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Design and Implementation of a Wireless Power Transfer System for Underwater Vehicles

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Resumo

O presente trabalho refere-se à otimização de um sistema de transferência de potência sem contacto em condições subaquáticas. A transferência de potência sem fios no oceano é especialmente crítica devido à energia dissipada na água salgada, sendo um meio condutor e responsável pela maior parte das perdas existentes, aumentando significativamente com a distância entre o recetor e o transmissor.

Um sistema típico de ressonância indutiva para transmissão de potência sem fios é tipicamente composto por um transmissor de potência, duas bobinas magneticamente acopladas e um recetor de potência. Um desenvolvimento cuidadoso das bobinas, nomeadamente a escolha de materiais e geometria, deve ser levado a cabo para reduzir as perdas. No entanto, de forma a otimizar a eficiência do sistema em geral, outros componentes devem ser desenvolvidos tendo tal em conta. Assim, o principal objetivo deste trabalho é otimizar o transmissor de potência. A aplicação alvo consiste no carregamento sem fios de veículos submarinos autónomos (AUVs), estando uma bobina no interior do AUV e a outra bobina no interior da estação de atracagem, ambas submersas no oceano.

A solução proposta para o transmissor desenvolvido consiste num inversor em ponte completa a operar em classe DE, de modo a reduzir as perdas por comutação. São utilizados condensadores em série com ambas as bobinas, sendo a frequência de operação estabelecida ligeiramente acima da ressonância de forma a fazer uso da impedância indutiva resultante. Os resultados obtidos a partir de testes feitos com uma configuração que representava o AUV e a estação de atracagem num ambiente de água salgada demonstram uma melhoria o desempenho do transmissor em termos de eficiência, tendo os testes revelado eficiências do inversor a atingir 99.5% e eficiências totais do sistema a atingir 86%, tendo em diferentes condições a potência de saída máxima testada atingido valores em torno de 120W.

Abstract

The present work addresses the optimisation of a power transfer system to achieve wireless power link in underwater conditions. Underwater wireless power transfer (WPT) in ocean is critically limited by the energy dissipated in salt water. This conductive medium is responsible for most of the power losses, significantly increasing with the distance between power transmitter (Tx) and receiver (Rx).

A typical inductive resonant system for WPT comprises a power Tx, two coupling coils and a Rx. A careful design of the coil pair with proper geometry and adequate choice of materials can effectively help in reducing these losses, but only up to some degree. However, in order to optimise the overall system efficiency, other components should be developed with that concern. Thus, the main goal of this thesis is to optimise the underwater power Tx. The target application is to develop a wireless charging solution to apply in autonomous underwater vehicles (AUVs) with one coil in the AUV and the other coil in a docking station, both underwater in the ocean.

The proposed solution for the developed Tx is a full bridge power inverter operating in class-DE, reducing the switching losses. Series capacitors are used both at the primary and secondary sides of the coupled coils, but with the operating frequency defined above resonance making use of the resulting inductive impedance. The results obtained from tests with a setup representing the AUV and the docking station with a salted water medium shown an increased Tx performance in terms of efficiency, revealing that the inverter efficiency could reach 99.5% and total system efficiency 86%. In different conditions, a maximum transmitted power of around 120W was reached.

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Miguel Silva

*“When wireless is perfectly applied the whole earth will be converted into a huge brain,
which in fact it is, all things being particles of a real and rhythmic whole.”*

Nikola Tesla, 1926

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Abbreviations and Symbols

AC	Alternated Current
AUV	Autonomous Underwater Vehicle
AWG	American Wire Gauge
C	Capacitance
DC	Direct Current
ESR	Equivalent Series Resistance
FIFO	First In First Out
L	Inductance
LDO	Low-Dropout Regulator
MPEP	Maximum Power-Efficiency Point
MPEPT	Maximum Power-Efficiency Point Tracking
OPAMP	Operational Amplifier
PBCLK	Peripheral Bus Clock
PI	Proportional Integral
PP	Parallel-Parallel
PS	Parallel-Series
PSS	Periodic Steady State
PWM	Pulse Width Modulation
R	Resistance
Rx	Receiver
SP	Series-Parallel
SPI	Serial Peripheral Interface
SPLL	System Phase Locked Loop
SS	Series-Series
Tx	Transmitter
UART	Universal Asynchronous Receiver Transmitter
USB	Universal Serial Bus
WPT	Wireless Power Transfer
ZVS	Zero Voltage Switching
ZVDS	Zero Voltage Derivative Switching

Chapter 1

Introduction

1.1 Context

Wireless power transfer (WPT) dates back to 1899, when Nikola Tesla first explored inductive coupling by means of resonant transformers, leading to what we nowadays know as Tesla coils [1]. In fact, Tesla has a major role in today's electrical reality, the use of alternated current (AC) in electrical distribution is one of his legacies, and also the invention of the three-phase induction motor, where the stator is the transmitter (Tx) and the rotor is the receiver (Rx), basically forming a transformer. Nowadays, there are several studies motivated by the historic experiments with the Tesla coils, which were originally used as a high-voltage high-frequency power source to generate electric sparks [2]. WPT is a related concept that recently started to be applied in an increasing number of emerging applications and is driving new interesting developments in several fields, some of which in consumer products used in a daily basis

Nowadays, the utilisation of wireless charging is becoming inevitable as the expansion of mobile devices where a battery must be present is fairly increasing [2]. The autonomy of these devices is one of the main problems to be tackled, so the development of strategies that allow for extended autonomy and, consequently, extended working periods is crucial. WPT holds a major role in this and represents a fundamental driver in the growing world of mobile electrical devices.

One of the most promising applications regards the wireless power charging of electrical vehicles due to their rapid growth as preferred in relation to traditional internal combustion engine vehicles. For that reason, being the battery capacity the main limitation in electrical vehicles, WPT like the one shown in Figure 1.1 represents a solution for making their use more efficient and extending their operation autonomy. As such, being inductive power transfer the most commonly used technology, there are several different ways of employing it to charge electrical vehicles, from stationary wireless charging, where the vehicle is parked in a way that a coil inside the vehicle is aligned to a coil outside the vehicle (often in the ground), to road powered wireless charging, where the vehicle is receiving electrical power as it moves through the road, which is done through rails on the road [2].

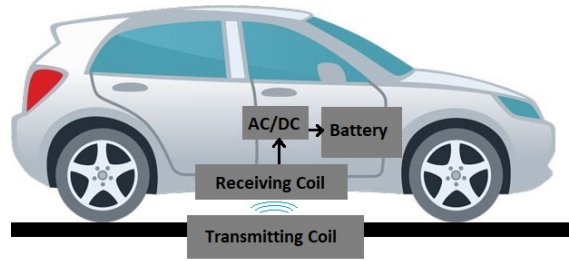


Figure 1.1: Electrical vehicle wireless charging

One different interesting application of wireless power transfer relies on the Maglev train technologies, which uses permanent magnets or electromagnets as means of suspension and propulsion through the magnetic field forces, offering great advantages such as reduction in maintenance through the elimination of wheels and tracks, reduction of noises and vibrations, prevents slipping and sliding, and others. In these vehicles, there is a power transfer from the ground for levitation, propulsion (as shown in Figure 1.2), electrical equipment powering and battery charging, which at high speeds is performed through a contactless linear generator that derives power from the travelling electromagnetic field. With the vehicle in motion, the levitation and propulsion coils generate a varying magnetic flux and the generator coils receive that varying flux that crosses the airgap [3].

