as these renewable sources' fuels are free.

While there are different approaches to this problem and various nations are betting on different types of energy, wind and photovoltaic (PV) plants are becoming increasingly popular. In the case of PV installations, new technology is being developed, especially in the area of photovoltaic panels. However, the problem of the intermittency of power generation still exists. Not only do these types of electric plants stop generating power at nighttime, but also they are not able to provide a constant power output, leading to the possibility of not satisfying the power required by the grid at that time. Some of the major issues concerning this problem are the effects of temperature and shading, which make the output power of PV installations nonlinear.

A prime example of shading is the appearance of clouds, blocking the sunlight received by the panel and therefore reducing the insolation it receives. This creates a noticeable drop in the power the panel is able to output, which in turn means that the grid has to supply the remaining energy found elsewhere. On top of that, this may lead to frequency changes in the grid which have to be regulated in order to keep the system stable.

A solution for PV installations would be to introduce a set of batteries that would be able to supply the energy needed when shading became present. The basic principle of operation is the following: batteries will be charged when the solar panel is able to supply excess power, but the grid does not have the demand for it, and will be discharged when a cloud appears, as the panel is not able to produce the same amount of energy with the sunlight available at that moment. Therefore, if insolation is not constant due to the appearance of a cloud, the batteries can supply the energy needed to provide a more constant power flow, thereby reducing the stress and regulation requirements of the grid.

In order to connect the battery to the rest of the plant, a bi-directional DC-DC converter has been designed with a “dual active bridge” (DAB) topology. As this design is to be constructed in a laboratory of low power —due to safety requirements— the design will be a proof-of-concept to demonstrate the capability of the idea in a real photovoltaic plant. The DAB converter is a perfect topology for this application because not only can it be used in big solar plants, but also in residential PV designs. This is due to its unique characteristics of enabling bi-directional power flow, providing galvanic isolation and achieving high efficiency with a high energy density. Figure 1.1 shows the main idea of the project in a graphical way that is easier to understand.

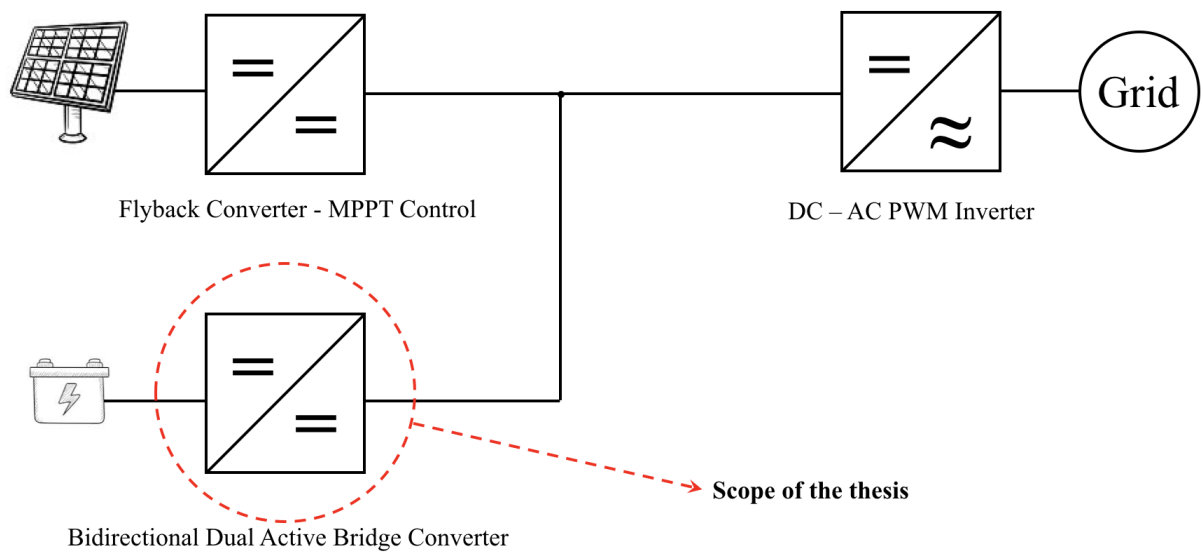


Figure 1.1.: Photovoltaic facility and the scope of the thesis

2. Applications

At this point in the thesis, it is useful to explain some of the applications in which this type of converter can be implemented. Obviously, as technology gets more and more advanced, the requirements for the components become stricter while always striving for more efficiency. In this chapter, we will study three applications where this converter could be implemented.

2.1. Intended application: Photovoltaic facility

The design of this converter was initially conceived for a photovoltaic facility and, specifically, one where there is energy storage in parallel to the main power deliverer, that is, the photovoltaic (PV) cells. The main idea is to reduce the variability of the power output from the PV plant in order to facilitate the regulation of the grid, because, as common sense and reality show us, the renewable sources are the first ones to enter the energy mix, as their fuel is free. Including before, for example, a coal plant not only would be more expensive in terms of fuel but also much more polluting.

Usually, insolation is one of the determining factors that fix the power output from the PV cells, that is why any change on its levels, like for example due to the appearance of a cloud, can result in a change of the power output. This drop in power means that the grid is suffering as the electrical market is one of the few where supply and demand must be exactly equal at all times. If part of the supply disappears for a small amount of time, other generators must provide that power.

For these reasons implementing batteries for energy storage to supply the needed power to obtain a more constant output is a good solution. The battery will need to be connected to a DC-Bus in parallel to the PV cells and then both of them will be connected to the grid through a DC-AC inverter. This inverter functions with pulse width modulation (PWM) techniques to convert the DC to AC and then filters the output to obtain a nice sinusoid ready to serve as input for the grid.

To maintain the voltage of the DC-Bus the battery will be implemented and its interconnection will be through a bi-directional DC-DC converter. Finally, the PV modules should also be connected to the DC-Bus through a DC-DC converter. Sometimes, this converter implements maximum power point tracking (MPPT) controls to maximize the power output of the PV cells.

Our chosen topology for the design of the battery to DC-Bus converter, the dual active bridge (DAB), fits this application perfectly as some of its key features match perfectly the requirements of the application. One of them is galvanic isolation needed for safety requirements and isolating the batteries from both the cells and the grid, meaning that in case of malfunctioning of the battery, current surges, or other problems, they will remain in the energy storage side, separating the rest of the circuit and confining the issue. Another essential feature of the topology is its bi-directionality, as the battery should provide power when it is needed, but also needs to recharge to be able to provide this power in another moment in the future. Of course, its high efficiency will not hinder a type of energy exploitation that has inefficient components such as the PV cells that have efficiencies of around 5 % to 45 % as it can be seen in Figure 2.1 provided by [1]. Finally, its high energy density means that not a lot of extra space will be needed in the facility as there is already a great amount of land being utilized.

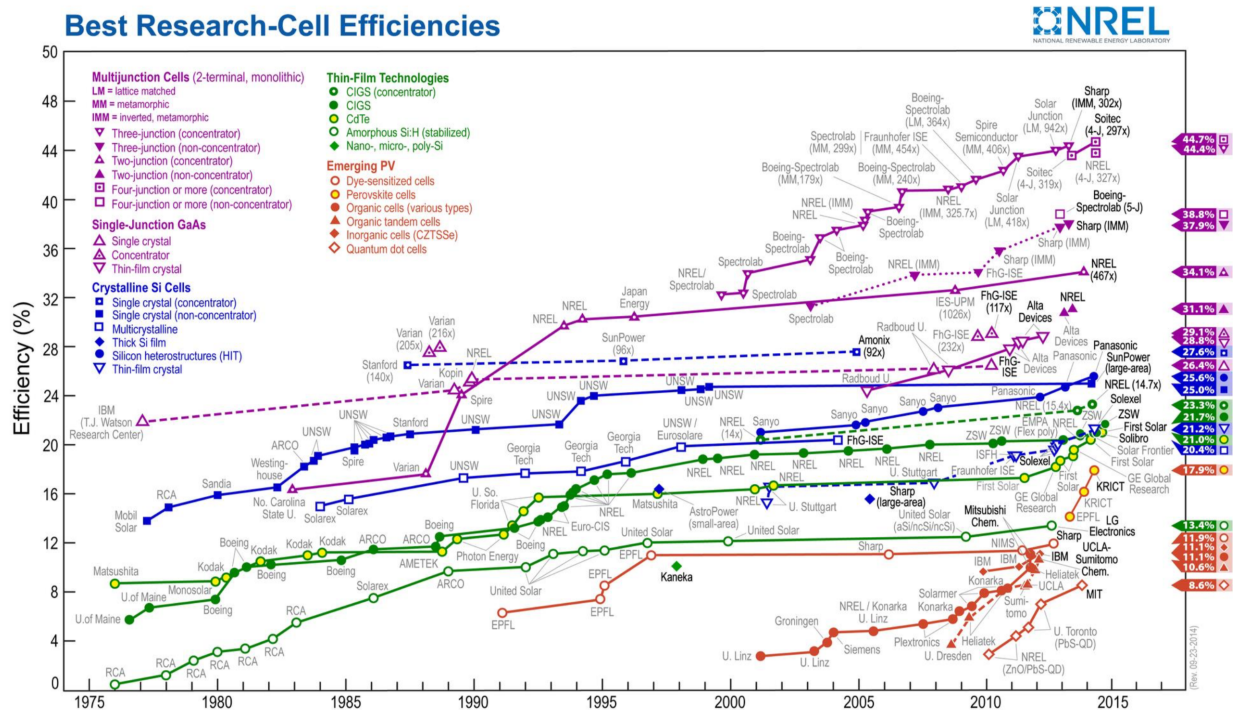


Figure 2.1.: Efficiency of the current PV cells and their materials

2.2. Other applications

Although the converter has been designed with a PV plant in mind, it is worth noting that this type of DC-DC converter is also of great usefulness in other applications, especially when implementing the DAB topology. Two of them will be explored in this section and some insight will be provided to understand the benefits of this converter in other applications.

2.2.1. V2G: Vehicle to grid

As the electric vehicle (EV) develops and it spreads to more households, so do the batteries that are inside these vehicles. There are some ideas concerning how the energy of these batteries can be utilized, but one of them stands out. This idea is that when one arrives home, if you decide to connect the EV to the grid, the energy stored in this battery can be utilized by the appliances in your home and reduce the consumption cost.

When people arrive home from work is normally when the domestic demand for electricity is higher as people use what consumes more electricity of the entire house, the stove and the oven. What is more, it is often used along with extra heating or cooling and the addition of entertainment systems such as televisions or music devices. All of this contributes to a big peak in the demand at that times of the day, and a big valley when it is deep at night when everyone is sleeping. By implementing the discharge of the EV's battery in those hours, every household would use some help and in that way would decrease the cost of electricity (because as there is more demand, the prices of the electricity rise) and the grid regulator can have a more predictable and more flat load curve; however, as the new load curves are unpredictable research is still ongoing [2]. Figure 2.2 portrays this fact, especially in the coldest months.

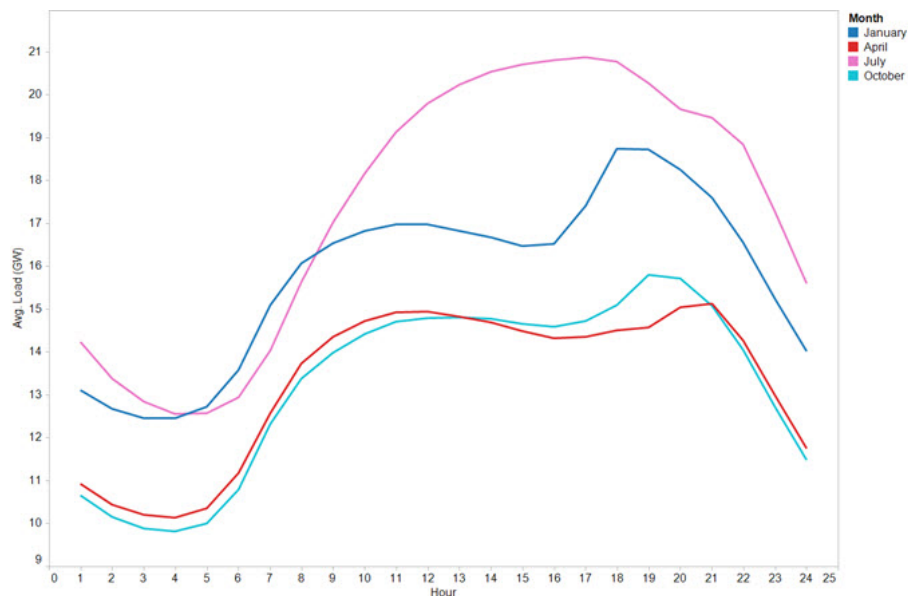


Figure 2.2.: Electrical daily demand for different months of the year. (Source: New England ISO)

In order to charge and discharge the battery a converter is needed and galvanic isolation is imperative, again for safety measures as, otherwise, any problem could result in fatalities. On top of that, the bi-directional nature of the converter is also imperative as it has to be charged and discharged. Finally, high energy efficiency means that less cost will be incurred. Generally, the grid comes in AC, but this DAB topology may be slightly adapted to fit those needs.

2.2.2. Aerospace

The last application we will discuss is aerospace. The aerospace field as a demanding one and for that reason weight and energy density are of enormous importance. In fact, this is one of the key features that the DAB topology can provide when implementing a DC-DC converter between the auxiliary circuit of the plane and the main electric circuit in charge of actuators and avionics amongst others.

In addition, when transmitting the electricity, DC has an edge over AC as losses are lower, especially in HV. This is due to the fact that in DC current the losses associated with any of the reactive elements such as cable inductances and capacitances are non-existent.

Again the galvanic isolation is of great importance due to safety reasons to limit the possibility of fatality of any of the passengers or crew from a potential electrical failure. Bi-directionality may come in handy when battery storage comes into play in this field.

3. Literature Review and Motivation

3.1. Photovoltaic background

Demand for renewable energy with low environmental impact and virtually zero fuel cost has been on the rise. For that reason, photovoltaic energy has been proposed as a solution to this problem, as the equipment needed for the exploitation of the free and non-depletable solar energy is rapidly advancing. Different materials and their performance for the construction of photovoltaic cells are being studied at the moment, such as silicon [3] or gallium [4]. However, this increase in performance is offset by the inability to provide energy during nighttime or when the cell receives low levels of sunlight such as in partially shaded conditions. In fact, when the cells have their received insolation reduced by the effect of shade -usually as a result of clouds- the maximum power point trackers (MPPT) have different peaks and their control becomes a much more complex task [5], [6].

These challenges become more apparent when the most likely future of electric production is heavily influenced by distributed generation [7]. In this context, the potential of photovoltaic energy harvesting shines. Currently, rooftop PV installation is one, if not the only, answer to the problem of distributed generation of electricity. From its modular design to the cell size availability, PV generation excels at diversified positioning, highlighting its placement on numerous rooftops and the energy reduction associated with it. If the policy regulating PV rooftop panels allows it, net metering becomes a reality, impacting directly the price the owner pays for electricity. Also, if the power supplied by this installation is enough, net metering can result in economic revenue for the installation owner when specific circumstances take place. However, by increasing distributed generation, and specifically PV energy production, its drawbacks emerge as a problem that must be solved.

A perspective that is not to go unnoticed is the one that Nourai et al. [8] reflect upon, the effect of energy storage systems on utilities. Battery implementation in the grid will change the way regulation works: while supply and demand must remain equal, now part of that demand—or supply—can come from storage solutions located at substations, large-scale generation plants or small production plants at rooftops. Furthermore, utilities are aware of this transition and are trying to maximize the benefits they will be able to obtain from the future of energy production and distribution through what is commonly known as “Smart Grids” [9]. Energy storage installation plays a big role in this future and these type of companies are making efforts to reap the benefits [10].

A major disadvantage of PV energy harvesting is its power fluctuation, which can contribute to grid stability issues, specifically if we are referring to voltage and frequency stability. In order to solve this problem, energy storage systems have been used to design new and modified power

plants with the aim to provide a less fluctuating power output. In [11] one can see the promising effects of battery energy storage systems (BESS) and their impact on how the grid views a PV facility with BESS implemented. One of the graphs from [11] has been borrowed as a reference and can be viewed in Figure 3.1.