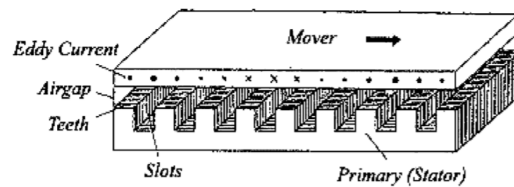


Figure 1.2: Maglev's linear induction motor [3]

Finally, the importance of such technology becomes evident in the underwater context, especially for charging autonomous underwater vehicles (AUVs). AUVs are an important component of underwater wireless networks and have their contribution and usefulness significantly increased when wireless charging is deployed. This provides them with higher capacity for long operations and reduces human intervention. One of the most common solutions for AUV wireless charging consists in a docking station as the one in Figure 1.3. This work will regard the AUV wireless charging application, which will further be described in detail. Other application for wireless power transfer in the underwater context concerns the charging of underwater sensors that can be present in the bottom of the ocean.

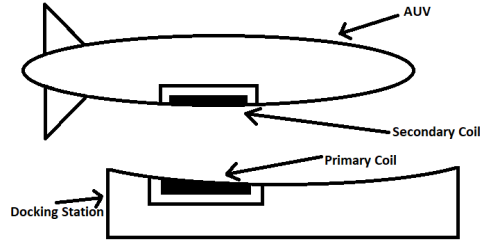


Figure 1.3: Docking station for AUV

1.2 Motivation

The deployment of underwater technologies has been receiving increased attention in distinct fields, for instance with the use of AUVs for monitoring deep-water rivers, interaction with underwater sensor networks as depicted in Figure 1.4, ocean floor mapping for exploration of minerals, oil/gas, etc [4]. In many of these applications, power might be the most limiting factor for success [5], potentially not taking the most out of the vehicle's autonomous capability [6], especially when high endurance is required. Nonetheless, to avoid any human intervention and since contact power transfer requires the design of a complex sealing structure [7], wireless powering approaches are required to transfer energy from and to the AUV, either to recharge it or to deliver energy to underwater deployed sensors, reducing the cost to replace the batteries, increasing reliability and sustainability [8].

Inductive power transfer has the same working principle of a transformer, except that the primary side is not fixed in relation to the secondary side, which leads to leaking inductance and increases power loss [9]. In addition, being this power transfer happening in seawater, there is an increasing problem concerning the medium conductivity [10]. Resonant inductive power transfer differs from the traditional inductive power transfer by employing resonant compensation networks, which minimise the effect of leakage inductance [7], allowing a higher power transfer [11]. This is the most used method for underwater applications, being more suitable than other methods like acoustic signals power transfer, which requires high precision components and are more expensive, or beamed power (lasers, microwaves, etc.), which show a large attenuation underwater [8].

The above described factors cause a decreased power transfer performance and lead to a focus on the optimisation of this process by researchers, with the goal of achieving the highest efficiency possible, to make this a deployable method.

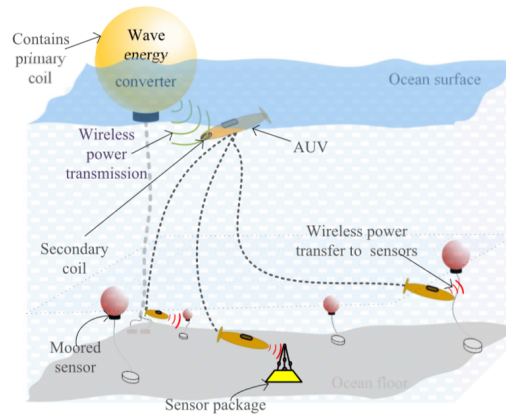


Figure 1.4: Underwater wireless network [12]

1.3 Objectives

The present work addresses the optimisation of a power transfer system to achieve wireless power link in underwater conditions. The system comprises a power transmitter, coupling coils and a power receiver. The basic overall systems are usually composed by the blocks depicted in the diagram of Figure 1.5.

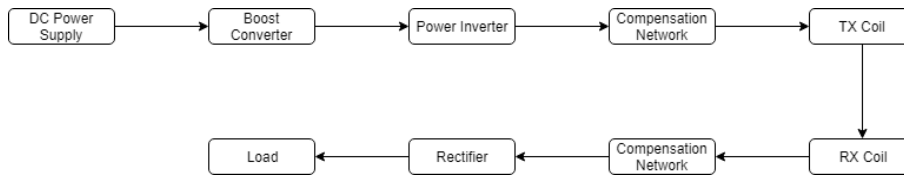


Figure 1.5: System block diagram

The main goal of this thesis is to optimise the hardware/software, realisation of the underwater powering system, designing and testing in a real-life scenario. That being said, the prime objectives of this work are:

- Control of a transmitter system, i.e. a full bridge transmitter system capable of monitor all system parameters, such as current sensing, voltage sensing, switching waveform monitoring, overload system protection.
- Design and development of a class-DE full bridge power inverter with improved efficiency.
- Design of a suitable control approach of the system to keep the power transfer efficiency at a maximum.
- Complete system integration and validation in a practical scenario.

1.4 Document Structure

The document is organised as follows.

Chapter 2 reports the scientific literature regarding different parts of WPT. After a presentation of the background on the fundamental aspects of inductive power transfer and its models follows the review of important components and characteristics of these systems. Among these are the coil design, losses, compensation networks, loads and control, being addressed with state-of-the-art referenced works.

In chapter 3, the coupled coils system to be used in the implementation is fully characterised using one of the theoretical models, the T-model. Further, with the obtained characterisation, the coupled coils system is studied relative to its operation, namely the efficiency and power capability depending on different loads and frequencies.

Next, chapter 4 presents a theoretical approach for system optimisation and first the class-DE power inverter topology is presented, analysed and simulated. Following, a new approach to the WPT system is taken and that power inverter topology is applied to it, with a theoretical analysis and consequent simulations.

Further, in chapter 5 the practical implementation of the previous chapter and the whole system realisation are presented, with hardware and software descriptions, tests and results discussions and the elaboration of a use case based on the results.

Following, in chapter 6 a control approach is presented with its architecture and simulation implementation, complemented with the respective obtained results.

Finally, chapter 7 describes the satisfaction of objectives and presents the future work to be carried.

Chapter 2

Wireless Power Transfer Literature Review

2.1 Background on Inductive Power Transfer

Equations (2.1) and (2.2) show Ampere's law and Faraday's law, respectively, which are the fundamental theories in which wireless power transfer is based [2], being important to analyse them in order to understand how the power flows between two coupled coils.

$$\nabla \times \mathbf{H} = \mathbf{J} \quad (2.1)$$

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \quad (2.2)$$

According to the differential form of Ampere's law, a time varying magnetic flux will be created by a time varying current in the Tx coil which, according to Faraday's law, will create an induced voltage in the Rx coil [2]. Consequently, the opposite effect will occur, from the Rx coil to the Tx coil, if the Rx coil has the possibility of conducting current through a closed circuit. Thus, a power flow between the two coupled coils is established.

Analysing the power flow mechanism between the two coils it is easy to conclude that the inductive power transfer is basically the same as a transformer, except for the high leakage inductances caused by the low coupling between the coils [2], which will further be approached. For this reason, resonant inductive power transfer is often deployed, as a way to tackle the reactance of these leakage inductances, improving the power factor and increasing the load power [2, 13–15].