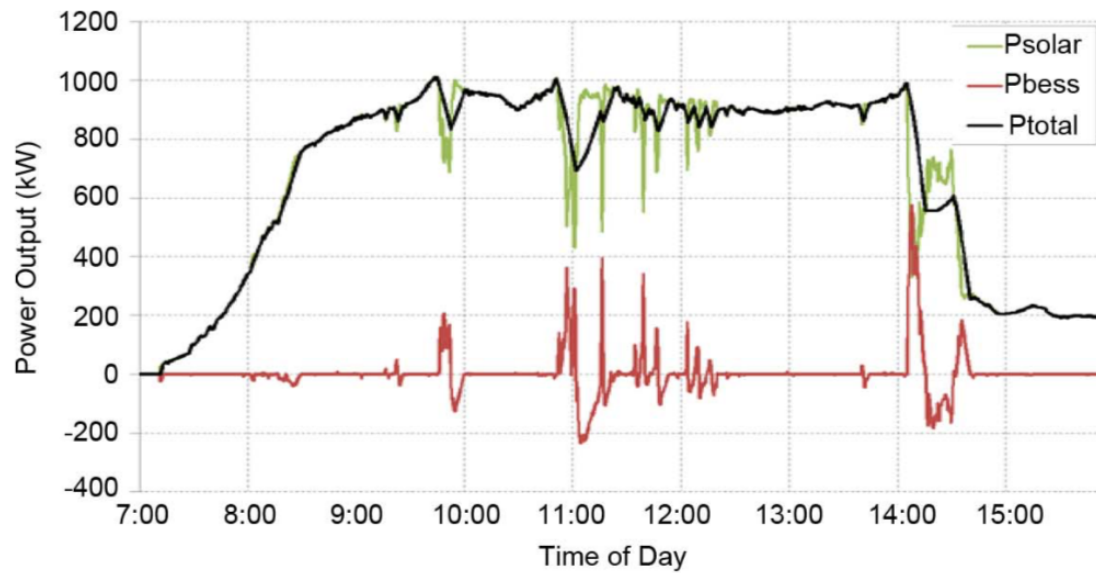


Figure 3.1.: Effect of energy storage (BESS) in PV plants

3.2. DC-DC Converters

Historically, the connection of DC-DC converters has been unidirectional and designed at line frequencies at 50 Hz or 60 Hz. If we forward some years, other topologies, various at higher frequencies, start appearing. A few of them include transformers that provided galvanic isolation. Therefore, it is useful to distinguish, as Karshenas et al. [12] do, between isolated and non-isolated DC-DC converters.

The basic structure of a DC-DC isolated converter is based on the power-flow through a transformer from or to a DC bus to another source of DC Voltage, typically a battery. To transform the DC voltage from one side of the transformer to the other, the voltage has to be inverted usually in the forms of square waves to be able to go through the transformer. The following step is to rectify the square wave on the other side in order to convert it back to the desired DC voltage, previously filtered by a low-pass filter. A visual description of the process is in Figure 3.2.

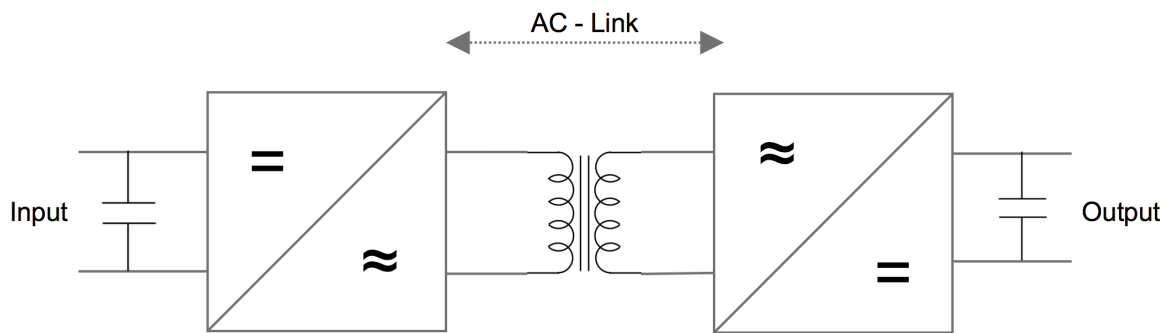


Figure 3.2.: Typical structure of an isolated DC-DC converter

3.2.1. Non-Isolated

Usually, the most basic converters such as the buck or the boost are not isolated and do not have bi-directional power flow capabilities; however, with slight modification and by using switches such as MOSFETs instead of diodes, the power is enabled to circulate in both directions. One of its most important drawbacks is that, without the transformer and its turns ratio, for big voltage differences these topologies becomes impractical. As an example Figure 3.3 represents a buck converter.

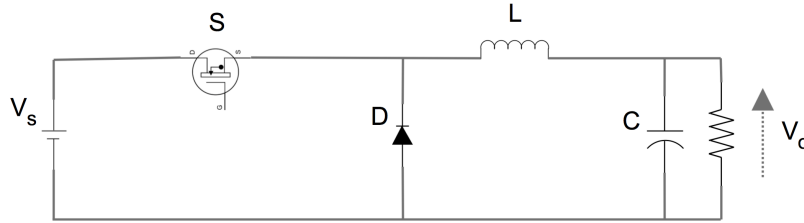
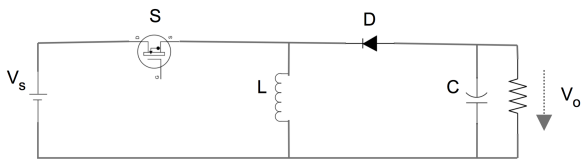
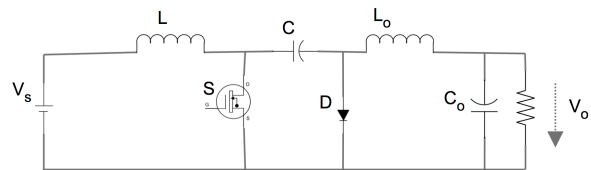


Figure 3.3.: Buck converter topology

As a next step, some other topologies were studied as a result of the interconnection of both of the simple topologies. This ended up in a buck-boost and in the Cuk, which can also be used in their bi-directional variants.



(a) Buck-Boost converter topology



(b) Bidirectional Cuk Converter

Figure 3.4.: Different converter topologies

3.2.2. Isolated

Among some of the more popular isolated DC-DC converters, the flyback and the forward topologies are the simplest. By only having one switch and one diode in the unidirectional and most basic version, they are derived from the buck-boost and the buck converter respectively. Among the advantages of these topologies is, first and foremost the simplicity of the circuit with a very low part count. This means that they are designed easily and it is simpler to control them as fewer elements take part in the conversion while having galvanic isolation.

These converters have been evolving to ones with more suitable characteristics to the applications they are catered to as time advanced. Bi-directional power flow is one of the key features several of these converters look for. From two switches with dual flybacks, or zeta-sepics, to 8 switches with the DAB, going through every number of switches in between, it has been noticed that the power capacity available for transfer is proportional to the number of switches[13]. For this reason, as well as the others that are described below, the chosen topology for the design and construction this thesis is going to focus on will be the dual active bridge converter (DAB). The reasons for the selection of this type of converter include the following:

- Bi-directional power flow
- Galvanic isolation
- High efficiency
- High energy density

These characteristics are essential for the desired application: the implementation of batteries in PV systems. Bi-directional power flow may be the most important one, as the battery should be able to charge and discharge in order to function correctly.

3.3. Soft-switching fundamentals

Efficiency is becoming more important nowadays in every part of the technological sphere. The electrical world is no different and, for that reason, new materials, topologies and loss reduction strategies are being researched and experimented with in order to fulfil the requirements users demand. Most of the knowledge aquired about soft switching comes from “*Fundamentals of Power Electronics*” by Erickson and Maksimovic [14].

Soft switching techniques are still being researched as their main objective is to minimize the losses related to the switching of devices that change states interrupting current and voltage. Commutation losses have traditionally been a limiting factor for the design of converters, as higher switching frequencies have advantages that make them very attractive. One, if not the most, important characteristic of going to higher frequencies is the reduction of the size of the converter and its individual parts. However, there has to be a trade-off between size and efficiency, as higher switching frequency means that commutation losses rapidly increase. Some of the main benefits of including soft switching techniques in the design of a DC-DC converter include:

- Reduction in switching losses in the semiconductors: This is the main objective of the strategy.
- Reduction in size and weight: As losses are lower, the amount of heat dissipated in the devices is also reduced, making them consequently smaller as there is no need for big heat sinks. On top of that, passive elements, that are related to frequency, can be smaller as the total value of their impedance can remain constant by switching at higher frequencies.
- Reduction in electro-magnetic Interference (EMI): At higher frequencies it is easier to filter the frequencies that incur electromagnetic interference.
- Utilization of parasitic elements such as leakage inductances of transformers: At high frequencies, magnetizing inductance results in a very big impedance, making the current through them negligible and making the usually disregarded leakage inductance the one in charge of transferring energy.
- Reduction in cost: For the whole reason that size and weight are reduced, the amount of material utilized also drops. This creates an associated cost shrinkage.

This also comes with a sacrifice: the currents and voltages that the elements sustain are higher, making conduction losses also higher.

The basic idea is that in hard switching scenarios, we usually consider that voltage and current are in steady state and generally speaking are constant at that moment. Then, at commutation, both of them —voltage and current— start increasing and decreasing respectively at turn-off. This results in switching losses. One can see the waveforms in Figure 3.5 for more clarity.

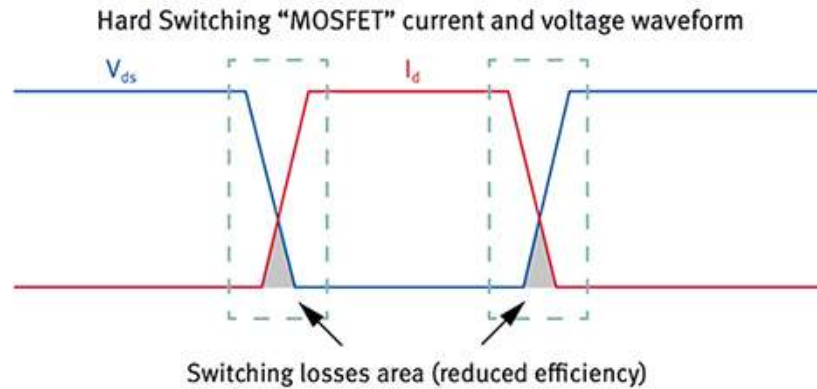


Figure 3.5.: Hard switching waveforms (courtesy of Infineon Technologies)

Therefore, the plan is to have the current or voltage be at a value of 0 before turning on or off the switches. In that way, the power losses will be non-existent - or at least very close to 0-. This results in a clear distinction between ZCS (zero current switching) and ZVS (zero voltage switching).

3.3.1. ZCS: Zero Current Switching

As the name implies, this type of soft switching bases its functioning on bringing the current to 0 before changing the switch state, that is, either from on to off or vice-versa. As one can see in Figure 3.6 provided by [15], this results in an overlap with zero current when the voltage is positive that will incur in switching losses.

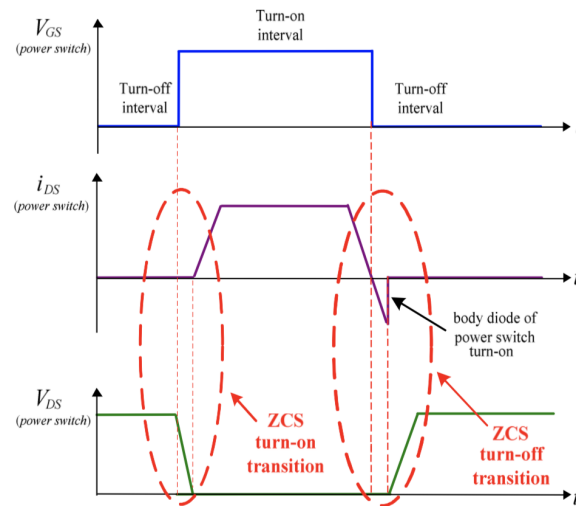


Figure 3.6.: Schematic ZCS waveforms

More often than not, this soft switching strategy is better suited in cases where inductors are already in series with the switching devices. Prime examples of this are filters with an inductor in series with a switching device and some types of resonant converters that use series tanks. However, in the majority of cases, ZVS is preferred.

3.3.2. ZVS: Zero Voltage Switching

Correspondingly, a ZVS strategy indicates that the voltage that the commutating device “sees” is brought to 0 before the turning on and off. In a like manner, Figure 3.7, again provided by [15], portrays the concept graphically. Now, the overlap occurs when the voltage is 0, and thus the switching losses are also reduced to 0 if treating with ideal components.

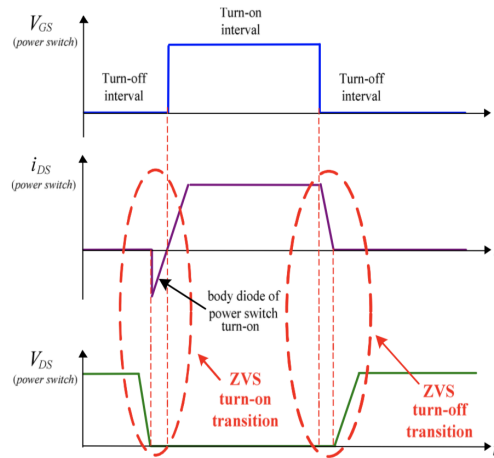


Figure 3.7.: Schematic ZVS waveforms

A zero voltage switching (ZVS) strategy has some key advantages over its counterpart, ZCS. Among them, we can highlight the fact that most of the switching devices, more specifically IGBTs and MOSFETs, have body diodes. This means that in case that the ZVS strategy is obtained, the reverse recovery charge in this body diode will not produce added losses as it will be cleared at 0 V, meaning no power losses.

Additionally, it is also worth mentioning that in case of using MOSFETs, the existence of parasitic capacitances and especially the drain-source capacitance can act as an ideal lossless snubber, clamping the voltage and making a ZVS not only for the MOSFET ideal switch but also for its body diode (as it will be explained later, if the current is negative in the device for that specific moment). This means that ZVS will be obtained without extra losses or significant compromises.

A snubber is an auxiliary circuit designed to provide an escape path to a current that usually is meant to remain constant like in the case of in the existence of an inductor. This may happen when a switch turns off and opens, interrupting the natural path of the current. In order to not cause a huge spike in the voltage, this circuit provides a path for the current to go. Usually, there are RC snubber circuits that may dissipate power, and sometimes with the appearance of a diode.

3.3.3. Dual Active Bridge Topology (DAB)

The basic topology of a dual active bridge (DAB) topology is defined by two active bridges linked through a transformer that is usually fit for high frequencies. As both of the bridges are active, hence the name, the power-flow can be bi-directional. Although there are multiple variations in the way of wave inversion and rectification such as pulse width modulation (PWM) techniques, the simpler version of generating square waves with a 50 percent duty cycle limits the added complexity of MOSFET commutation control. In addition, and due to the fact that this project is a proof of concept for a big, high power facility, it has been discovered that square waves are more beneficial in case of high power applications. By imposing square waves on the transformers windings, the “dead” time or phases where the voltage is zero is reduced, and although this can reduce switching losses, it also reduces the total amount of power transmitted. Moreover, since ZVS techniques will be implemented, this is not a big problem.

Bi-directional capabilities alone would not lead us to the selection of the DAB as our preferred converter, but due to safety measures, galvanic isolation is extremely valuable. By having the battery electrically isolated, if any problem were to occur in the battery, the majority of electricity-related accidents would be minimized, at least concerning human safety. High efficiency is a desired quality for this converter as electricity arriving from the solar panel has to go through the MPPT converter and also through the converter being designed (our DAB). On top of that, when there is a shortage of power, the battery has to liberate the energy previously obtained to the grid going through, again, our DAB converter and the following DC to AC inverter. Finally, the smaller size and weight achieved by this converter make its implementation on rooftops more convenient and, in the case of its utilization in a large PV plant, reduce the impact and the cost of the DC-DC interconnection.

The dual active bridge topology has been heavily researched in various instances [16]. This fact is especially interesting as DAB was created a number of years ago, around 1990 [17]. More recent materials used in switching devices for the most part, such as SiC MOSFETs have permitted the utilization of this topology in several applications. Moreover, zero voltage switching (ZVS) and other soft switching strategies for the reduction of losses [18] make the utilization of this topology viable at high frequencies reaching high efficiencies.

At the same time, the control of the power remains essentially the same: the relative angle between both of the square waves defines the amount and the direction of the energy flow at that instant of time. In this process, the leakage inductance (and any other inductances placed in series with it) of the transformer plays an important role, as it is in charge of transmitting the power. It is worth noting that the power flows from the leading to the lagging square wave that is, if power needs to flow from the low voltage (the battery) to the high voltage side (the DC bus), the square wave created on the LV side should be leading the one from the HV bridge. The circuit topology can be seen in Figure 3.8.

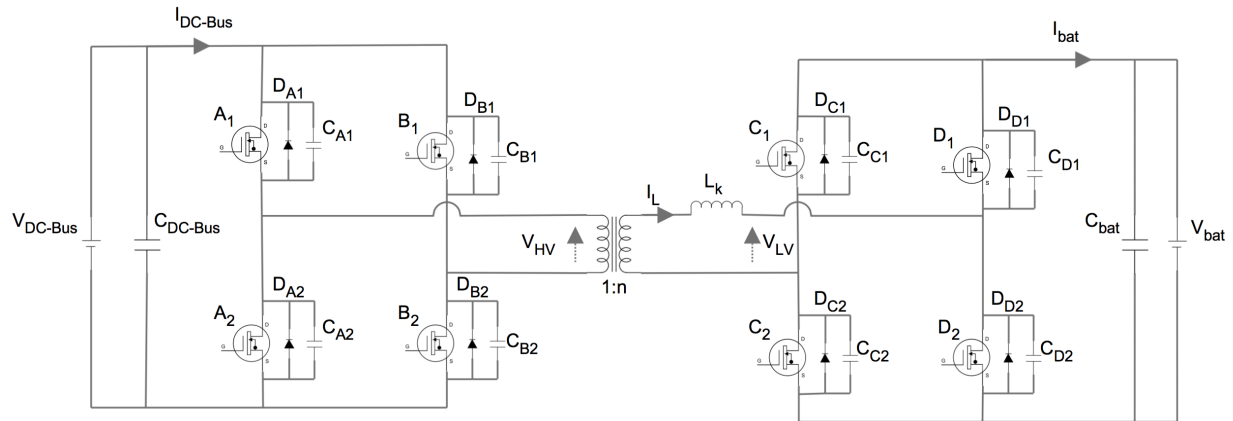


Figure 3.8.: Dual active bridge topology

3.4. Steady state model

The DAB converter consists of two full bridges at both ends of a transformer, one in the HV side—or the DC-Bus—and another in the LV or battery side. These bridges are active, and therefore each is able to create a different square wave, which in this case are high frequency. The two bridges are connected through a transformer and a coupling inductance. This inductance can be the leakage inductance of the transformer, or in case that a bigger value is needed, another one should be put in series with it. The addition of both inductances will be referred as the coupling inductance.

By incorporating the coupling inductance and by appropriately controlling the switching devices, the high frequency square waves are phase shifted, allowing for power flow from one HF bridge to the other. The power flows from the side with the leading square wave to the one which square wave is lagging.

In order to depict the theoretical waveforms, we will assume that $M = \frac{V_{bat}}{nV_{DC-Bus}} < 1$. This will mean that two modes are available. When the power flows from the HV side to the LV, or buck mode, that is when the battery is charging and the opposite situation.

3.4.1. Buck mode

As specified in [16], the main waveforms of the DAB converter in steady state are shown in Figure 3.9 when the battery is charging, or buck mode. Also, in the same article, the various key parameters are defined.

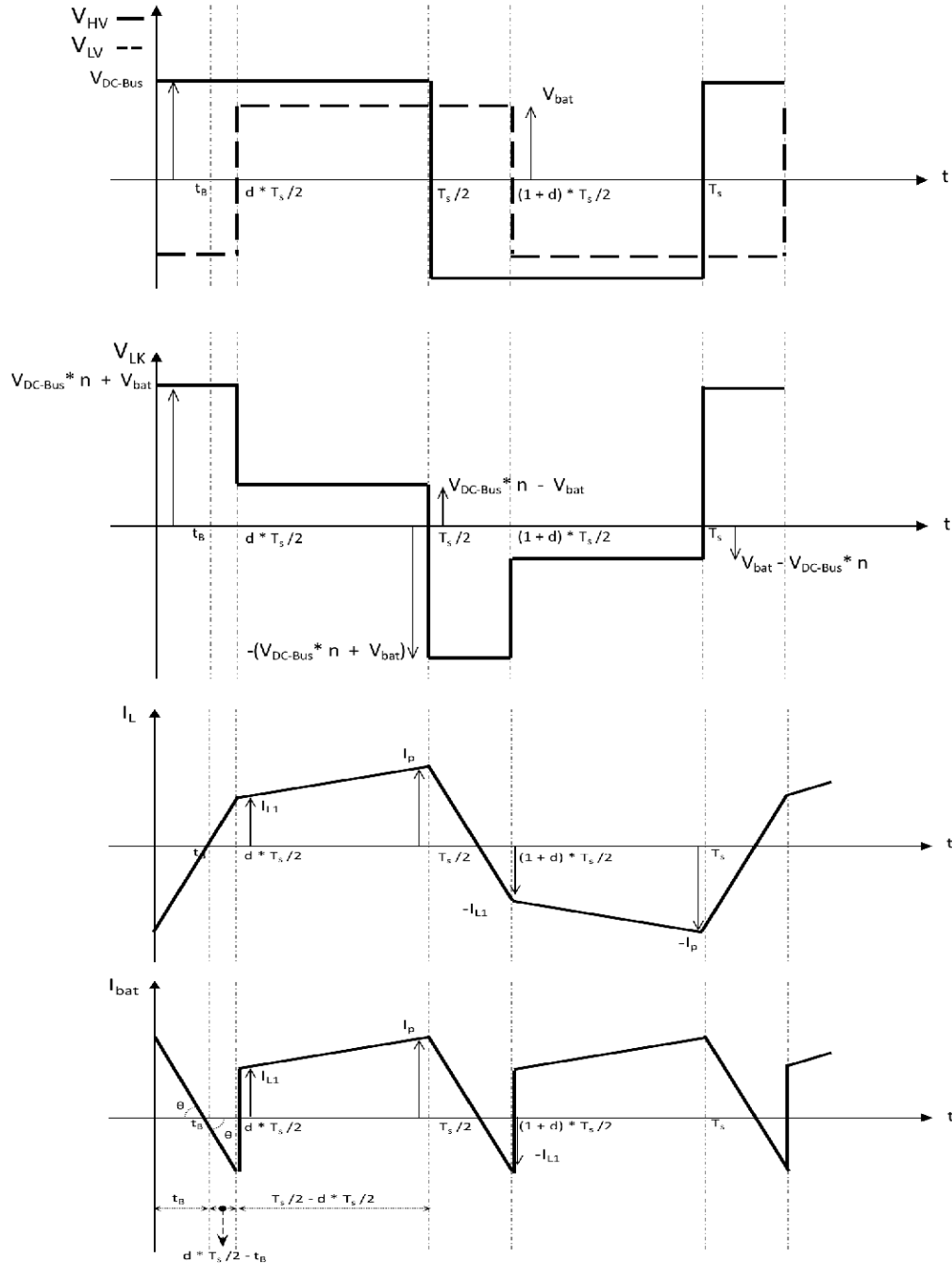


Figure 3.9.: Essential wave forms for the DAB in buck mode

We define the inductor and output power by parameters in the domain of time and current, which also depend on the system parameters such as the DC bus and battery voltages, the coupling or leakage inductance, the switching frequency, the turns ratio, or the phase shift between square waves (d). We should bring to attention the fact that d is the percentage of the semi-period that one wave is phase shifted with respect to the other one. This is due to the reason that the output current (in this case entering the battery) is symmetric around half of the period.

- Current:

$$I_P = \frac{1}{4L * f_s} [nV_{DC-Bus} + V_{bat}(2d - 1)] \quad (3.1)$$

$$I_{L1} = \frac{1}{4L * f_s} [nV_{DC-Bus}(2d - 1) + V_{bat}] \quad (3.2)$$

- Time:

$$t_B = \frac{nV_{DC-Bus} + V_{bat}(2d - 1)}{4 * f_s [nV_{DC-Bus} + V_{bat}]} \quad (3.3)$$

In order to calculate the power delivered by the converter, we need to calculate the average power. It is useful that, as the battery voltage can be considered constant, the output power will only be the product between V_{bat} and the average current output (I_{bat}). This calculation needs the integration of the curve, but, by splitting the areas and some basic geometric relationships, the output current average turns out to be:

$$I_{bat} = \frac{nV_{DC-Bus}}{2L * f_s} (d - d^2) \quad (3.4)$$

And therefore the ideal converter power is:

$$P_{bat} = \frac{nV_{DC-Bus}V_{bat}}{2L * f_s} (d - d^2) \quad (3.5)$$

By normalizing the output power we obtain the following expression that we can plot to locate maximum power point at $d = 0.5$. This makes sense because a phase shift of more than 90° would mean that power would start flowing in the opposite direction again.

$$P'_{bat} = (d - d^2) \quad (3.6)$$

In order to validate the simulations and to make calculations for the power losses in subsequent sections, both the average and RMS (root mean square) values of the currents through the transistors and its body diodes are calculated and included in the tables 3.13.2

Table 3.1.: RMS current values through the devices
when power flows from the HV side to the LV side or buck mode

Device	RMS Current expressions
HV Transistor	$I_{RMSHVTransistor} = \sqrt{\frac{1}{T_s} \left[\frac{I_{L1}^2}{3} * \left(\frac{dT_s}{2} - t_B \right) + \left(\frac{T_s}{2} - \frac{dT_s}{2} \right) \left(I_{L1}^2 + \frac{(I_p - I_{L1})^2}{3} + I_{L1}(I_p - I_{L1}) \right) \right]}$
HV Diode	$I_{RMSHVDiode} = \sqrt{\frac{1}{T_s} \left[\frac{I_p^2}{3} * t_B \right]}$
LV Transistor	$I_{RMSLVTransistor} = \sqrt{\frac{1}{T_s} \left[\frac{I_{L1}^2}{3} * \left(\frac{dT_s}{2} - t_B \right) \right]}$
LV Diode	$I_{RMSLVDiode} = \sqrt{\frac{1}{T_s} \left[\frac{I_p^2}{3} * t_B + \left(\frac{T_s}{2} - \frac{dT_s}{2} \right) \left(I_{L1}^2 + \frac{(I_p - I_{L1})^2}{3} + I_{L1}(I_p - I_{L1}) \right) \right]}$

Table 3.2.: Average current values through the devices
when power flows from the HV side to the LV side or buck mode

Device	RMS Current expressions
HV Transistor	$I_{AvgHVTransistor} = \frac{\frac{1}{2} * I_{L1} * \left(\frac{dT_s}{2} - t_B\right) + \frac{1}{2} * (I_{L1} + I_p) * \left(\frac{T_s}{2} - \frac{dT_s}{2}\right)}{\frac{T_s}{2} - t_B}$
HV Diode	$I_{AvgHVDiode} = \frac{1}{2} I_p$
LV Transistor	$I_{AvgLVTransistor} = \frac{1}{2} I_{L1}$
LV Diode	$I_{AvgLVDiode} = \frac{\frac{1}{2} * I_p * t_B + \frac{1}{2} * (I_{L1} + I_p) * \left(\frac{T_s}{2} - \frac{dT_s}{2}\right)}{\frac{T_s}{2} - \frac{dT_s}{2} + t_B}$

On top of that, the RMS value of the coupling inductor current can be calculated similarly and ends up being the following:

$$I_{RMS_{L_k}} = \sqrt{\frac{2}{T_s} \left[\frac{I_{L1}^2}{3} * \left(\frac{dT_s}{2} - t_B\right) + \left(\frac{T_s}{2} - \frac{dT_s}{2}\right) \left(I_{L1}^2 + \frac{(I_p - I_{L1})^2}{3} + I_{L1}(I_p - I_{L1}) \right) + \frac{I_p^2}{3} t_B \right]} \quad (3.7)$$

3.4.2. Boost mode

Similarly, the calculations in boost mode, or what is the same, when the power flows from the LV bridge to the HV bridge, that is when the battery is discharging into the DC-Bus. As in the previous mode, both the waveforms and equations have been deducted as in [16]. Figure 3.10 depicts the waveform for boost mode when $M = \frac{V_{bat}}{nV_{DC-Bus}} < 1$.

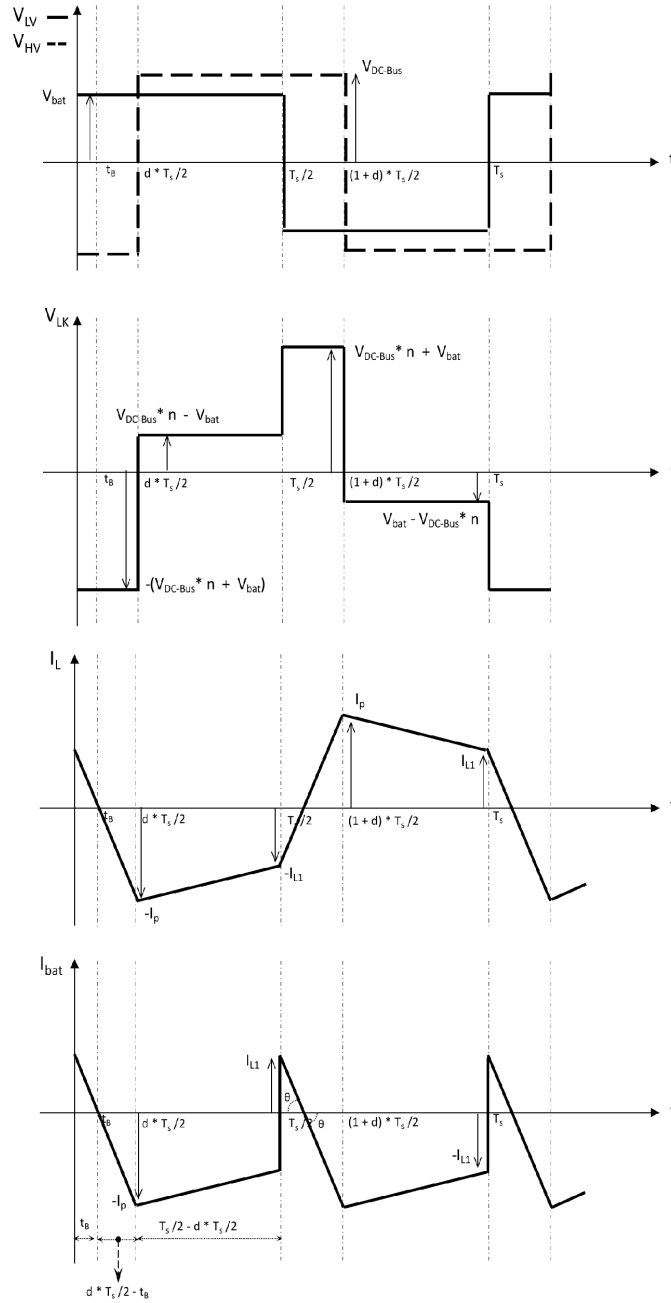


Figure 3.10.: Essential wave forms for the DAB in boost mode

All of the suppositions made in buck mode stay the same with the exception that now, the

LV square wave now leads the HV bridge one. This can be seen in Figure 3.10. By applying the same concepts in the calculations with the appropriate changes, the equations for the current result in the following:

- Current:

$$I_P = \frac{1}{4L * f_s} [nV_{DC-Bus} + V_{bat}(2d - 1)] \quad (3.8)$$

$$I_{L1} = \frac{1}{4L * f_s} [nV_{DC-Bus}(2d - 1) + V_{bat}] \quad (3.9)$$

As calculated, these current equations happen to be the same in both the buck and boost mode in value but interchanged in the time spectrum. However, the t_B changes and can be seen in equation (3.10):

- Time:

$$t_B = \frac{nV_{DC-Bus}(2d - 1) + V_{bat}}{4 * f_s [nV_{DC-Bus} + V_{bat}]} \quad (3.10)$$

Again, calculating the average for the output current I_{bat} and integrating in a similar way as in the buck mode, the average value results in:

$$I_{bat} = \frac{nV_{DC-Bus}}{2L * f_s} (d^2 - d) \quad (3.11)$$

Likewise, the ideal converter power is:

$$P_{bat} = \frac{nV_{DC-Bus}V_{bat}}{2L * f_s} (d^2 - d) \quad (3.12)$$

As it is easily seen, when normalizing the power again:

$$P'_{bat} = (d^2 - d) \quad (3.13)$$

Interpreting this result and combining it with the one in buck mode, we obtain equation (3.14), and normalizing it we obtain (3.15). It is easy to see that the maximum power is at $\pm d = 0.5$ and the sign only refers to the direction of the power flow. Figure 3.11 shows this in a graphical way. Note that in the sinusoid, the maximum and minimum are in $d = 0.5$ and $d = -0.5$ respectively.

$$P_{bat} = \frac{nV_{DC-Bus}V_{bat}}{2L * f_s} d(1 - |d|) \quad (3.14)$$

$$P'_{bat} = d(1 - |d|) \quad (3.15)$$

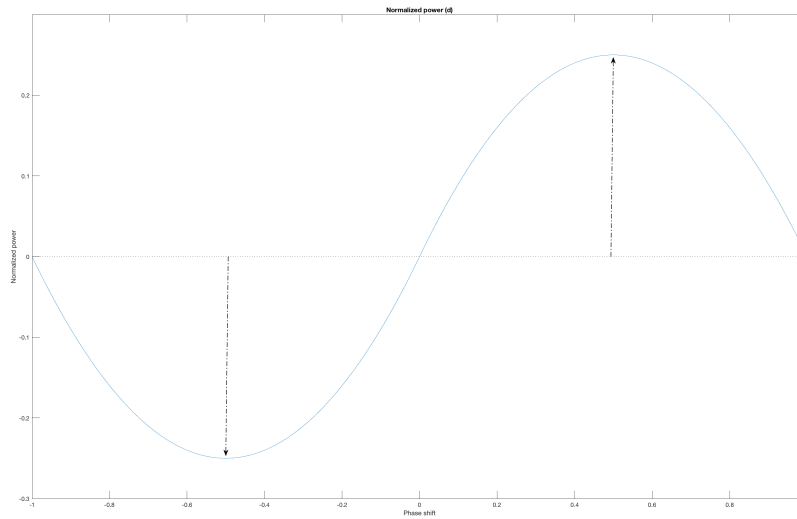


Figure 3.11.: Normalized power

Again, we calculate the RMS current through the inductor:

$$I_{RMS_{L_k}} = \sqrt{\frac{2}{T_s} \left[\frac{I_p^2}{3} * \left(\frac{dT_s}{2} - t_B \right) + \left(\frac{T_s}{2} - \frac{dT_s}{2} \right) \left(I_p^2 + \frac{(I_p - I_{L1})^2}{3} - I_p(I_p - I_{L1}) \right) + \frac{I_{L1}^2}{3} t_B \right]} \quad (3.16)$$

And, subsequently, the rest of the RMS and average current values in boost mode as listed in 3.3 and 3.4

Table 3.3.: RMS current values though the devices
when power flows from the LV side to the HV side or boost mode

Device	RMS Current expressions
HV Transistor	$I_{RMSHVTransistor} = \sqrt{\frac{1}{T_s} \left[\frac{I_p^2}{3} * \left(\frac{dT_s}{2} - t_B \right) \right]}$
HV Diode	$I_{RMSHVDiode} = \sqrt{\frac{1}{T_s} \left[\frac{I_{L1}^2}{3} * t_B + \left(\frac{T_s}{2} - \frac{dT_s}{2} \right) \left(I_p^2 + \frac{(I_p - I_{L1})^2}{3} - I_p(I_p - I_{L1}) \right) \right]}$
LV Transistor	$I_{RMSLVTransistor} = \sqrt{\frac{1}{T_s} \left[\frac{I_p^2}{3} * \left(\frac{dT_s}{2} - t_B \right) + \left(\frac{T_s}{2} - \frac{dT_s}{2} \right) \left(I_p^2 + \frac{(I_p - I_{L1})^2}{3} - I_p(I_p - I_{L1}) \right) \right]}$
LV Diode	$I_{RMSLVDiode} = \sqrt{\frac{1}{T_s} \left[\frac{I_{L1}^2}{3} * t_B \right]}$

Table 3.4.: Average current values through the devices
when power flows from the LV side to the HV side or boost mode

Device	RMS Current expressions
HV Transistor	$I_{AvghVTransistor} = \frac{1}{2} I_p$
HV Diode	$I_{AvghVDiode} = \frac{\frac{1}{2} * I_{L1} * t_B + \frac{1}{2} * (I_{L1} + I_p) * \left(\frac{T_s}{2} - \frac{dT_s}{2} \right)}{\frac{T_s}{2} - \frac{dT_s}{2} + t_B}$
LV Transistor	$I_{AvglVTransistor} = \frac{\frac{1}{2} * I_p * \left(\frac{dT_s}{2} - t_B \right) + \frac{1}{2} * (I_{L1} + I_p) * \left(\frac{T_s}{2} - \frac{dT_s}{2} \right)}{\frac{T_s}{2} - t_B}$
LV Diode	$I_{AvglVDiode} = \frac{1}{2} I_{L1}$

3.4.3. ZVS limits

As the efficiency of a converter is extremely important, various techniques are designed to improve this parameter. Our choice has been to impose ZVS conditions to achieve this purpose.