2.1.1 Coupling Coefficient

The coupling coefficient can be viewed as a measure of how much of the Tx coil flux is used in the inductive link with Rx [8], and may be affected by a number of physical factors, such as the distance between the coils and their misalignment, as well as coil design factors, which will further

analysed. The coupling coefficient describes the relation between the mutual inductance and the Tx and Rx coils inductances, as can be seen in (2.3).

$$k = \frac{L_m}{\sqrt{L_p L_s}} \quad (2.3)$$

Furthermore, the coupling factor is well known to be related with the efficiency [1] and output power characteristics [9] of the wireless power transfer, being one of the prime characteristics found on the literature to be optimised. Equation (2.4) shows the relationship between the coupling coefficient and the efficiency [4].

$$\eta = \frac{k^2 Q_{Tx} Q_{Rx}}{\left(1 + \sqrt{(1 + k^2 Q_{Tx} Q_{Rx})}\right)^2} \quad (2.4)$$

However, in some cases, in a situation where the separation between the two coils is shorter than it was designed for [1], a too high coupling factor could make the system operate in the over coupled region. Even though the efficiency would be maximised at the resonant frequency, the output power peak at the same frequency observed in the under coupled region, in its frequency characteristics, would diverge and have two different peaks away from the resonant frequency in the over coupled region [1, 16]. This phenomenon is called in literature as frequency splitting. In the under coupled region, a higher coupling coefficient is related to a higher power transfer, and the opposite is observed in the over coupled region, at the same frequency, due to the frequency splitting [16].

As it can be seen in (2.4), the quality factor for reactive components has a direct influence in the system efficiency, being an important parameter in inductive wireless power transfer analysis and design. It describes the ratio of stored energy and dissipated energy of an inductor [17], as can be seen in (2.5). The quality factor of the resonator is typically limited by the quality factor of the coil [18].

$$Q_{coil} = \frac{\omega L_{coil}}{R_{coil}} \quad (2.5)$$

In addition, the quality factor is useful to evaluate the effect of higher order harmonics, since a low quality factor is characterised by responding to higher harmonics [19], whereas a higher quality factor has a more negligible response to higher harmonics [20]. Thus, a high quality factor might be used to compress the over coupled region and extend the under coupled region [16]. This value may be decreased in situations where high losses through Eddy currents are verified, like in the case where, for example, the coils are close to a metal shell [9, 17, 21], Eddy currents are created due to a varying magnetic field actuating over a conductive material. Another situation that could decrease its value is in the case where the coils are surrounded by salted water since it increases the permittivity of the medium, causing a lower self-resonance frequency, making the reactive and resistive part of the impedance deviate at a lower frequency, resulting in a lower optimal frequency of the quality factor [14, 22], as can be seen in Fig. 2.1.

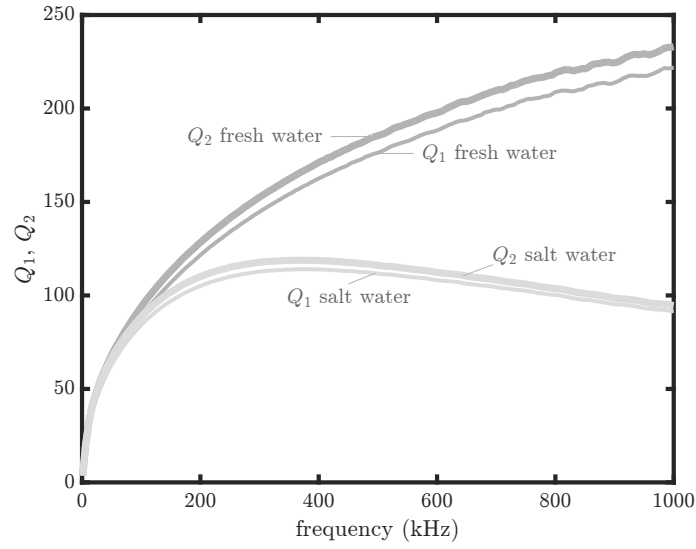


Figure 2.1: Measurement of the Q in two coils in fresh water and salted water [22]

2.1.2 Inductive Power Transfer Modelling

In order to accurately analyse and simulate inductive power transfer systems both in stationary and transient state, analytical models have been defined [2].

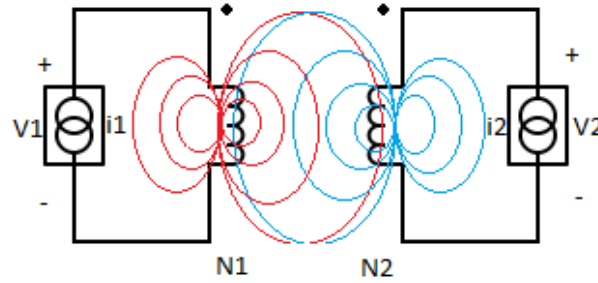


Figure 2.2: Coupled coils

First, the coupled coils system in Figure 2.2 must be analysed, where:

$$v_1 = N_1 \left(\frac{d\phi_{11}}{dt} + \frac{d\phi_{12}}{dt} - \frac{d\phi_{21}}{dt} \right) = \frac{L_{l1} di_1}{dt} + \left(\frac{L_{12} di_1}{dt} - \frac{L_{21} di_2}{n} \right) \quad (2.6)$$

$$V_2 = -N_2 \left(\frac{d\phi_{22}}{dt} + \frac{d\phi_{21}}{dt} - \frac{d\phi_{12}}{dt} \right) = -\frac{L_{l2} di_2}{dt} - \left(\frac{L_{21} di_2}{dt} - \frac{nL_{12} di_1}{dt} \right) \quad (2.7)$$

Because:

$$\phi = \int_s \mathbf{B} d\mathbf{S} \quad (2.8)$$

$$\mathbf{B} = \mu \mathbf{H} \quad (2.9)$$

$$\oint_l \mathbf{H} d\mathbf{l} = Ni \quad (2.10)$$

It is known that only a part of the magnetic flux is shared between the two coils, depending on the coupling coefficient k . L_{l1} and L_{l2} are the leakage inductances and L_{12} and L_{21} are the inductances of primary and secondary sides, respectively, sharing magnetic flux:

$$L_{l1} = \frac{N_1 \phi_{11}}{i_1}, \quad L_{l1} = (1 - k) L_1 \quad (2.11)$$

$$L_{l2} = \frac{N_2 \phi_{22}}{i_2}, \quad L_{l2} = (1 - k) L_2 \quad (2.12)$$

$$L_{12} = \frac{N_1 \phi_{12}}{i_1} \quad (2.13)$$

$$L_{21} = \frac{N_2 \phi_{21}}{i_2} \quad (2.14)$$

Having defined the basic equations of a two coupled coils system, a complete transformer model can be created. In this model, as can be seen in Figure 2.3, the two leakage inductances, the magnetising inductance and transformation ratio are defined, making it an useful choice for designing the coils of the system and for circuit-based analyses [2].