In order to meet ZVS conditions during commutation, two requirements must be satisfied:

- At turn on: the device body diode must be conducting.
- At turn off: the device current at that instant is greater than 0.

These conditions are easily understood if we think about them:

For the first condition, or turn on state, the body diode must be conducting. Hence, the voltage across this body diode is 0 or, in the case of a real diode, a forward voltage drop of typically of around 0.7 to 1.2 V. In this case, the body diode needs to be conducting in order to act as a clamp. Only after this, the ideal switch—that is part of the MOSFET— can turn on at ZVS or very low voltage if we take into account the forward drop.

For the second condition, since the body diode allows current to flow backwards, at turn off, the current must be flowing with a positive sign having for a reference the MOSFET typical path, that is, on an N-channel MOSFET, from drain to source. Otherwise, if the current is flowing through in the opposite direction, when the ideal switch turns off, the current could start flowing through the body diode, defeating the purpose and the control for ZVS.

If we translate these constraints into equations I_{L1} is the limiting current; therefore, we impose the constraint in equation (3.2) or (3.9), and obtain:

$$I_{L1} = \frac{1}{4L * f_s} [nV_{DC-Bus}(2d - 1) + V_{bat}] \geq 0 \quad (3.17)$$

Taking into account that the ratio M remains $M = \frac{V_{bat}}{nV_{DC-Bus}} < 1$, by solving the inequation for the phase-shift d we obtain:

$$d \geq 0.5 - \frac{M}{2} \quad (3.18)$$

However, if the ratio $M = \frac{V_{bat}}{nV_{DC-Bus}} > 1$, the limiting current is I_p making the constraint:

$$I_p = \frac{1}{4L * f_s} [nV_{DC-Bus} + V_{bat}(2d - 1)] \geq 0 \quad (3.19)$$

Resulting in a phase-shift of:

$$d \geq 0.5 - \frac{1}{2M} \quad (3.20)$$

Despite the fact that these constraints are correct for an ideal situation, in reality, MOSFETs have parasitic capacitances. The C_{ds} is of vital importance as it is the one that will act as a snubber and will clamp the diode.

Explained in a more precise way, not only will the body diode clamp the ideal MOSFET switch, but also, the snubber capacitance will clamp the body diode, making the losses of releasing the reverse recovery charge disappear. This means that at turn on, the losses of the MOSFET are virtually non-existent as the snubber capacitor is ideal and it clamps the body diode, which also clamps the switch, making the power effectively 0. Of course, needless to say, all of this analysis is considering the possible stray inductances, capacitances and resistors negligible.

Still, we have not have taken into account the charge that has to be forced in and out of these snubber capacitors. In order to do so, we will have to size the coupling inductor appropriately as the energy that it holds must be enough to charge and discharge the capacitors. To do so, the research followed by Kheraluwala [17] will be very helpful in our analysis.

The energy delivered from one full bridge to the other one can be obtained from two different parts: First, from the bridge itself, or in our case, the referred HV to the LV side. Secondly, from the coupling inductor, which is in charge of transmitting the power and interconnecting both bridges. The next equations represent these two statements respectively.

$$E_{delivered} = \int_0^{t_B} V'_{DC-Bus} i_L dt \quad (3.21)$$

$$E_{delivered} = \frac{1}{2} L_k I_{min}^2 \quad (3.22)$$

If we also take into account that, since when the MOSFETs turn off, the circuit behaves in a resonant manner while the energy stored in the inductor charges/discharges the snubber capacitors until the voltage of both capacitors reaches the value of the opposite rail. For that reason, the current of the coupling inductor is twice that which circulates through a snubber capacitor:

$$i_L = 2C_s \frac{dv_{C_s}}{dt} \quad (3.23)$$

If we now substitute this expression in the integral above and substitute the increment of the voltage of one of the capacitors for V_{bat} :

$$E_{delivered} = \int_0^{t_B} V'_{DC-Bus} 2C_s \frac{dv_{C_s}}{dt} dt = V'_{DC-Bus} 2C_s \int_0^{t_B} dv_{C_s} = 2V'_{DC-Bus} V_{bat} C_s \quad (3.24)$$

Equalizing this result to equation (3.22), simplifying and solving for I_{min} :

$$\frac{1}{2} L_k I_{min}^2 = 2V'_{DC-Bus} V_{bat} C_s \rightarrow I_{min} = 2 \sqrt{\frac{nV_{DC-Bus} V_{bat} C_s}{L_k}} \quad (3.25)$$

If we now incorporate the constraints of ZVS commutation for $M < 1$:

$$I_{L1} = \frac{1}{4L * f_s} [nV_{DC-Bus}(2d-1) + V_{bat}] \geq 2 \sqrt{\frac{nV_{DC-Bus} V_{bat} C_s}{L_k}} \quad (3.26)$$

And solving the inequation for the phase-shift d we obtain:

$$d \geq 0.5 - \frac{M}{2} + \frac{4L_k}{nV_{DC-Bus} T_s} \sqrt{\frac{nV_{DC-Bus} V_{bat} C_s}{L_k}} \quad (3.27)$$

However, if the ratio $M = \frac{V_{bat}}{nV_{DC-Bus}} > 1$, the limiting current is I_p making the constraint:

$$I_p = \frac{1}{4L * f_s} [nV_{DC-Bus} + V_{bat}(2d-1)] \geq 2 \sqrt{\frac{nV_{DC-Bus} V_{bat} C_s}{L_k}} \quad (3.28)$$

Again, resulting in a phase-shift d of:

$$d \geq 0.5 - \frac{1}{2M} + \frac{4L_k}{V_{bat} T_s} \sqrt{\frac{nV_{DC-Bus} V_{bat} C_s}{L_k}} \quad (3.29)$$

These values represent the boundaries for ZVS operation for M ratios under and over 1. It is also worth noting that a similar thought process is valid when calculating the minimum current

at the other side of the transformers referring the values appropriately to the HV side and instead of charging and discharging the snubber capacitors of the LV MOSFETs, doing so with the HV ones, although, the limiting condition is usually on the LV side due to the values of these snubber capacitors.

In some cases, we could additionally incorporate capacitors in parallel to the C_{ds} parasitic capacitor of the MOSFETs themselves. These snubber capacitors will allow us to reduce the $\frac{dv}{dt}$, helping with possible turn-off losses in the transients and lower electromagnetic interference (EMI). This will also reduce the current through the MOSFETs and reduce the possibility of switching device failure due to over-voltage. However, this entails a trade-off with losses in the never ideal capacitors and a bigger inductor to store more energy to charge or discharge these capacitors.

4. Design factors

In this section, several of the design parameter choice selection will be explained. A detailed although schematic exposition of the parameter decisions will be portrayed as this is an iterative process and all of the parameters affect the rest of them, making the unraveling of the specific details impractical.

As a baseline, the original specifications will be determined and then the rest of the devices and parameters will be clarified in a logical order.

4.1. Preliminary specifications

As this project has been designed as a proof of concept, most of the parameters were chosen with a general idea of achievable in the up-to-scale real-life design. Nevertheless, some of them have been adapted to the university laboratory requirement for safety purposes.

Among the realistic specifications we have the following:

- The battery's voltage: A standardized voltage was to be picked. In general, these include 12 Volts and its multiples. A battery of 48 Volts was chosen.
- The DC-Bus voltage: As the intended purpose of this converter is to stabilize the output power of a PV cell to the grid through a DC-Bus, it was clear, that the limiting factor was the connection to the grid.

A DC to AC inverter is needed when connecting the PV facility to the grid. This inverter function with Pulse Width Modulation (PWM) techniques. PWM basic principle of operation is to, by comparing a sinusoid with a triangular wave, encode this sinusoid in a train of pulses by varying the length of these pulses. This is how, by switching with a HF wave can obtain a sinusoid with some other harmonics associated with it.

As one may expect in the process of creating this sine wave, there are two important factors: the filtering stage and the width modulation. The former will provide a cleaner and more sinusoid-like wave to the grid, but it will cause some voltage drop. The latter will also cause a voltage drop, as encoding the sine wave will most definitely not use the full amplitude of the DC voltage provided - as by having pulses go through a value of 0 they will not utilize the entire voltage value-.

For these reasons, and by choosing to design for the typical values for grid frequency and voltage values (at a user or LV level), the average DC-Bus voltage was chosen to be 190 Volts for

a 60 Hz and 120 Volts of output.

The up to scale specification will be the power that the converter delivers. This restriction is due to the fact that for specified voltage values at both input and output; regardless of which is DC-Bus and which is the battery, as the DAB is a bi-directional converter; the current will be set. As it is commonly known, what is dangerous for the human health is the amount of current flowing, the path flow through the body and the duration of this incidence. For all of these reasons, the converter power output was to be close to 100 Watts. The specific number is not very important as it only viewed as a safety measure and other safety measures were taken.

4.2. Battery model

The battery was modelled as a constant voltage source. However, as the battery charges and discharges, the value of its voltage also changes. This was not taken into account during the research as we only focus on the steady state of the converter. For that reason, the battery value will be changed manually during simulations. This is only considered when the battery is acting as a supplier of power.

In order not to cause any trouble with the simulation, the same battery was modelled as a resistor with an appropriate value to draw the desired amount of power for the predetermined voltage.

The equation used for this is the following:

$$|R| = \frac{V_{bat}^2}{P_{bat}} = \frac{nV_{DC-Bus}V_{bat}^3}{2L * f_s} d(1 - |d|) \quad (4.1)$$

4.3. DC-Bus model

The case of the DC-Bus model is relatively similar to the battery. In this instance, the DC-Bus voltage will be considered constant at a value of 190 Volts. The reason is that when a control scheme is implemented, it will control the voltage value of the bus by injecting power from the battery or deciding to charge the battery and reduce the voltage of the DC-Bus and maintain it relatively constant. We will need to consider the voltage constant and consider negligible the change in voltage, as that will be the objective of the control scheme.

When speaking of power being output to the DC-Bus, the procedure is identical to the battery above, with the only difference that now the power will flow in the opposite direction. As one can see in the expression above, the value of the resistor will always be considered positive, regardless of the direction of the power because, if otherwise, the resistor will be generating power instead of drawing it.

4.4. Switching frequency

When considering the switching frequency, the iterative process starts. Some of the parameters we need to consider are the switching losses, the size of the components.

- Switching losses will be accounted for in one of the following sections but the basic principle is that although we have minimized them as much as possible, the existence of parasitics in the circuit as well as operation in cases of low load, will cause losses. These losses will be determined by how many times does the MOSFETs commute per second, that is, the switching frequency.
- The size of the components will also be determined by the switching frequency, especially the transformer and the coupling inductance. The transformer will need to be designed for high frequency operation but, as a benefit, we obtain a lighter and smaller device. The coupling inductance will need to store energy for the transfer of power between the bridges. As one will see in the following section, the inductor size is directly linked with the switching frequency. Of course, the switching devices are greatly impacted by the frequency that they operate at, but this will be studied in more detail later on.

For these reasons a switching frequency of $f_s = 80kHz$ has been chosen.

4.5. Inductor sizing

When accounting for the inductor sizing, the fact that it is a reactive element justifies that its value is closely related to the value of the frequency. What is more, as it is also a function of the output power, we should design it for the worse case scenario, that is, when the power delivered is maximum at a phase-shift $d = 0.5$.

An optimal value inductance is selected after solving equation (3.14) , for the inductance value, as the next expression portrays:

$$L_k = \frac{d(1-d)nV_{DC-Bus}V_{bat}}{2f_sP_{bat}} \quad (4.2)$$

The result of substituting the values explained in the previous expression is that the coupling inductance should have a value of $L_k = 30.28\mu H$

4.6. Switching devices: MOSFETs

When choosing switching devices, MOSFETs were chosen primarily because of the convenience of their availability and their benefits when implementing ZVS strategies. The existence of parasitic capacitances and specifically, the drain to source one, provides a lossless snubber capacitance that will act as a clamp when commutating. On top of that, since a high frequency was chosen to control their switching, a device that was able to withstand this type of stress was mandatory.

In comparison to other switching devices such as BJT, the low drain-source on state resistance RDS_{On} meaning a low voltage drop and hence a more efficient converter. What is more, MOSFETs are much smaller in size for the same specifications making their use completely aligned with a high density power converter.

If we make the comparison between a MOSFET and another state of the art switching device such as an IGBT the advantages for our design are apparent: first of all MOSFETs are comfortable working at high frequencies and where IGBTs tend to shine is in the 10 to 20 kHz range. Furthermore, as previously mentioned, the fact that the MOSFET has a drain to source capacitance helps to lower the $\frac{dv}{dt}$ that it will suffer, making the ZVS easier without adding extra snubber capacitors as you would need with an IGBT. This will also make the circuit layout easier as high $\frac{dv}{dt}$ and $\frac{di}{dt}$ create high levels of EMI.

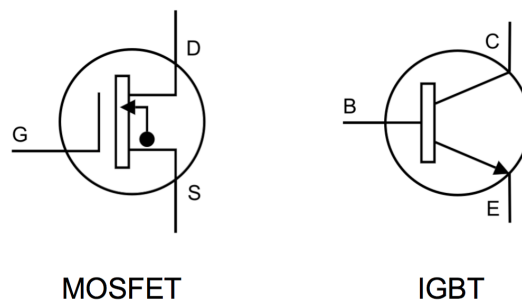


Figure 4.1.: MOSFET and IGBT models

4.7. Transformer

The transformer's main aspects of consideration are the turns ratio, the core material and the difference between the magnetizing inductance value and the coupling inductance value.

First and foremost the turns ratio of the transformer should consider that the ratio M should be of a value close to 1 when the battery is charged or discharged. For this reason a Value of $n = 0.25 \approx \frac{48}{190}$ has been chosen. When the battery is charged, $M = \frac{V_{bat}}{nV_{DC-Bus}} > 1$ and when its voltage is lower and has reached the lowest voltage we allow (24 V), $M = \frac{V_{bat}}{nV_{DC-Bus}} < 1$.

The core material has been decided to be a ferrite core as it withstands high frequencies better than other materials, however, the saturation level is a lot lower than in other materials and a trade-off is to be made.

The difference in the value of the coupling inductance (including the transformer's leakage inductance) should be big enough to be able to consider the magnetizing inductance negligible. If this is not the case, we risk saturating the core. For that reason, a value where the magnetizing inductance does not seem to have noticeable effects on the circuit at the selected frequency is around 250 times the value of the coupling inductance. This is unusual for a transformer of this low power, but the risk of core saturation leaves us no choice but to try and find a solution that will be discussed later on.

4.8. Battery capacitor

A capacitor was needed to provide a more steady voltage source in the battery. As it will be discussed later again in the DC-Bus capacitor, the main characteristic looked for was to have a voltage ripple specification lower than 5%.

For that reason, various calculations were taken into account, but more detail will be provided in the DC-Bus capacitor as, because of its changing nature, it makes more sense to explain it there. The calculations are totally transferable to this case just by changing the corresponding value of voltage, that is, substituting V_{DC-Bus} by V_{bat} and obtaining:

$$C_{bat} = \frac{\Delta Q}{0.05 * V_{bat}} \quad (4.3)$$

From this equation, we obtain a battery capacitor value of $3.8\mu F$.