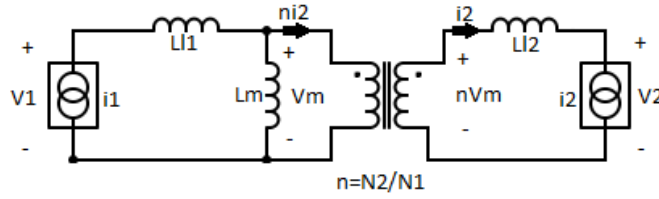


Figure 2.3: Transformer model

From the previous analysis, the following equations can be derived, where L_m is the magnetising inductance that is used in the coupling and depends directly on the coupling coefficient, as shown in (2.3), and R_m is the magnetic resistance of the magnetic coupling coil:

$$L_{12} = \frac{N_1 \phi_{12}}{i_1} = \frac{N_1 \phi_m}{i_1}, \quad i_2 = 0, \quad L_{12} = \frac{N_1}{i_1} \frac{N_1 I_1}{R_m} = \frac{N_{12}}{R_m} \quad (2.15)$$

$$L_{21} = \frac{N_2 \phi_{21}}{i_2} = \frac{N_2 \phi_m}{-i_2}, \quad i_1 = 0, \quad L_{21} = \frac{N_2}{i_2} \frac{N_2 i_2}{R_m} = \frac{N_{22}}{R_m} \quad (2.16)$$

$$L_{21} = n^2 L_{12} = n^2 L_m, \quad L_{12} = L_m \quad (2.17)$$

$$v_1 = \frac{L_{l1} di_1}{dt} + \left(\frac{L_{12} di_1}{dt} - \frac{L_{21} di_2}{n} \right) = \frac{L_{l1} di_1}{dt} + L_m \left(\frac{di_1}{dt} - \frac{n di_2}{dt} \right) = \frac{L_{l1} di_1}{dt} + v_m \quad (2.18)$$

$$v_2 = -\frac{L_{l2} di_2}{dt} - \left(\frac{L_{21} di_2}{dt} - \frac{n L_{12} di_1}{dt} \right) = \left(\frac{n L_m di_1}{dt} - \frac{n^2 L_m di_2}{dt} \right) - \frac{L_{l2} di_2}{dt} = n v_m - \frac{L_{l2} di_2}{dt} \quad (2.19)$$

$$v_m = L_m \left(\frac{di_1}{dt} - \frac{ndi_2}{dt} \right) = \frac{L_m di_m}{dt}, \quad i_m = i_1 - ni_2 \quad (2.20)$$

A detailed transformer model including components like resistances for conduction and core loss and parasitic capacitances as can be seen in Figure 2.4 could also be built for a more accurate analysis [2].

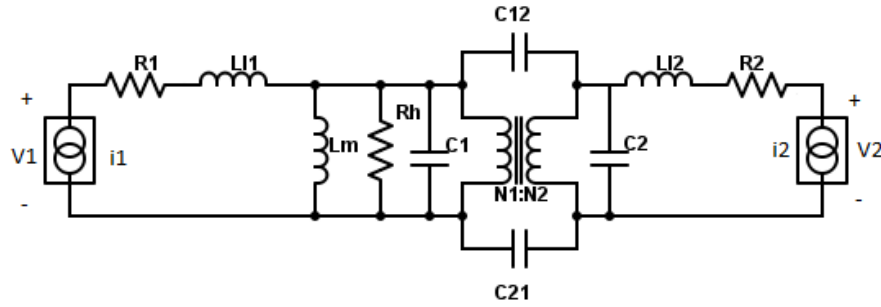


Figure 2.4: Detailed transformer model

The next model to be defined is the M-model, which has the advantage of simplicity and preferred for equation-based analyses, not being suited for circuit based analyses though [2].

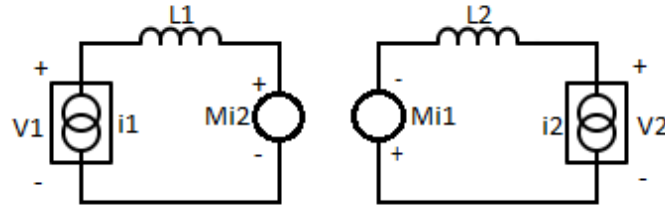


Figure 2.5: M-model

$$v_1 = \frac{L_1 di_1}{dt} - \frac{M di_2}{dt}, \quad v_2 = \frac{M di_1}{dt} - \frac{L_2 di_2}{dt} \quad (2.21)$$

$$v_1 = \frac{L_{l1} di_1}{dt} + L_m \left(\frac{di_1}{dt} - \frac{ndi_2}{dt} \right) = \frac{(L_{l1} + L_m) di_1}{dt} - \frac{nL_m di_2}{dt} = \frac{L_1 di_1}{dt} + \frac{M di_2}{dt} \quad (2.22)$$

$$v_2 = \left(\frac{nL_m di_1}{dt} - \frac{n^2 L_m di_2}{dt} \right) - \frac{L_{l2} di_2}{dt} = \frac{nL_m di_1}{dt} - \frac{(n^2 L_m + L_{l2}) di_2}{dt} = \frac{M di_1}{dt} - \frac{L_2 di_2}{dt} \quad (2.23)$$

$$M = nL_m, \quad L_1 = L_{l1} + L_m, \quad L_2 = L_{l2} + n^2 L_m \quad (2.24)$$

As it can be seen in Figure 2.5, there are three parameters describing the model. L_1 and L_2 are the inductances viewed from both sides and M is the mutual inductance, obtained through the transformation ratio n and the magnetising inductance L_m [2].

Finally, it is possible to define the T-model from Figure 2.6, which derives from the M-model and merges the simplicity of the M-model and the possibility of circuit-based analysis of the transformer model [2]. This model, however, does not display some parameters, such as the transformation ratio and the coupling factor and, besides that, it is not an exact equivalent circuit because there is no isolation [2]. Thus, from the equations of the M-model, the following equations are defined:

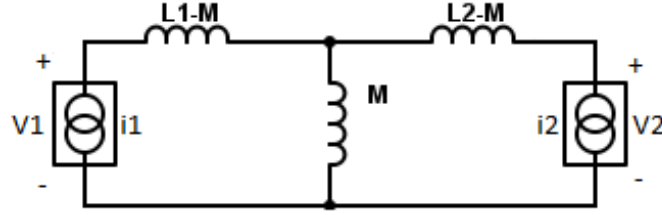


Figure 2.6: T-model

$$v_1 = \frac{L_1 di_1}{dt} - \frac{M di_2}{dt} = \frac{(L_1 - M) di_1}{dt} + M \left(\frac{di_1}{dt} - \frac{di_2}{dt} \right) \quad (2.25)$$

$$v_2 = \frac{M di_1}{dt} - \frac{L_2 di_2}{dt} = -\frac{(L_2 - M) di_2}{dt} + M \left(\frac{di_1}{dt} - \frac{di_2}{dt} \right) \quad (2.26)$$

2.2 State of the Art

2.2.1 Power Loss

Knowing that one of the major goals is to achieve high efficiencies to make such systems economically and sustainably valuable, when analysing and designing an inductive power transfer system, it is of prime importance to analyse and quantify the losses and to determine how such losses occur and how to minimise or prevent them. This is an especially critical matter when the surrounding medium is seawater. Due to its high electrical conductivity, when compared to air or distilled water, the losses will be considerably increased [1], as it will further be explained, causing the efficiency to drop significantly.

An inductive power transfer system presents three major types of power loss, namely semiconductor loss in the electronic circuits, copper loss in the coils and Eddy current loss in the transmission medium [9].

Most of the inductive power transfer systems are comprised of an inverter bridge on the primary side for generation of a high frequency wave form to magnetise the coupling coils and a rectifier bridge on the secondary side to convert the signal to a continuous signal [15] that can be used to charge a battery, for example. The inverter bridge and rectifier bridge losses are described in (2.27) and (2.28), where T_s is the voltage rising and falling times.

$$P = 2V_g i_g T_s f \quad (2.27)$$

$$P = 2i_g^2 R_{ds_{on}} + \left(4 \frac{\sqrt{2}}{\pi}\right) i_L V_{drop} \quad (2.28)$$

To minimise these semiconductor losses, the inverter and the rectifier are often soft-switched, which means that the semiconductor switches between on and off when either the current or the voltage is close to zero [2]. Other types of power devices could be present, like a DC/DC converter, which would increase the total switching losses.

In what copper loss in the coil windings concerns, those happen due to the small but significant resistive characteristic of the windings and the typical coil resistance values registered in literature experiments range from 0.15 Ω to 1.3 Ω . The values of this coil resistivity present an approximately linear relation with the system efficiency [9]. The resistance of a coil in seawater can be divided into DC resistance R_{dc} , AC resistance R_{ac} and radiation resistance R_{rad} [23]. Experiments in [12] and [23] show that R_{ac} increases with frequency but is the same in air and seawater. Furthermore, this value depends on the coil configuration and quality factor of the wire [12]. In addition, though R_{rad} has a very small value in air without increasing with frequency [12, 23], it has a different behaviour in seawater due to the conductive nature of seawater which allows the creation of electric currents induced by the magnetic field that surrounds the coil [12], causing the highest increase with frequency amongst all resistances [23]. This value increases exponentially when the frequency is greater than a determined value (200 kHz in a simulation from [12], where the coil radius is 10 cm, the wire radius is 0.4 mm and the coil had 20 turns).

Finally, medium losses should be accounted. As stated before, efficiency in seawater is decreased compared to the one in air, and this difference becomes more evident as the separation between the coupled coils increases, being that phenomenon caused by Eddy current losses in the more conductive medium [1]. Besides the increase in medium losses by increasing the distance between coils, there will also be an increase in medium losses with an increasing lateral misalignment between the coupled coils, with small losses for small lateral misalignments but rapidly increasing losses for bigger lateral misalignments [24]. Furthermore, medium losses are strongly dependent on the operating frequency, since a higher frequency leads to a higher induction in the conductive material, creating Eddy currents and increasing medium losses, which can be represented by an increased series resistance in the equivalent electrical model [4, 8, 19, 21]. However, lower frequencies decrease the coupling coefficient [8, 9]. For this reason, the typical frequency values used in literature for underwater applications are much smaller than those used for in air power transference and range from 19 kHz to 215 kHz. Some experiments found in literature show that the medium losses become much more critical above a certain frequency, being less significant at lower frequencies. For example, in [7] simulations show that medium losses are less significant at frequencies below 30 kHz.

2.2.2 Coil Design

Coil design has a major role in the project and development of an inductive power transfer system, since it will directly affect important system parameters such as the coupling coefficient, losses

and efficiency. Some different variables must be approached carefully when designing the coils for an inductive power transfer system, namely the used materials, the coil shape, the number of winding turns, the coil diameter and the coil inductance.

Relative to the materials used in the coils construction, which mainly regard the core and the windings, some are regularly found to be used in the literature experiments for delivering the best results.

One of those materials is ferrite, commonly used in the core due to its high magnetic permeability leading to a reduction of the magnetic leakage and an increased coupling efficiency [25], thanks to a bigger magnetic flux concentration in the core. Besides that, Eddy currents are minimised, especially if the coils are surrounded with ferrite [1]. In addition, it is valuable to reduce the use of other conductive materials in the construction near the coils, and some materials like plastic and foam could be used [1]. However, some studies suggest that coreless coils could be used to avoid hysteresis loss on the core [9,26], which represents a trade-off and results in a lower coupling coefficient due a higher leakage flux [9, 13, 14].

Regarding the windings, the most commonly used material for an increased performance is Litz wire, whose diameter in literature experiments typically range from 0.6 mm to 2.6 mm. Simulation results in [12] suggest that the self-inductance and mutual inductance is higher for smaller coil wire diameters. The reason why Litz wire is often the preferred material for the windings is because the current is equally distributed between the wire strands and the alternating current resistance R_{ac} that results from the skin effect at high frequency and proximity effect is minimised [10,26]. As discussed earlier, besides reducing the copper losses, reducing the resistance of the coil will allow Litz wire to improve the quality factor [14]. Additionally, the windings can be actively or naturally cooled to maintain the resistance values within their operating temperature range [10].

Concerning the coil shape, that is another variable of great relevance on the coil design and will largely depend on the application, from the docking station type (it can, for example, be funnelled like in Figure 2.7 or half pipe, etc.) to the AUV appliance and where and how will the coils be located on the AUV. Having that said, the coil shape can vary between planar or nonplanar, it may contain or not contain a core (as discussed previously) and it can have several different winding arrangements like cylindrical, helical, spiral, etc.

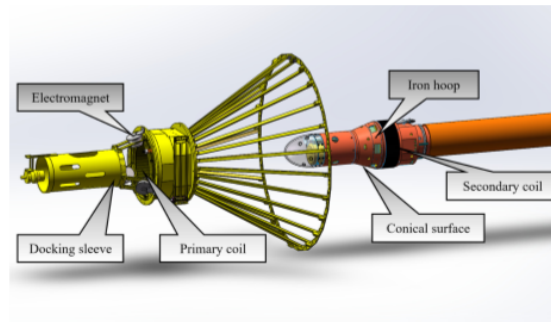


Figure 2.7: Funnelled type docking station [27]

Comparing the helical and spiral coils, some experiments show that the coupling coefficient and mutual inductance between helical coils is much lower than in spiral coils [8, 12], which is also evident when the distance between both coils increases, where the efficiency in the spiral coils is higher [12]. In [12], simulations with helical and spiral coils both with 20 turns, 10 cm radius and 0.4 mm wire radius show that the coupling coefficient at 1 cm separation is 0.3 in the helical coil and almost 0.8 in a spiral coil.

The next coil design variable to be analysed is the number of turns. The typical number of turns used in literature ranges from 7 to 50 turns and experiments show that the number of turns affects transferred power and efficiency, having an optimum point [13]. Furthermore, simulations in [8] with number of turns ranging between 2 and 20 show that increasing the number of turns improves the coupling coefficient, with 2 turns spiral coil corresponding to a 0.22 coupling coefficient and 20 turns spiral coil corresponding to a 0.78 coupling coefficient. The coupling coefficient values are only demonstrative, since they will vary greatly with many design characteristics.

Regarding the coil diameter, it typically ranges from 6 cm to 35 cm in literature, and a larger diameter coil may transfer power at longer distances since the coupling coefficient is dependent on the relation between coil diameter and distance between coils. Besides, [12] suggests that to achieve maximum power transfer, the diameter of the Tx coil should be made bigger than the one of the Rx coil, which could also allow to mitigate bigger misalignments between the coils during the docking process [4].

Finally, the typical coil inductance values used in literature range from 10 μH to 1000 μH , and simulations results from [9] obtained with resonant state maintained by adjusting the resonant capacitance suggest that the self-inductance of coils does not have a strong effect on the efficiency, as long as their values are not too low, whereas the output power is closely related to the ratio of the Rx to Tx self-inductances, and for a bigger ratio the output power was higher.