4.9. DC-Bus capacitor

Imposing a maximum ripple specification is mandatory for the choice of an output capacitor. More so in case of providing power to a DC-Bus that will then be controlled by the current value of its voltage. As it was explained in the battery capacitor, the chosen specification for the output ripple will be 5% in both cases. When calculating the capacitance value, the procedure is identical in both cases.

A good strategy for calculating the capacitor value is to first assume that the capacitor is big enough to get all of the ripple of the battery current I_{bat} . That way, if we refer to Figure 3.9, we can obtain the ripple by “lowering” I_{bat} by its average value. In this way, we obtain a wave with an average of 0. This is what we call ripple.

Now, we need to calculate the charge the capacitor needs to be able to hold. Hence, and for convenience purposes, we have chosen the negative triangle in absolute value to calculate this charge. Now, this triangle has increased in height by the average value of I_{bat} and its width has also increased. To calculate this time we make use of similar triangles obtaining:

$$\Delta t = \frac{(I'_{L1} + I_{bat_{avg}}) \left(\frac{dT_s}{2} - t_B \right)}{IL1'} \quad (4.4)$$

Since the capacitor is on the HV side, every current must be reflected to that side. Thus, we are using $IL1'$ which is nothing else than $IL1 * n$.

We now calculate the charge that the capacitor needs to hold by calculating the area of the new triangle:

$$\Delta Q = \frac{1}{2} \Delta t (I'_{L1} + I_{bat_{avg}}) \quad (4.5)$$

Basing our calculations on the basic capacitor equation, we finally arrive at an expression for the capacitor value.

$$C_{DC-Bus} = \frac{\Delta Q}{0.05 * V_{DC-Bus}} \quad (4.6)$$

From this equation, we obtain a DC-Bus capacitor value of $0.2425 \mu F$.

5. Device Choice

This chapter's objective is to try to look for real components that could be implemented when constructing the converter. Some of the decisions driving the choice of the different components and elements will be briefly explained. The datasheets of the selected elements will be attached in an appendix for convenience purposes.

Although the final construction was not possible due to different limitations such as the time constraint, the components chosen should be sufficient for the construction of the converter. However, for obvious reasons, they have not been tested and any problems that have not been accounted for in the design will have to be dealt with.

5.1. Battery

A section was included for the battery to explain the some of the different alternatives of batteries available to us and the various pros and cons of their technologies. In the end, a possible battery will be chosen as an example for our specifications. This component is not specifically chosen for the main reason that the converter is designed for these specifications, however, while the battery voltage is within our constraints, the application may determine the type of battery used.

There are four main rechargeable battery technologies in the market. Their main benefits and drawbacks are:

- **Lead acid:** The electrodes are mainly made from lead usually mixed with another additive like calcium to give it mechanical strength. The electrolyte is sulfuric acid. Some of the characteristics of this type of battery are its low energy density, meaning heavy batteries and that it is not environmentally friendly. However, there are some advantages associated with his technology, various speeds of charging are allowed, overcharging is not an issue and its cost per kWh is one of the lowest in the battery market.
- **Nickel - Cadmium:** Typically the positive electrode is made of Nickel hydroxide and the negative electrode is composed of Cadmium hydroxide. Its main features are that unlike other types of batteries, such as lead acid, that decrease their voltage when they are discharging, NiCd batteries maintain their 1.2 Volts per cell until this is almost depleted. Additionally, they have a low internal resistance. This allows for quick charging and discharging as well as high power outputs. However, it is not as energy dense as some of the newer battery technologies and it can self-discharge if not in use. However, one of the biggest problems is its environmental effect as it is not possible to recycle them.
- **Nickel metal hydride:** The positive electrode uses nickel oxide hydroxide and the negative electrodes use a hydrogen-absorbing alloy. This type of battery has a high energy density, almost comparable to lithium-ion, and have around 50% the capacity of NiCd batteries. In addition, they are a lot more environmentally friendly than its other counterparts, but its charging control its complex and does not overcharge well generating big amounts of heat when charging and discharging.
- **Lithium-ion:** This battery is one of the most recent battery technologies to be explored, and for that reason, it is still immature. Nevertheless, it has some significant advantages over its competitors being energy density its strong suit. However, it is still an expensive alternative that requires complex charging control and is prone to self-discharging. If a correct charging control is not in place disastrous consequences such as fire appearance can occur.

Because of some of the characteristics stated above, we have chosen a Ni MH battery as an example of a battery for our application, as the technology is well studied, it has sufficient energy density for us and is environmentally friendly and as the application is to implement it on a PV facility it seems in line with the scope of the project of trying to improve clean sources of energy

and make them more viable. However, if we were implementing this converter in an electric vehicle or an aerospace project a Lithium-ion would be better suited as weight is extremely important.

As a first approximation, we have considered that our battery capacity should be of around $1Ah$. This number has been calculated by accounting for the maximum converter output power $100W$, the time this power could be delivered at once, estimating a standard maximum time of cloud covering of 15 minutes and accounting for 1 extra of them as a backup. Of course, the 48 V battery value should be included to make the calculation.

$$Battery\ Capacity = \frac{Max.\ Power(W) * (Total\ time\ of\ cloud\ appearance + backup)(hrs)}{Battery\ voltage(V)} = \frac{100 * 0.5}{48} \approx 1Ah \quad (5.1)$$

Another option will be to, instead of using a battery, to use an ultracapacitor that will serve as the battery.

As 48 V batteries are hard to come by, we have chosen a 12 V FDK battery and will put four of them in series to obtain the full 48 Volts.

The rest of the parameters can be found in the data-sheet in Appendix A.

5.2. Switching devices: MOSFETs

The switching devices that have been chosen are MOSFETs, as it was explained in some sections previous to this one. When choosing the devices specifically four key factors were considered:

- The voltage rating
- The current rating
- The switching frequency
- Loss related parameters

Of course, the analysis is identical for the HV bridge than for the LV one. For convenience purposes, only the HV bridge will be explained further. Please note that some of the values will be referred to the datasheet in Appendixes B and C.

When speaking about the voltage rating, the drain to source voltage should be higher than the maximum voltage allowed in the DC-Bus. This voltage should include the ripple that it was designed for. For that reason and by making some simple calculations we find that the voltage rating should be superior to 199.5 volts. Accounting for some voltage transients, we include a security factor of $SF = 1.2$. Multiplying both numbers we obtain a 239.4 Volts rating. The immediately superior standardized voltage is 250 Volts, which is our first design choice.

Similarly, the current that flows through our MOSFETs in the HV side should be lower than the rating that we choose. As the peak current is of around 1.24 Amperes the current rating should be higher than that. We apply the same reasoning as with the voltage and incorporate another security factor but this time higher, of $SF = 1.3$, since current could be more dangerous and could have higher spikes due to the resonant nature of part of the circuit commutation. If we consider that commutation is only a small fraction of the period, allowing for a steady state current of 1.7 Amps would not be a problem as the current allowed for transients (10s in the data-sheet). All in all, the value that the current should surpass is $I_{peak} * SF = 1.24 * 1.3 = 1.612A$ which is lower than the 1.7 A allowed in the data-sheet. Instantaneous values should be lower than 3 A. We now have our second design choice.

Switching frequency is not a problem in these types of power MOSFETs, but we will have to take into account the amount of heat that they will dissipate. This has been accounted for by possibly having heat sinks and allowing for high dissipation power in the MOSFETs, in this case, 1.56 W, which is a 1.56 % of the power per MOSFET. If we put this into perspective if all the devices dissipated that much power, only in the HV bridge we would be having losses of around 6.25 V or 6.25 %. This would mean that this bridge would be taking up almost all of the losses in the entirety of the circuit as we will see when seeing efficiencies higher than 90 %.

Finally, the loss related parameters that make up for significant losses such as the resistance from drain to source when conducting ($R_{ds_{on}}$), the forward voltage of the body diode or the rise and fall times. The criteria followed was to get as low levels as possible in all of them and to, in case of

choosing between two, choose the one with lowest conduction losses and switching losses should be minimized with the ZVS technique. The rest of the parameters can be found in the data-sheet in Appendixes B and C.

5.3. Snubber capacitors

As no high transients were detected when simulating in steady-state, it was assumed that the values of the parasitic capacitors, especially the drain-source capacitor C_{ds} was sufficiently large to reduce undesirable over-voltages and facilitate ZVS transitions.

5.4. Transformer

Two main parameters determine the preliminary design of the transformer: the turns ratio and the frequency. As it was explained before, a turns ratio of $n = 0.25$. This also means that the transformer could be turned around and end up with a $n = 4$ turns ratio. The material chosen was ferrite in order to reduce the losses related to frequency.

When designing the transformer a limitation was found after having chosen a material for the core, ferrite, and the turns ratio. Since the design parameter for this converter was the coupling inductor, if the strategy was to reduce the leakage inductor, the magnetizing inductance and the coupling inductance would end up being of the same magnitude, defeating the purpose of considering it negligible. However, if the strategy was to make the L_m as big as possible to have a desired difference between it and the coupling inductor, we could not find cores with high enough A_L or inductance per turn. For that reason and for constraints with time, the choice of the transformer has been left for future work, where a brief description of possible alternatives will be discussed.

5.5. Inductor

As the transformer could not be fully designed, the inductor of choice was to utilize the transformers leakage inductance and add an inductor that was closest to the value that would end up in a coupling inductance of $L_k = 30.28\mu H$ or superior. However, no specific inductor was chosen as it would be closely related to the transformer.

5.6. Battery capacitor

As explained before, the desired ripple for the battery capacitor is a 5 %. This affected the capacitance value that we obtained which ended up being $3.8\mu F$.

We chose devices that would minimize the stray inductances and resistances by opting for ceramic capacitors with little to no legs so that they could be directly placed on top of a printed circuit board (PCB). The reasoning behind choosing ceramic capacitors was similar, basically reducing as much as possible the equivalent series resistor (ESR). This will not only imply a closer ripple to the one designed for, but also reduce the overall losses in the converter.

Usually one of the main constraints associated with ceramic capacitors are the low capacitance value that they can reach, however, for our specific design the needed values have been found and therefore, this is no longer a disadvantage.

We have tried to minimize the ESR and the ESL (equivalent series inductance) among the different capacitors available, ending with a ceramic capacitor from TDK with a capacitance of $4.7\mu F$, which is the immediate value above our designed value of capacitance and an ESR of approximately $5m\Omega$.

The rest of the values can be found in the data-sheet attached in Appendix D.

5.7. DC-Bus capacitor

In a like manner, the DC-Bus capacitor was designed for the same ripple factor, but now, resulting in a capacitance value of $0.2425\mu F$.

The process was almost identical to the battery capacitor, and for a better reason, the ripple specification should be tried to be respected as a control scheme will then use the voltage values to function properly. For those reasons, the strategy of limiting the values of ESR and ESL remain the same. And since the capacitor is smaller, the fact that is ceramic does not pose a threat.

After following the said strategy we end up with another TDK capacitor, but this time with a $330nF$ and an approximate ESR value of $40m\Omega$.

The rest of the values can be found in the data-sheet attached in Appendix E.

5.8. Power loss calculation

A theoretical approach to the power loss incurred in the converter is necessary to obtain a preliminary approximation of the efficiency of the converter. We will consider negligible the losses of the cables and alike and will only consider the switching losses and the ESR losses in the capacitor. Since we do not have data for the transformer, algebraic expressions will be shown but will not be taken into account.

When speaking about the switching device losses we should distinguish between switching and conduction losses:

$$P_{HVTransistorConduction} = I_{HVTransistorRMS}^2 * R_{DSonHV} \quad (5.2)$$

$$P_{HVDiodeConduction} = V_{FwdDiodeHV} * I_{HVDiodeAvg} * \frac{tb}{Ts} \quad (5.3)$$

$$P_{HVTransistorSwitching} = 0.5 * V_{DC-Bus} * I_{PHV} * (t_{fHV} + t_{rHV}) * fs \quad (5.4)$$

The basic components of these losses are the switching loss itself, and the conduction losses through the transistor and the body diode. The reverse recovery charge of the diode has not been included because of the clamping effect that ZVS conditions impose. The area that the switching transition creates in triangle shapes, as one can see reflected in equation (5.4) . In reality, due to the imposed ZVS conditions, this losses will be much lower.

The transformer losses can be accounted by multiplying the inductor current by the value of the copper resistance and taking into account the skin and proximity effect. Another part of the losses of the transformer are those due to the core loss - by accounting for hysteresis and Eddy or Foucault related losses-, which can be calculated by a curve that models them with some parameters related to the core itself, however, none of these have been taken into account as we do not have a transformer choice:

$$P_{TransformerCu} = I_{LRMS}^2 * R_{Cu} \quad (5.5)$$

$$P_{Transformercore} = K * f^\alpha * B^\beta * Volume \quad (5.6)$$

When considering the reactive elements, we have to take into account the ESR of both the capacitors and the possible resistive loss in the coupling inductance.

$$P_{ESR} = I_{Capacitor_{RMS}}^2 * ESR \quad (5.7)$$

$$P_{Inductor} = I_{L_{RMS}}^2 * R_L \quad (5.8)$$

Efficiency values will be discussed in the simulation chapter.

6. Description of Research Results: Simulation

After all of the theoretical study, we try to validate the model by simulating it on a SIMULINK scheme and we will compare it to the theoretical results. Various simulations will be done in different situations to validate the model.

6.1. Charging the battery - Buck mode

First of all, we will consider the battery as fixed, with a value of 48 and the HV bridge being at the DC-Bus voltage at 190 V. We will consider the first case to be when the battery is discharging, and therefore, the power flow is in the direction of the HV bridge, that is, from the HV bridge to the LV bridge. Therefore, the simulation should be similar to that of the theoretical approach referred as Figure 3.9.

Since the M ratio is very close to 1, 1.0105 to be precise, the waveform for the inductor current should be relatively flat and any slight upward and downward direction may be due to the little difference from the M ratio of value 1 or even for the low losses incurred in the simulation, as the components have been tried to be represented as close to reality as possible. The first simulation can be viewed in Figure 6.1.

For a value of $d = 0.28$, the theoretical values are $I_P = 2.7225$ A, $I_{L1} = 2.7968$ A, $I_{bat} = 1.9766$ A, $I_{BUS} = 0.4993$ A and $I_{RMS} = 2.4889$ A. The calculated input power is 94.8745 W. These can be viewed in Figure 6.1. The slight differences may be due to the ESR of the capacitors of the losses in the real MOSFETs.

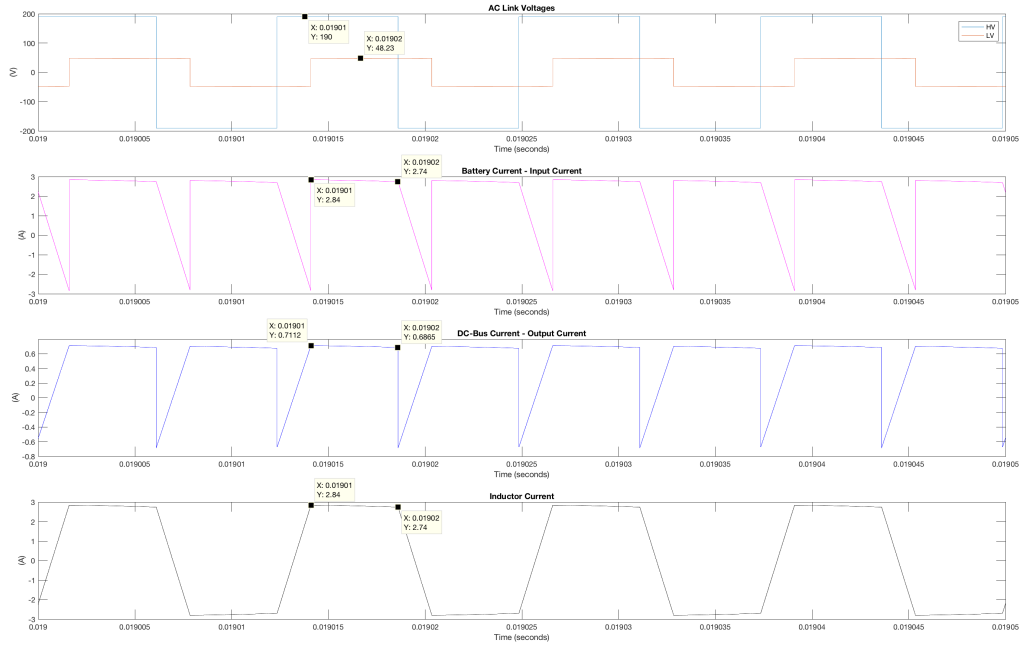


Figure 6.1.: Simulation results of the DAB converter when charging and $d = 0.28$. $V_{HV} = 190V$, $n = 1: 4$, $V_{LV} = 148V$, $P_{in_{max}} = 100/0.85$ W, $f_s = 80kHz$, $C_{BUS} = 330$ nF, $C_{bat} = 4.7$ μF ; and $L = 30.28$ μH .

An important thing to note is that in order to account for possible losses, an efficiency of 85 % has been assumed to calculate the input power hoping to end up in a 100 W output, although as one will see, the output power ends up being higher due to higher calculated and simulated efficiency.