2.2.3 Resonant Compensation Networks

As mentioned before, due to the high leakage inductances caused by the low coupling between the coils [2], resonant circuits are often used, as a way to tackle the reactance of these leakage inductances, improving the power factor and increasing the load power [2, 13–15]. This way, the power that would be lost on the unused magnetic field is stored in a resonant circuit. As such, two coils can exchange power with high efficiency, even though their magnetic fields are not tightly coupled, as long as both coils have the same resonant frequency and are operating in that frequency [2]. That frequency is calculated as in (2.29) that shows the frequency at which the coil impedance matches the capacitor impedance.

$$f = \frac{1}{2\pi\sqrt{LC}} \quad (2.29)$$

The compensation circuits may differ in the number of used capacitors and inductors as well as their configuration, whereas the most common compensation circuits only use two capacitors [2], one on the Tx side and the other on the Rx side. The most commonly studied compensation circuits

in literature for inductive power transfer applications can be seen in Figure 2.8 and are classified relative to the arrangement of the capacitor on the primary and secondary sides, which can be in series or in parallel with the coil: SS (series-series), SP (series-parallel), PS (parallel series) and PP (parallel-parallel).

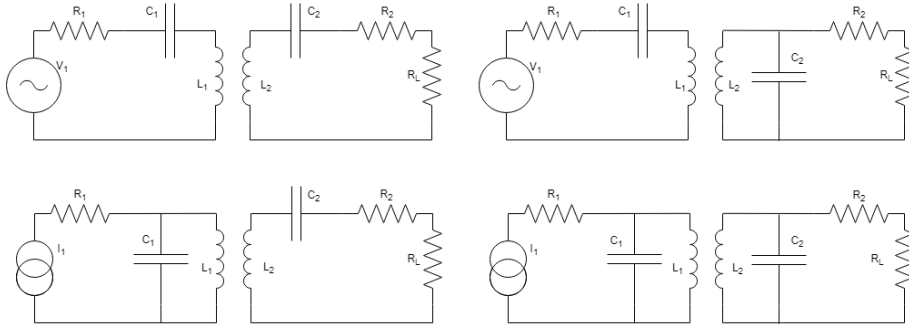


Figure 2.8: Most common compensation circuits

The different arrangements present some advantages and disadvantages relative to each other, and a careful analysis should be made in order to select the most suitable compensation circuit for a specific application.

The efficiency equation for SS, SP, PS and PP can be seen from (2.33) to (2.36), respectively, being dependent of variables like the angular frequency ω , the mutual inductance L_m , the capacitances C_1 and C_2 , the loss resistances R_1 and R_2 and the load resistance R_L [28].

In the SS structure, as seen in the simplification performed in Figure 2.9, the resistance R_{2eq} represents the equivalent resistance of the secondary reflected to the primary side, which is [4]:

$$R_{2eq} = \frac{(\omega_0 L_m)^2}{(R_2 + R_L)} \quad (2.30)$$

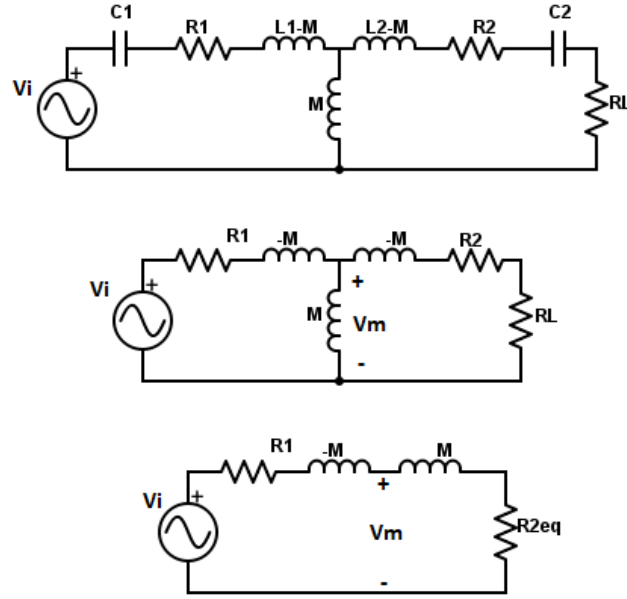


Figure 2.9: Secondary impedance reflected to the primary side (third circuit) after simplification of the T-model circuit (first circuit) at resonance frequency (second circuit)

From (2.30), it is possible to determine the input and output power:

$$P_i = \frac{1}{2} \frac{R_2 + R_L}{R_1 (R_2 + R_L) + (\omega_0 L_m)^2} V_i^2 \quad (2.31)$$

$$P_o = \frac{1}{2} \frac{(\omega_0 L_m)^2 R_L}{\left(R_1 (R_2 + R_L) + (\omega_0 L_m)^2 \right)^2} V_i^2 \quad (2.32)$$

leads to the efficiency equation:

$$\eta_{SS} = \frac{P_o}{P_i} = \frac{(\omega_0 L_m)^2 R_L}{(R_2 + R_L) \left(R_1 (R_2 + R_L) + (\omega_0 L_m)^2 \right)} V_i^2 \quad (2.33)$$

with the same being performed with the remaining structures, obtaining the following efficiencies [28]:

$$\eta_{SP} = \frac{\omega^2 L_m^2 R_L}{\left(\left(R_1 + j\omega L_1 + \frac{1}{j\omega C_1} \right) \left(R_2 + j\omega L_2 + R_L // \frac{1}{j\omega C_2} \right) + \omega^2 L_m^2 \right)} \frac{1}{1 + j\omega C_2 R_L} \quad (2.34)$$

$$\eta_{PS} = \frac{\omega^2 L_m^2 R_L}{\left(\left(\left(R_1 + j\omega L_1 \right) // \frac{1}{j\omega C_1} \right) (R_2 + R_L) + \omega^2 L_m^2 \right) (R_2 + R_L)} \quad (2.35)$$

$$\eta_{PP} = \frac{\omega^2 L_m^2 R_L}{\left(\left((R_1 + j\omega L_1) // \frac{1}{j\omega C_1} \right) \left(R_2 + j\omega L_2 + R_L // \frac{1}{j\omega C_2} \right) + \omega^2 L_m^2 \right)} \cdot \frac{1}{1 + j\omega C_2 R_L} \cdot (R_2 + j\omega L_2 + R_L // \frac{1}{j\omega C_1}) \quad (2.36)$$

On a first circuit analysis, it can be concluded that in SS and in PS, the load current will be consistent with the coil current, whereas in SP and PP that does not apply [28]. Furthermore, the fact that PP and PS require that a current source is used and that they are deviated from the resonant state as the load resistance varies, causing a fluctuant output, make them less suitable for practical application, leaving us with SS and SP [28, 29].

Starting by analysing the SS structure, studies suggest that it is more suitable for close range, higher frequency and smaller load [6, 28, 30, 31]. As it can be seen in Figure 2.10, above a certain load resistance, the power efficiency decreases as the load increases [28]. In addition, as can be seen in Figure 2.12, though the power transfer is very low for low values of mutual inductance it increases with an increasing mutual inductance, having its maximum at the resonant frequency. At some point, however, if the mutual inductance keeps being increased that maximum at the resonant frequency does not verify, due to the frequency splitting phenomenon described before caused by the over coupling [30]. Figure 2.11 shows that the efficiency is clearly higher at the resonant frequency, having a poorer performance at different frequencies. There is a trade-off between efficiency and transferred power and when maximum power is demanded, even though the efficiency decreases with the load, this structure offers the highest power delivery [6]. Additionally, the SS structure has the advantage of not having its resonance affected by the load value [20]. Finally, Figure 2.13 shows that this topology has a large fluctuation in the output voltage as the load varies [29].