Since the M ratio is very close to 1, 1.0105 to be precise, the waveform for the inductor current should be relatively flat and any slight upward and downward direction may be due to the little difference from the M ratio of value 1 or even for the low losses incurred in the simulation, as the components have been tried to be represented as close to reality as possible. The first simulation can be viewed in Figure 6.1.

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And for a value of $d = 0.0208$, we are in the ZVS boundary and the theoretical values are $I_P = 0.1545$ A, $I_{L1} = 0.2555$ A, $I_{bat} = 0.1997$ A, $I_{BUS} = 0.0504$ A and $I_{RMS} = 0.2057$ A. Now, the calculated input power is 9.5850 W. Figure 6.2 shows the validity of this values, but due to the nature of their small values, more relative error was incurred in them. The value calculated in equation (3.28) gives us a value of $I_L = 0.1543$ A which is very close to the value of $I_P = 0.1545$ A, which makes sense as it is the first of I_P and I_{L1} to reach 0, or in this case the calculated I_L value due to the snubber capacitors.

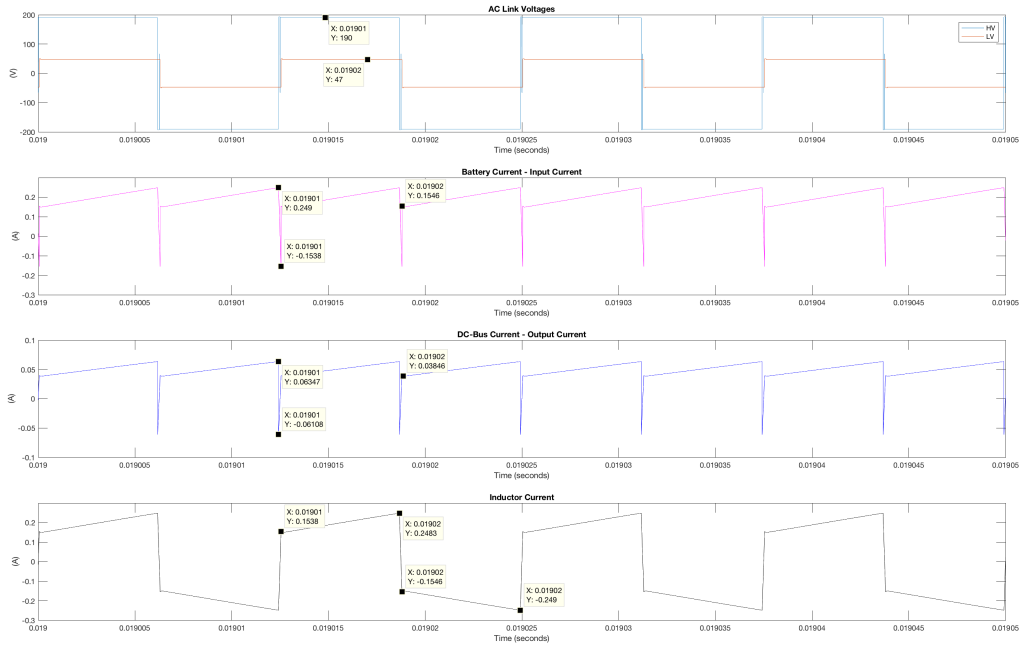


Figure 6.2.: Simulation results of the DAB converter when charging and in ZVS conditions with $d = 0.0208$.

VHV = 190V, $n = 1: 4$, VLV = 148V, $P_{in_{max}} = 100/0.85$ W, $f_s = 80$ kHz, $C_{BUS} = 330$ nF, $C_{bat} = 4.7$ μ F; and $L = 30.28$ μ H.

Finally, for a maximum power flow, occurring at $d = 0.5$, the theoretical values are $I_P = 4.9022$ A, $I_{L1} = 4.9538$ A, $I_{bat} = 2.4511$ A, $I_{BUS} = 0.6192$ A and $I_{RMS} = 4.0237$ A. The calculated input power is 117.6519 W, which is equivalent to the maximum power that we have fixed in our parameters by forcing $P_{in_{max}} = 100/0.85 = 117.647$ W. These values correlate well with those of the simulation provided in Figure 6.3. The theoretical efficiency is 92.99 % while the simulated efficiency is 97.30 %, nevertheless, the converter's efficiency should be around these values but experimentation is considered mandatory to confirm these results.

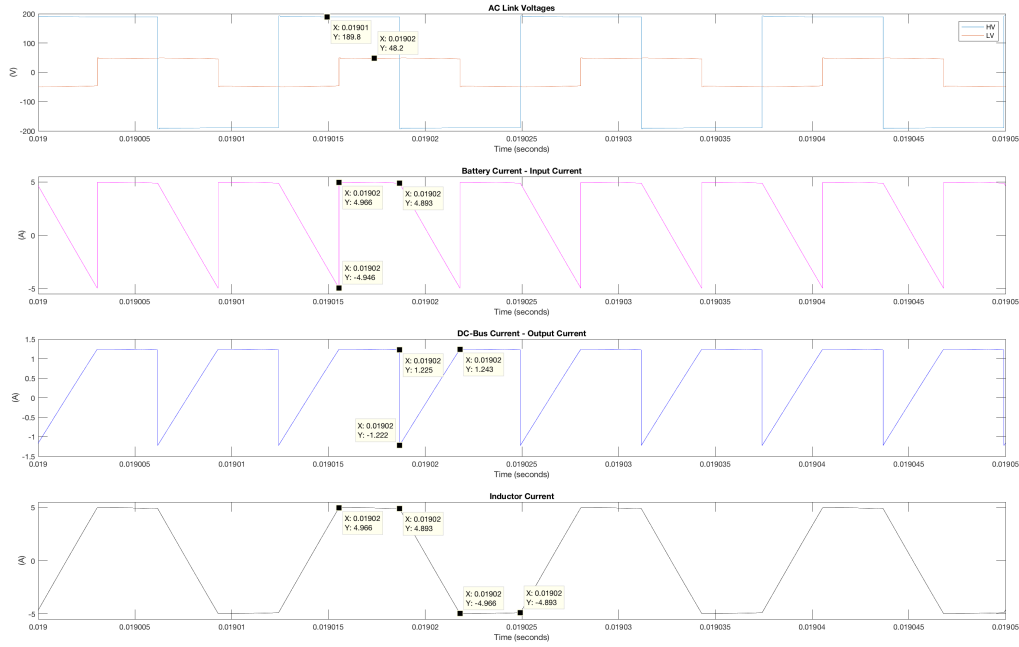


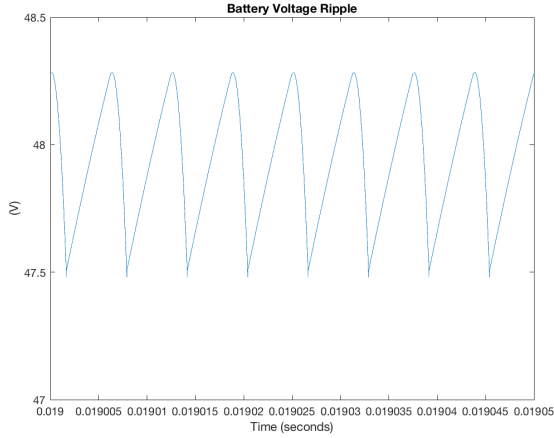
Figure 6.3.: Simulation results of the DAB converter when charging and maximum power flow conditions $d = 0.5$.

$V_{HV} = 190V$, $n = 1: 4$, $V_{LV} = 148V$, $P_{in_{max}} = 100/0.85$ W, $f_s = 80kHz$, $C_{BUS} = 330$ nF, $C_{bat} = 4.7 \mu F$; and $L = 30.28 \mu H$.

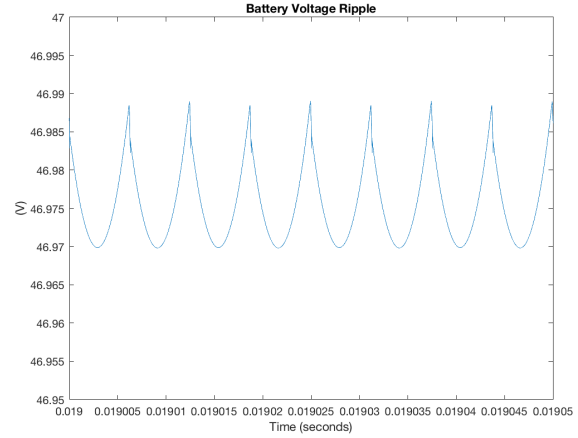
After all of the different points of operation simulated, we can state that the model is a good representation of reality. In the future, when the converter is constructed, the values may be somewhat different due to parasitics and other factors that we could not have taken into account in the theoretical approach or the simulation.

6.1.1. Battery voltage ripple - Output voltage ripple

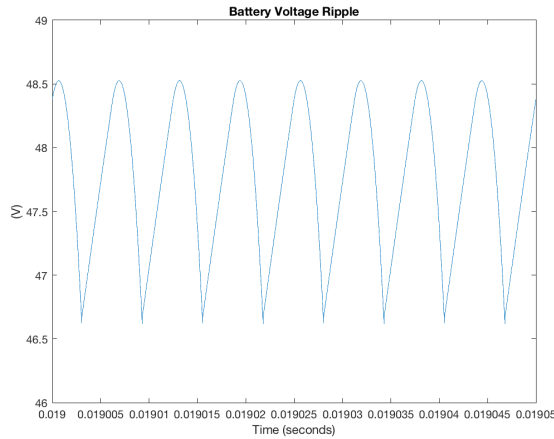
We will now show the output ripple, which in this case is the battery's voltage ripple. All of them should be well below the 5 % mark as the capacitor we chose was a little bit oversized as it was the first one available which suited our needs. Figure 6.4 represents the battery ripple for the different phase shift values simulated before.



(a) Voltage ripple with $d = 0.28$



(b) Voltage ripple with $d = 0.0208$



(c) Voltage ripple with $d = 0.5$

Figure 6.4.: Battery voltage ripples for the different values of phase shift d simulated.

As one can see the ripple in all of the cases is under the 5 % mark which would entail a voltage ripple of 2.4 Volts peak to peak.

6.2. Discharging the battery - Boost mode

In this mode, the battery should be charging, and therefore, the power flows from the LV bridge to the HV bridge. For this reason, the LV voltage should be leading the HV voltage. As both values of the DC-Bus and the battery have not changed, the shape of the waveforms will remain almost flat, especially the inductor current. This should correlate well with the waveform of Figure 3.10.

We have to keep in mind that some of the values like some of the average currents and average power are considered in absolute value as this only shows the direction of the power flow.

For a value of $d = 0.36$, the theoretical values are $I_p = 3.5151$ A, $I_{L1} = 3.5812$ A, $I_{bat} = 2.2589$ A, $I_{BUS} = 0.5707$ A and $I_{RMS} = 3.0933$ A. The calculated input power is 108.4280 W. These can be viewed in Figure 6.1. Again, we assume that the little differences may be due to the ESR of the capacitors or the losses in the real MOSFETs. In Figure 6.5 we can see the simulated waveforms that correlate well to the theoretical values.

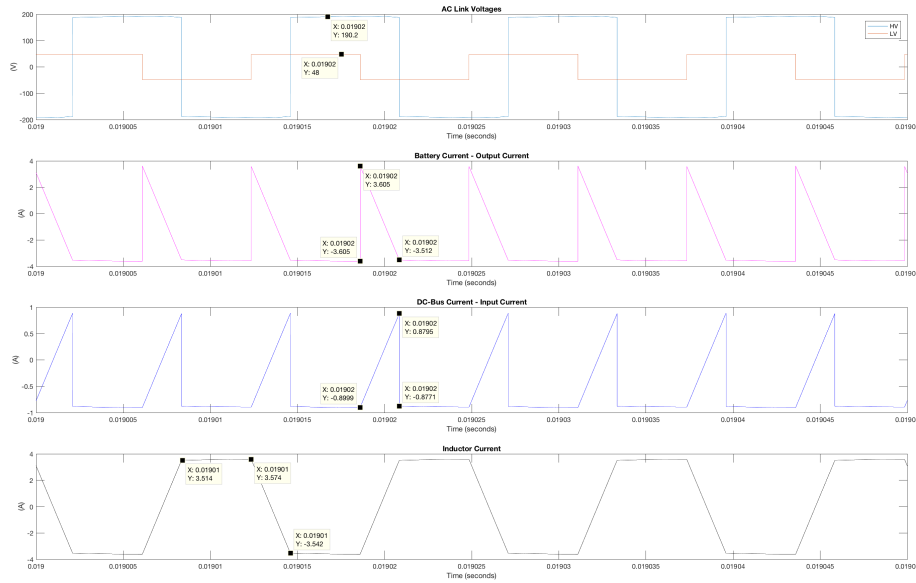


Figure 6.5.: Simulation results of the DAB converter when discharging and $d = 0.36$. $V_{HV} = 190$ V, $n = 1: 4$, $V_{LV} = 148$ V, $P_{in_{max}} = 100/0.85$ W, $f_s = 80$ kHz, $C_{BUS} = 330$ nF, $C_{bat} = 4.7$ μ F; and $L = 30.28$ μ H.

And for a value of $d = 0.0208$, we are in the ZVS boundary and the theoretical values are $I_P = 0.1545$ A, $I_{L1} = 0.2555$ A, $I_{bat} = 0.1997$ A, $I_{BUS} = 0.0504$ A and $I_{RMS} = 0.2057$ A. Now, the calculated input power is 9.5850 W. Figure 6.6 depicts this values, but due to the nature of their small values, more relative error was incurred in them. As it occurred in the buck mode, the value calculated in equation (3.28) gives us a value of $I_L = 0.1543$ A which is very close to the value of $I_P = 0.1545$ A. We consider this simulation valid, as the result as close enough.

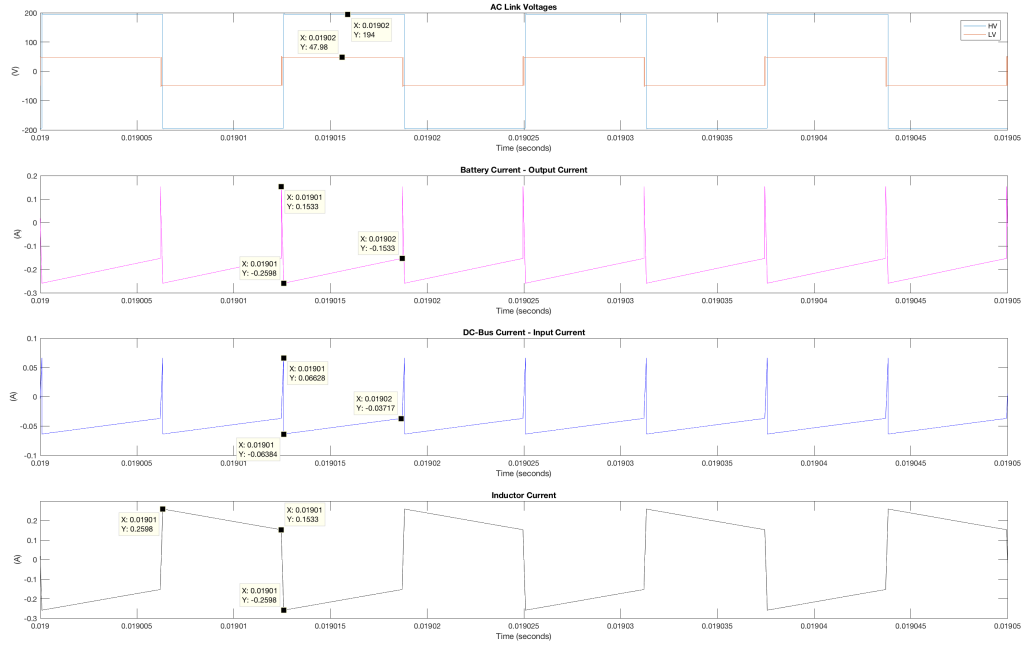


Figure 6.6.: Simulation results of the DAB converter when discharging and in ZVS conditions with $d = 0.0208$.

$V_{HV} = 190$ V, $n = 1: 4$, $V_{LV} = 148$ V, $P_{in_{max}} = 100/0.85$ W, $f_s = 80$ kHz, $C_{BUS} = 330$ nF, $C_{bat} = 4.7$ μ F; and $L = 30.28$ μ H.

Again, we fix the phase shift to $d = 0.5$ to obtain the maximum power flow, which is the same as in the buck case, $P_{in_{max}} = 100/0.85 = 117.647$ W. The theoretical values correlate well with those of the simulation provided in Figure 6.7 and are $I_P = 4.9022$ A, $I_{L1} = 4.9538$ A, $I_{bat} = 2.4511$ A, $I_{BUS} = 0.6192$ A and $I_{RMS} = 4.0237$ A. The calculated input power is also 117.6519 W. The theoretical efficiency is 91.75 % while the simulated efficiency is 97.28 %, nonetheless, this results should be taken with caution as there is a great deal of difference between them, but the converter's efficiency should be around these values. Experimentation will be needed to confirm this.

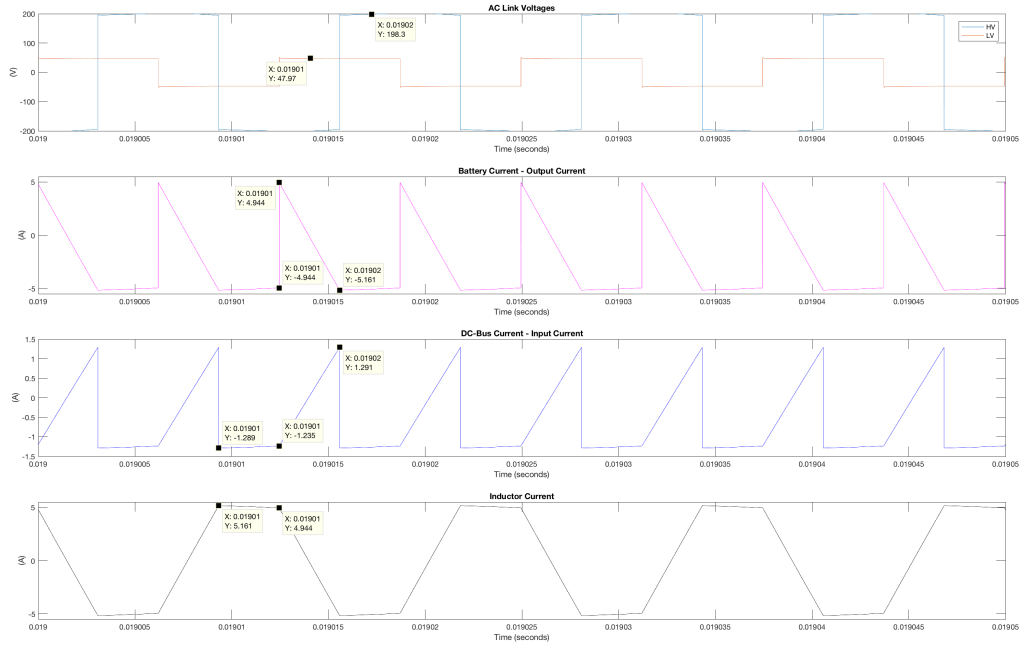
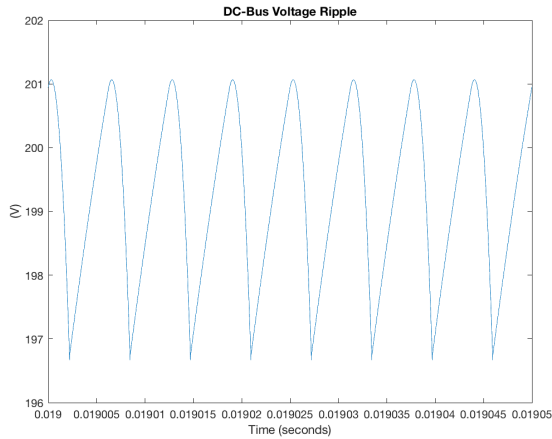


Figure 6.7.: Simulation results of the DAB converter when discharging in maximum power flow conditions with $d = 0.5$. $V_{HV} = 190V$, $n = 1: 4$, $V_{LV} = 148V$, $P_{in_{max}} = 100/0.85$ W, $f_s = 80kHz$, $C_{BUS} = 330$ nF, $C_{bat} = 4.7 \mu F$; and $L = 30.28 \mu H$.

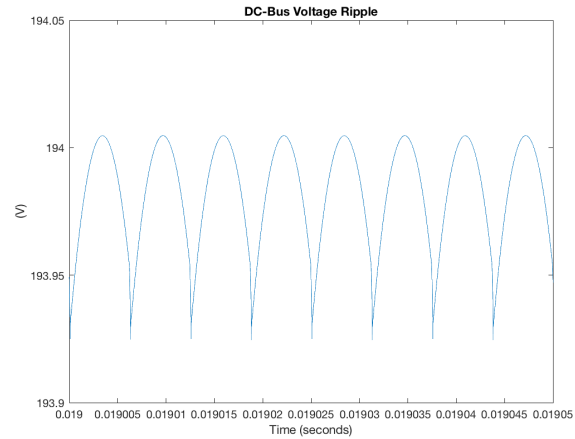
We end up considering the boost model as valid as well, as the simulations match up nicely with the ideal model.

6.2.1. DC-Bus voltage ripple - Output voltage ripple

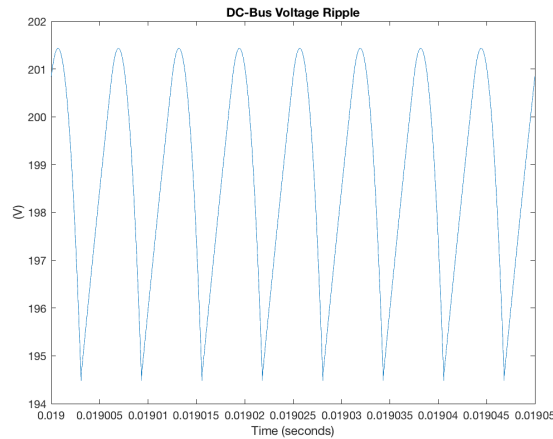
The DC-Bus' voltage ripple is an important parameter, as a future control scheme will use the values of the DC-Bus voltage to regulate the phase shift, and therefore, the amount of power that enters or leaves the battery. The voltage ripple for all cases should as well be under the 5 % mark as the capacitor we chose was oversized for the same reason as before. Figure 6.8 represents the battery ripple for the different phase shift values simulated in the previous cases for the boost mode.



(a) Voltage ripple with $d = 0.36$



(b) Voltage ripple with $d = 0.0208$



(c) Voltage ripple with $d = 0.5$

Figure 6.8.: DC-Bus voltage ripples for the different values of phase shift d simulated.

The ripple in all of the cases falls below the 5 % target which in this case would mean a voltage ripple of 9.5 Volts peak to peak in the DC-Bus.

7. Limitations

During the course of the design of this converter, we have encountered various limitations, especially concerning the design of the transformer. This converter bases its principle of operation in the existence of a coupling inductance that works as the main element that transfers power between both bridges, also known as AC-link.

As it has been explained during this thesis, usually the leakage inductance of a transformer is not big enough to satisfy the requirements of frequency, power, voltages and in some cases the snubber capacitors. The reality is that usually, an extra inductor is placed in series with the transformer's leakage inductance (referred to one side) in order to achieve the value needed for the coupling inductance. For that reason, we normally design a transformer for the desired turns ratio and a magnetizing inductance that is big enough to limit the current flow through the magnetizing branch. However, usually, this types of transformers have leakage inductances of values that are approximately 2 % to 5 % of the magnetizing inductance. This means that even when we design for a big enough magnetizing inductance, the leakage inductance will always be of a value of the same order of magnitude (around 20-50 %), when we would need much higher values to consider negligible this branch.

On top of that, if this condition is not satisfied, a dual active bridge topology will lead to almost certain saturation of the transformer's core, making the transformer only valid for a few periods, which at high frequencies can result in transformers lasting only a few seconds.

The first attempt would be to try to enlarge the magnetizing inductance as much as possible to have room to add inductance to the leakage inductance to get to the desired coupling inductance value. In consequence, a good proposition to augment the magnetizing inductance would be to choose a gapped core as this inductance is almost proportional to the gap size (if we consider negligible the reluctance of the core, which is usually the case), but as one can see intuitively, the more we open the gap - even if it is just a little- the more magnetic leakage will take place. For that reason, we reject this option.

If, on the contrary, we tried to design the transformer assuming a value for the leakage inductance in relationship to the magnetizing inductance, we would find one of two things: either the assumption was too optimistic, as we have just explained before for the typical values of this relationship, or the transformer has the characteristics of a bigger and higher power one, which would not only not satisfy the safety constraints of the laboratory, but will most likely have some other disadvantages like more losses, weight and bigger dimensions.

The specific parameters for the design of this converter have lead to an extremely difficult design of the transformer, but, after researching the problem, some solutions have been found and

will be discussed in the following section: future work, to see how, with less of a time constraint, this issue could be addressed.

8. Future work

As the full scope of the project has not been completed, some future work is mandatory. This prospective work will be detailed in the following sections.

8.1. Transformer design

As we have explained before, one of the limitations encountered in this project has been the inability to design a transformer that satisfied our needs. Hence, more research should be made in terms of the design of a transformer that can do so.

If this solution cannot be materialized, we have researched and learned, that by implementing resonant features to the converter, like by placing a capacitor in series with the coupling inductance, we would create a resonant series tank that would reset the flux of the magnetizing inductance. However, more research and understanding would need to be taken in this area to adapt the equations accordingly.

8.2. Control design

As it has been referred throughout this thesis, in order to control the power flow direction and quantity, we need to control the phase shift between both active bridges. To do so, we would need to implement a control scheme that controls it.

As a first approximation, a good idea would be to create a closed loop and start by implementing a P (proportional) control. Since this may not be enough, the next logical step would be to get a PI (proportional integral) control that tried to minimize the error. In case that this does not work, a PID (proportional integral derivative) should be put in place, however, this is likely not the case as the only thing that the control would need to do is to measure the DC-Bus voltage and compare it with some threshold values of for example 15 volts up or down, that is, from 175 down, the battery should deliver power to the DC-Bus and for voltage values of above 205, it should charge the battery. Of course, overcharging directives should also be put in place for a correct charging of the battery.

8.3. Construction

Finally, when the transformer is designed, the hardware could be bought and assembled, tested at light conditions first, to the test it at full power. Of course, as this project itself is a proof of concept, this model should need to be revalued for more realistic power values and the parameters should

need to be recalculated. Both the control and the switching of the MOSFETs will be controlled with microcontrollers.

When constructing some extra considerations such as the dead time necessary to implement between the top and bottom MOSFET of the same leg. This dead time should be higher than the MOSFET switching time and the charging time of the snubber capacitor. Among the other factors, EMI should also be taken into account when designing the printed circuit board or PCB and heat dissipation should be also considered in case heat sinks or even fans need to be implemented.

9. Conclusions

Implementing energy storage in our world through isolated DC-DC converters is becoming a reality, especially in the renewable energy and smart grids areas. After having studied the dual active bridge topology and having viewed its advantages and drawbacks, it is concluded that while not having completed the full scope of the experiment, the initial results are promising. With more time on our hands, this project should be completed and more insights may be discovered.

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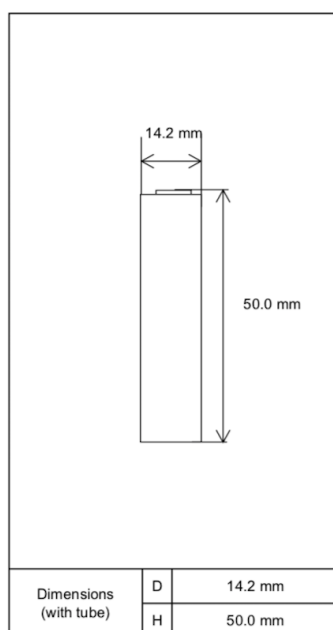
APPENDIXES

A. Battery data sheet

FDK
TWICELL

Cell Type HR-AAUC

Specifications



Type : Nickel-Metal Hydride Battery			Size : AA
Capacity ₁₎	Typical	1200mAh	
	Minimum	1100mAh	
Nominal Voltage			1.2V
Charging Current x Time		Fast Charge ₂₎	1200 mA x about 1.1h
Ambient Temp.	Charge Condition	Fast Charge ₂₎	0°C - 40°C
	Discharge Condition		0°C - 50°C
	Storage Condition	Less than 30days	-20°C - 50°C
		Less than 90days	-20°C - 40°C
		Less than 1year	-20°C - 30°C
Internal Impedance ₃₎ (after discharge to E.V.=1.0V)			Approx. 20mQ(at 1000Hz)
Weight ₄₎			Approx. 22 g
Size ₄₎ : (Diameter) x (Height)			14.2(D) x 50.0(H) mm

1)Single cell capacity under the following condition.

Charge : 120mA×16h, Discharge : 240mA(E.V.=1.0V) at 20°C

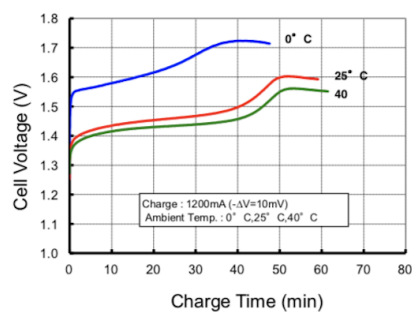
2)Use recommended charging system.

3)After a few charge and discharge cycles under the above 1) condition.

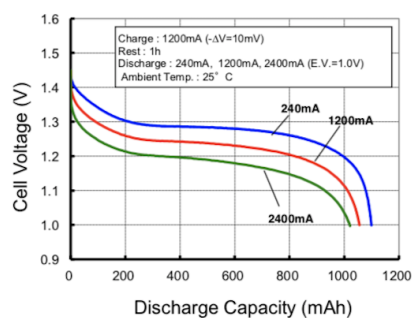
4)With tube.

Typical Characteristics

Charge



Discharge



2013/7

B. HV MOSFET data sheet



Si4434DY

Vishay Siliconix

N-Channel 250-V (D-S) MOSFET

PRODUCT SUMMARY

V _{DS} (V)	R _{rDS(on)} (Ω)	I _D (A)
250	0.155 at V _{GS} = 10 V	3.0
	0.162 at V _{GS} = 6.0 V	2.9

FEATURES

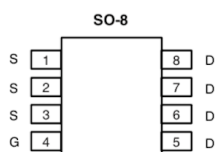
- Halogen-free According to IEC 61249-2-21 Definition
- PWM-Optimized TrenchFET® Power MOSFET
- 100 % R_g Tested
- Avalanche Tested



RoHS
COMPLIANT
HALOGEN
FREE
Available

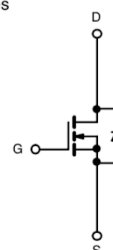
APPLICATIONS

- Primary Side Switch In:
 - Telecom Power Supplies
 - Distributed Power Architectures
 - Miniature Power Modules



Top View

Ordering Information: Si4434DY-T1-E3 (Lead (Pb)-free)
Si4434DY-T1-GE3 (Lead (Pb)-free and Halogen-free)



N-Channel MOSFET

ABSOLUTE MAXIMUM RATINGS T_A = 25 °C, unless otherwise noted

Parameter	Symbol	10 s	Steady State	Unit
Drain-Source Voltage	V _{DS}	250		V
Gate-Source Voltage	V _{GS}	± 20		
Continuous Drain Current (T _J = 150 °C) ^a	I _D	3.0	2.1	A
		2.4	1.7	
Pulsed Drain Current	I _{DM}	30		
Continuous Source Current (Diode Conduction) ^a	I _S	2.6	1.3	
Avalanche Current	I _{AS}	13		
Single Pulse Avalanche Energy	E _{AS}	8.4		mJ
Maximum Power Dissipation ^a	P _D	3.1	1.56	W
		2.0	1.0	
Operating Junction and Storage Temperature Range	T _J , T _{stg}	- 55 to 150		°C

THERMAL RESISTANCE RATINGS

Parameter	Symbol	Typical	Maximum	Unit
Maximum Junction-to-Ambient ^a	R _{thJA}	33	40	°C/W
		65	80	
Maximum Junction-to-Foot (Drain)	R _{thJF}	17	21	

Notes:

a. Surface Mounted on 1" x 1" FR4 board.

C. LV MOSFET data sheet



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March 2015

FDD5612

FDD5612

60V N-Channel PowerTrench® MOSFET

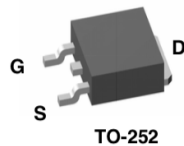
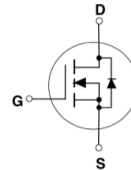
General Description

This N-Channel MOSFET has been designed specifically to improve the overall efficiency of DC/DC converters using either synchronous or conventional switching PWM controllers.

These MOSFETs feature faster switching and lower gate charge than other MOSFETs with comparable $R_{DS(ON)}$ specifications. The result is a MOSFET that is easy and safer to drive (even at very high frequencies), and DC/DC power supply designs with higher overall efficiency.

Features

- 18 A, 60 V. $R_{DS(ON)} = 55 \text{ m}\Omega$ @ $V_{GS} = 10 \text{ V}$
 $R_{DS(ON)} = 64 \text{ m}\Omega$ @ $V_{GS} = 6 \text{ V}$
- Optimized for use in high frequency DC/DC converters.
- Low gate charge.
- Very fast switching.