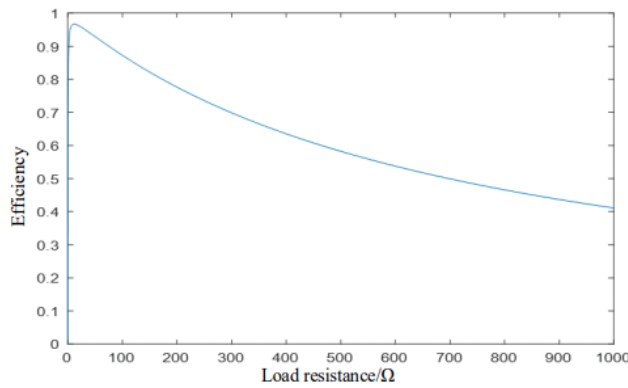


Figure 2.10: Simulation results for efficiency variation with load resistance with SS structure. Not in an underwater condition [28]

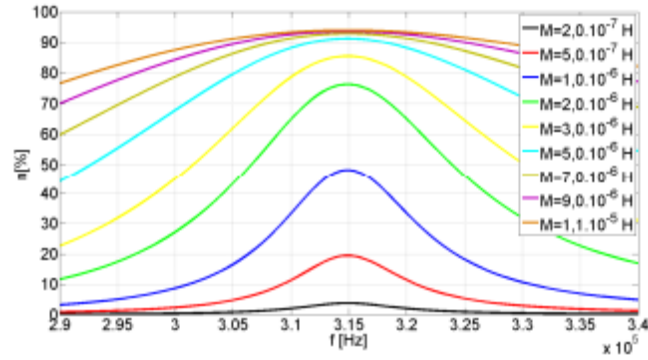


Figure 2.11: Simulation results for efficiency variation with frequency for various mutual inductances with SS structure. Not in an underwater condition [30]

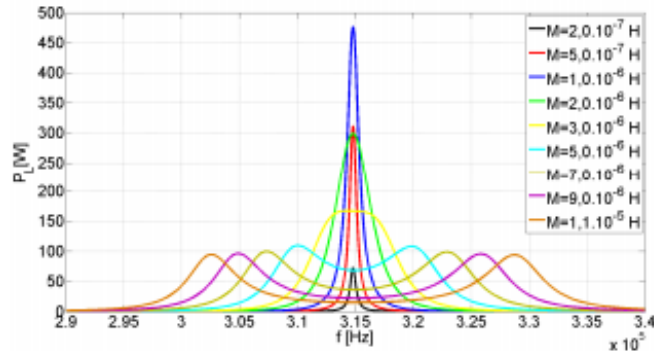


Figure 2.12: Simulation results for load power variation with frequency for various mutual inductances with SS structure. Not in an underwater condition [30]

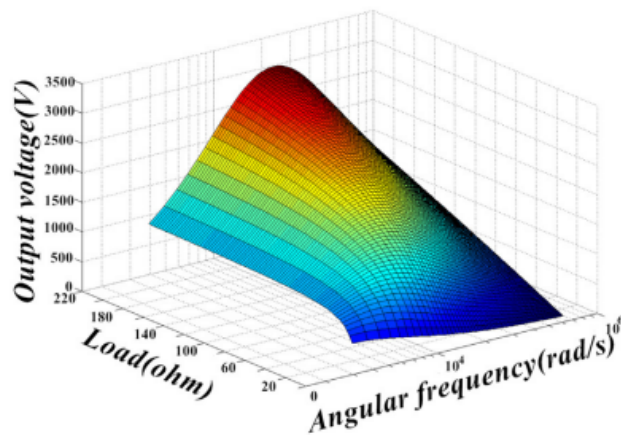


Figure 2.13: Simulation results for output voltage variation with load for various frequencies with SS structure [29]

Taking now a closer look to the SP structure, studies suggest that it is more suitable for lower

frequency, longer distances and larger load resistances [6, 28, 30, 31]. As it can be seen in Figure 2.14, the power transfer efficiency is optimised for higher loads, which is a great difference between the SS and the SP structures [28]. Relative to efficiency and transferred power, both increase with the rise of mutual inductance, being the transferred power shifted to frequencies higher than the resonant frequency [30], as can be confirmed in Figure 2.15 and Figure 2.16. Compared to the same results for SS structure, even though the efficiency in the SP structure shows a constant value for a wide range of frequencies, with the same mutual inductances, the SP structure reaches much inferior values for transferred power [30]. In addition, a poorer efficiency is verified in this structure [28, 30]. In terms of output voltage fluctuation with load variations, the SP structure keeps a more stable output voltage than the SS structure, as can be seen in Figure 2.17.

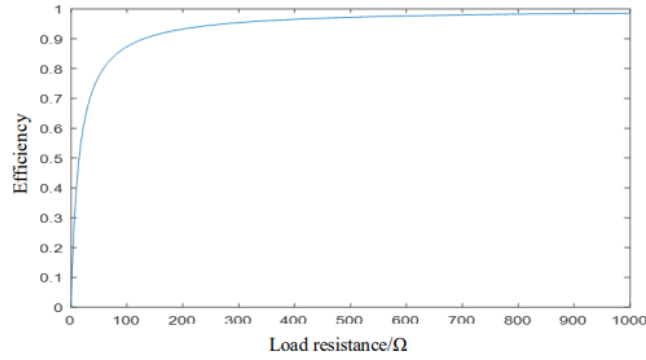


Figure 2.14: Simulation results for efficiency variation with load resistance with SP structure. Not in an underwater condition [28]

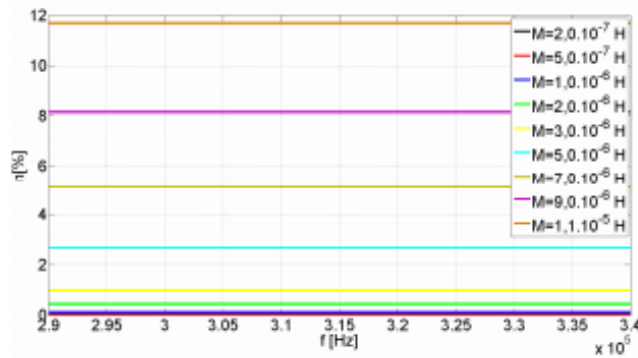


Figure 2.15: Simulation results for efficiency variation with frequency for various mutual inductances with SP structure. Not in an underwater condition [30]

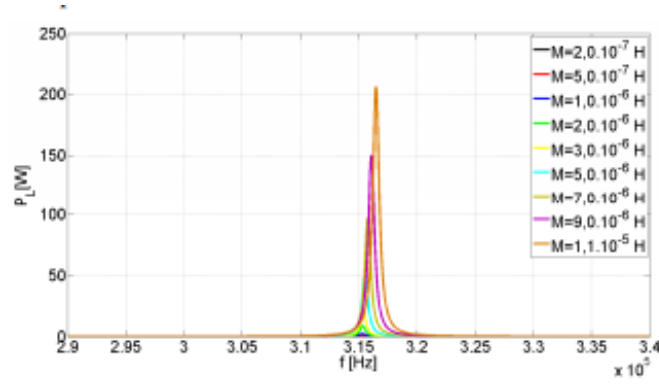


Figure 2.16: Simulation results for load power variation with frequency for various mutual inductances with SP structure. Not in an underwater condition [30]