Absolute Maximum Ratings $T_A = 25^\circ\text{C}$ unless otherwise noted

Symbol	Parameter	Ratings	Units
V_{DS}	Drain-Source Voltage	60	V
V_{GS}	Gate-Source Voltage	± 20	V
I_D	Drain Current – Continuous (Note 1)	18	A
	Drain Current – Pulsed (Note 1a)	5.4	
	Drain Current – Pulsed	100	
P_D	Maximum Power Dissipation (Note 1)	42	W
	(Note 1a)	3.8	
	(Note 1b)	1.6	
T_J, T_{STG}	Operating and Storage Junction Temperature Range	-55 to $+175$	$^\circ\text{C}$

Thermal Characteristics

$R_{\theta JC}$	Thermal Resistance, Junction-to-Case (Note 1)	3.5	$^\circ\text{C/W}$
$R_{\theta JA}$	Thermal Resistance, Junction-to-Ambient (Note 1a)	40	$^\circ\text{C/W}$
		96	

Package Marking and Ordering Information

Device Marking	Device	Reel Size	Tape width	Quantity
FDD5612	FDD5612	13"	16mm	2500 units

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FDD5612 REV. 1.3

FDD5612

Electrical Characteristics $T_A = 25^\circ\text{C}$ unless otherwise noted

Symbol	Parameter	Test Conditions	Min	Typ	Max	Units
Drain-Source Avalanche Ratings (Note 1)						
W_{DSS}	Single Pulse Drain-Source Avalanche Energy	$V_{DD} = 30\text{ V}$, $I_D = 5.4\text{ A}$			90	mJ
I_{AR}	Maximum Drain-Source Avalanche Current				5.4	A

Off Characteristics

BV_{DSS}	Drain-Source Breakdown Voltage	$V_{GS} = 0\text{ V}$, $I_D = 250\text{ }\mu\text{A}$	60			V
$\frac{\Delta BV_{DSS}}{\Delta T_J}$	Breakdown Voltage Temperature Coefficient	$I_D = 250\text{ }\mu\text{A}$, Referenced to 25°C		62		mV/ $^\circ\text{C}$
I_{DSS}	Zero Gate Voltage Drain Current	$V_{DS} = 48\text{ V}$, $V_{GS} = 0\text{ V}$			1	μA
I_{GSSF}	Gate-Body Leakage, Forward	$V_{GS} = 20\text{ V}$, $V_{DS} = 0\text{ V}$			100	nA
I_{GSSR}	Gate-Body Leakage, Reverse	$V_{GS} = -20\text{ V}$, $V_{DS} = 0\text{ V}$			-100	nA

On Characteristics (Note 2)

$V_{GS(th)}$	Gate Threshold Voltage	$V_{DS} = V_{GS}$, $I_D = 250\text{ }\mu\text{A}$	1	2.4	3	V
$\frac{\Delta V_{GS(th)}}{\Delta T_J}$	Gate Threshold Voltage Temperature Coefficient	$I_D = 250\text{ }\mu\text{A}$, Referenced to 25°C		-6		mV/ $^\circ\text{C}$
$R_{DS(on)}$	Static Drain-Source On-Resistance	$V_{GS} = 10\text{ V}$, $I_D = 5.4\text{ A}$ $V_{GS} = 6\text{ V}$, $I_D = 5\text{ A}$ $V_{GS} = 10\text{ V}$, $I_D = 5.4\text{ A}$, $T_J = 125^\circ\text{C}$		36 42 64	55 64 103	m Ω
$I_{D(on)}$	On-State Drain Current	$V_{GS} = 10\text{ V}$, $V_{DS} = 5\text{ V}$	20			A
g_{FS}	Forward Transconductance	$V_{DS} = 5\text{ V}$, $I_D = 5.4\text{ A}$		15		S

Dynamic Characteristics

C_{iss}	Input Capacitance	$V_{DS} = 30\text{ V}$, $V_{GS} = 0\text{ V}$		660		pF
C_{oss}	Output Capacitance	$f = 1.0\text{ MHz}$		79		pF
C_{rss}	Reverse Transfer Capacitance			36		pF

Switching Characteristics (Note 2)

$t_{d(on)}$	Turn-On Delay Time	$V_{DD} = 30\text{ V}$, $I_D = 1\text{ A}$		8	16	ns
t_r	Turn-On Rise Time	$V_{GS} = 10\text{ V}$, $R_{GEN} = 6\text{ }\Omega$		4	8	ns
$t_{d(off)}$	Turn-Off Delay Time			24	38	ns
t_f	Turn-Off Fall Time			4	8	ns
Q_g	Total Gate Charge	$V_{DS} = 30\text{ V}$, $I_D = 5.4\text{ A}$		7.5	11	nC
Q_{gs}	Gate-Source Charge	$V_{GS} = 10\text{ V}$		2.5		nC
Q_{gd}	Gate-Drain Charge			3		nC

Drain-Source Diode Characteristics and Maximum Ratings

I _S	Maximum Continuous Drain-Source Diode Forward Current				2.7	A
V _{SD}	Drain-Source Diode Forward Voltage	V _{GS} = 0 V, I _S = 2.7 A (Note 2)		0.8	1.2	V

Notes:

1. $R_{\theta JA}$ is the sum of the junction-to-case and case-to-ambient thermal resistance where the case thermal reference is defined as the drain tab.

$R_{\theta JA}$ is the guaranteed design while $R_{\theta JA}$ is determined by the user's design. $R_{\theta JA}$ has been used to determine some of the maximum ratings.



a) $R_{\theta JA} = 40^\circ\text{C/W}$ when mounted on a 1 in^2 pad of 2oz copper.



b) $R_{\theta JA} = 96^\circ\text{C/W}$ when mounted on a 0.076 in^2 pad of 2oz copper.

Scale 1 : 1 on letter size paper

2. Pulse Test: Pulse Width $\leq 300\text{ }\mu\text{s}$, Duty Cycle $\leq 2.0\%$

FDD5612 Rev. 1.3

FDD5612

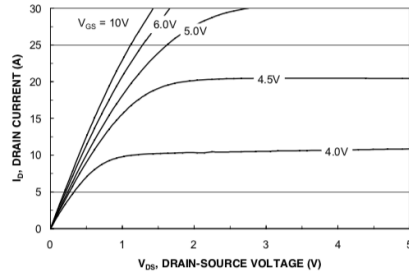
Typical Characteristics

Figure 1. On-Region Characteristics.

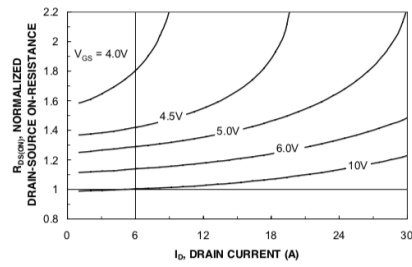


Figure 2. On-Resistance Variation with Drain Current and Gate Voltage.

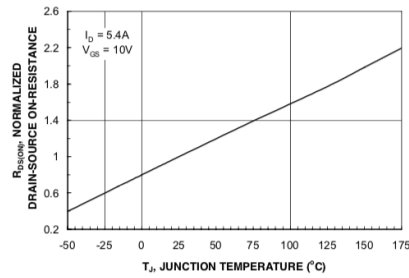


Figure 3. On-Resistance Variation with Temperature.

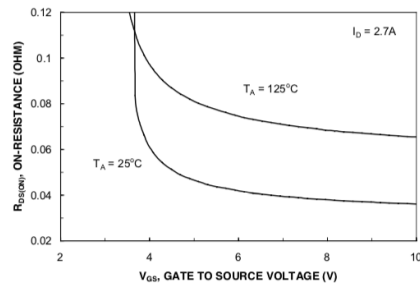


Figure 4. On-Resistance Variation with Gate-to-Source Voltage.

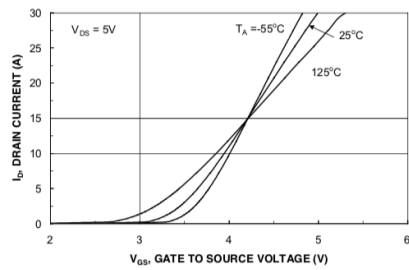


Figure 5. Transfer Characteristics.

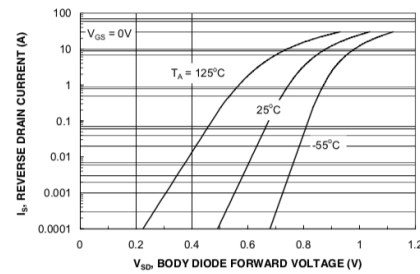
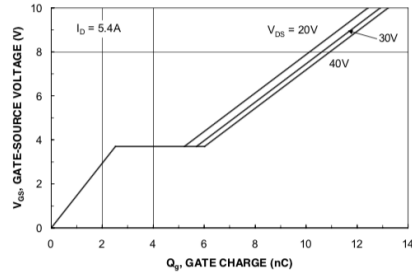
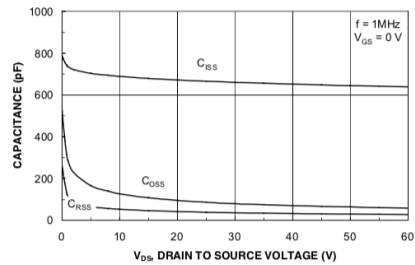
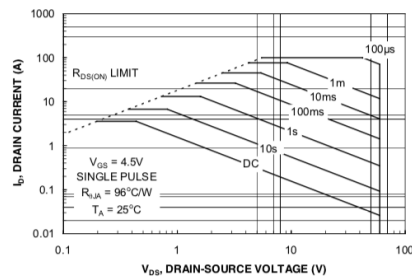
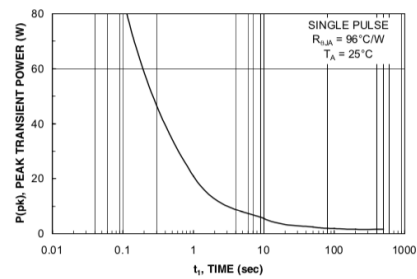
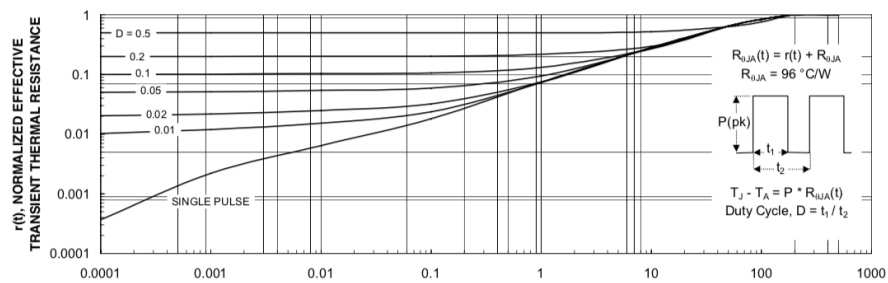


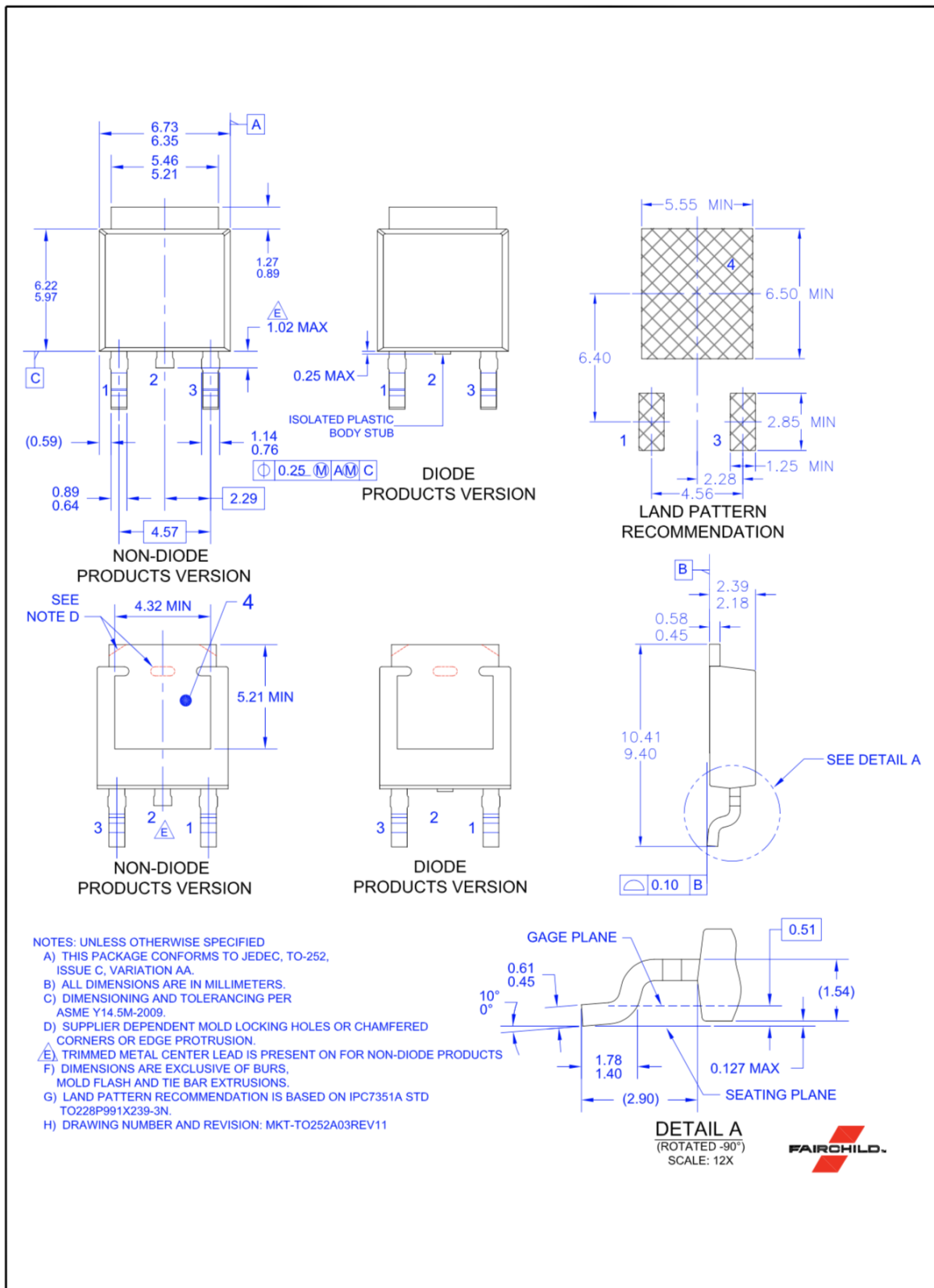
Figure 6. Body Diode Forward Voltage Variation with Source Current and Temperature.

FDD5612

Typical Characteristics**Figure 7. Gate Charge Characteristics.****Figure 8. Capacitance Characteristics.****Figure 9. Maximum Safe Operating Area.****Figure 10. Single Pulse Maximum Power Dissipation.****Figure 11. Transient Thermal Response Curve.**

Thermal characterization performed using the conditions described in Note 1b.
Transient thermal response will change depending on the circuit board design.

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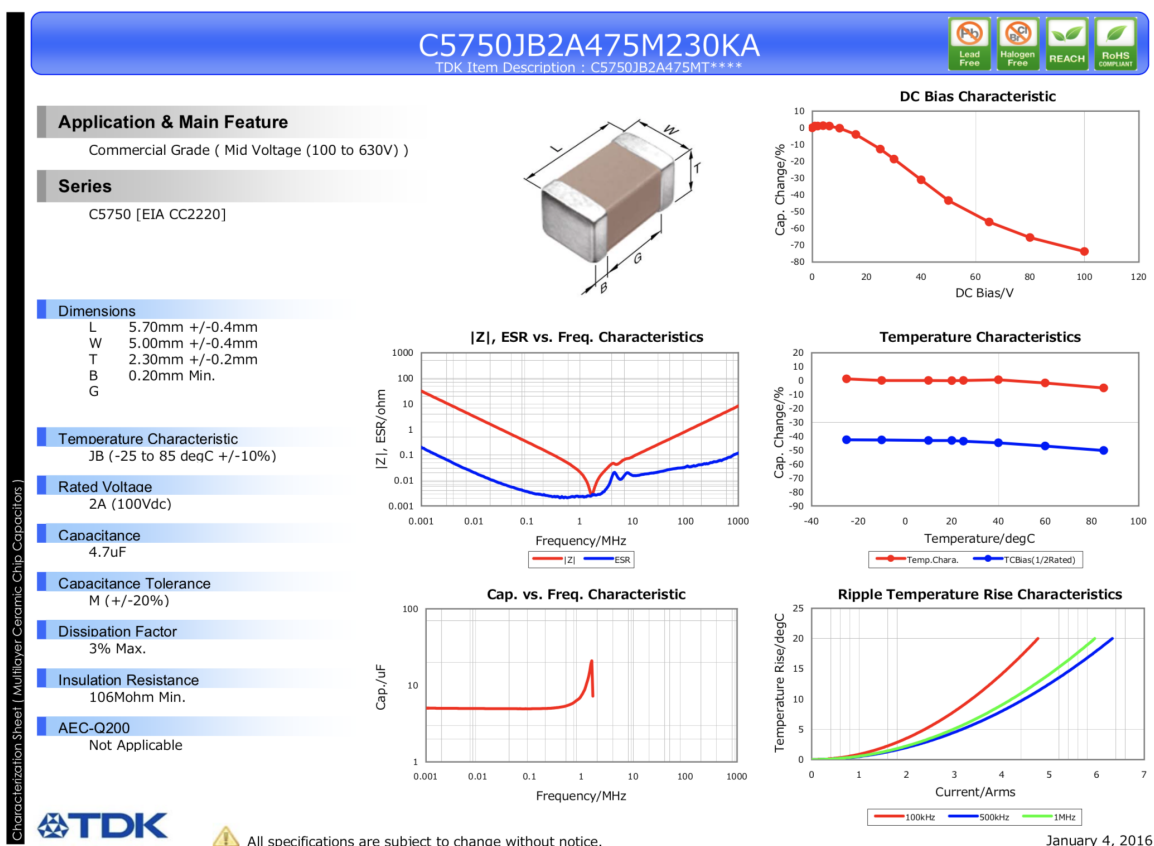
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D. Battery capacitor data sheet



E. DC-Bus capacitor data sheet