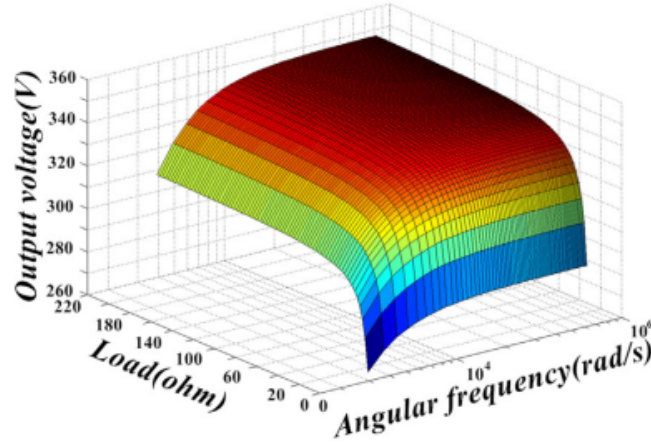


Figure 2.17: Simulation results for output voltage variation with load for various frequencies with SP structure [29]

The capacitor values typically used in literature can range from 10nF to 800nF. For compensation proposes, a variable capacitor could be used at the Tx side to set the resonant frequency to its precise value [12]. Furthermore, knowing that high operating frequencies may incur in high medium and switching losses, a larger capacitance and lower frequency can be used and still maintain the magnitude of the output power [9], as can be seen in Figure 2.18. The value of the Tx side capacitor is calculated differently for the SS structure and the SP structures whereas the Rx side capacitor may be calculated in the same way for all the topologies as depicted from (2.37) to (2.39), respectively [29].

$$C_{1ss} = \frac{1}{\omega_0^2 L_1} \quad (2.37)$$

$$C_{1sp} = \frac{L_2}{\omega_0^2 (L_1 L_2 - L_m^2)} \quad (2.38)$$

$$C_2 = \frac{1}{\omega^2 L} \quad (2.39)$$

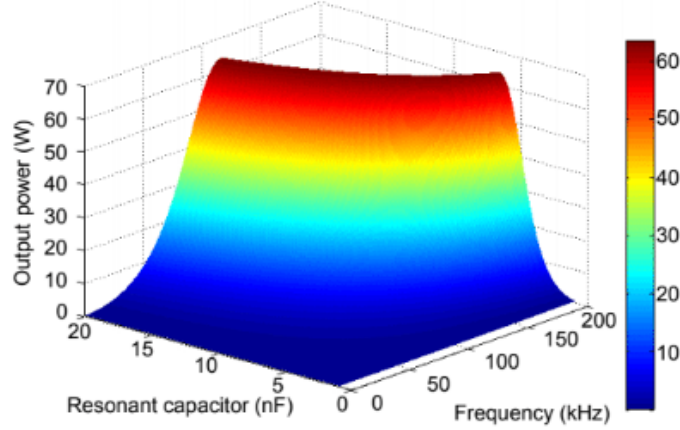


Figure 2.18: Simulation results for output voltage variation with load for various frequencies with SP structure [29]

2.2.4 Load

The typical values for underwater transferred power found in literature range from 30 W to 400 W. Regarding the loads, some studies suggest that, generally, bigger loads lead to slightly higher efficiency but slightly smaller output power [9], which, as seen before, does not verify when SS type compensation structures are used.

It is important to note that a voltage regulator can be used, usually after the rectifier, because during the charging process of a battery, as the state of charge changes, the equivalent load changes, possibly resulting in a fluctuant output if a voltage regulator is not employed [11, 15]. During these systems analyses, it is often assumed that the load is fixed in steady state. However, this assumption does not correctly translate to reality, since there is a variation in the coupling coefficient when the vehicles are approaching the docking station as well as when they are docked due to oscillations and misalignments [6]. In addition, having a variable impedance seen from the transmitter caused by some variables such as the relative position between the coils, unpredictable load variation demands, charging level dependent impedance, temperature-sensitive parameters, long-term capacity drops, component ageing [20], among other mechanical or electrical uncertainties, the resonance of the system will also be affected, being especially severe when there is a high quality factor and power can only be transferred in a narrow frequency bandwidth with almost no tuning capacity, degrading the power delivery as well as the system efficiency [4, 11].

2.2.5 Control

As it generally happens for most electronic power systems, when designing an inductive power transfer system, a special care should be given to the control system in order to maximise its

performance while preventing unsafe situations, either for the system or its users. Due to the previously mentioned parameter deviations in practical implementations, adaptive approaches should be used to keep the system in the desired operation when resonance or other parameters deviate from what was expected [11].

For this reason, inductive power transfer systems basic variables such as the battery voltage, the charging current and battery temperature should be monitored and the charging sequences should be controlled [21]. Additionally, the batteries consumption and charging historic could be recorded for battery management and charging control, abnormal charging could be alerted and the amount of battery charge could be controlled to avoid overcharging [21].

In terms of performance, the most important system variables to be controlled are the power transfer efficiency and the transferred power. Regarding the transferred power control, there are two main methods to perform it, namely the frequency control method and the voltage control method.

In the frequency control method, the goal is to modulate the operating frequency around the resonant frequency, which is where the transferred power peaks [32]. In this method, when the charging is initiated, the frequency should be set to a significantly far away value from the resonant frequency, in order to avoid unsafe situations. Then, the transferred power is slowly increased by making the frequency gradually closer to the resonant frequency, while the battery voltage and temperature maintain below their limit values, being the current monitored throughout the process and gradually made closer to its set point [32]. Care should be taken when systems have a high quality factor and, consequently, a very steep resonant peak, which adds the risk of making the system operate in unsafe power levels very rapidly with small frequency changes [32].

In the voltage control method, less commonly employed, knowing that near resonance the power has an almost linear behaviour with voltage variation, the transferred power is controlled by varying the Tx bridge or DC/DC converter voltage as required by the load, allowing a frequency independent power level control, and could be used in cases where the range of required power levels is too wide and cannot be reached by frequency variation alone being [32].

Finally, the most advantageous method would be to perform the power control by combining both of the previously described methods, and the flexibility and fine control capability of the frequency control method could be merged with the wider range of transferred power and flexibility for different loads and power demands shown by the voltage control method, providing the capability of rough tuning with the voltage control and fine tuning with frequency control [32].

For the power transfer efficiency, the goal is generally to maximise it. To achieve that, the maximum power-efficiency point (MPEP) in an efficiency to secondary voltage graph should be tracked (MPEPT) in order to always maintain the highest possible power transfer efficiency by deploying the optimal secondary voltage [12]. This could be observed in Figure 2.19.

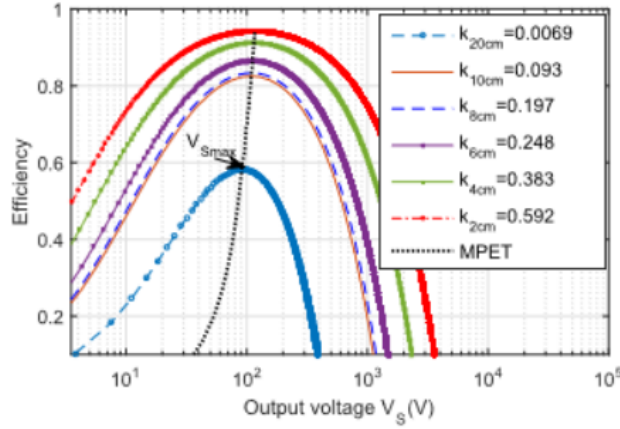


Figure 2.19: MPEP for various coupling coefficients [12]

This MPEP tracking is critical due to the dynamic behaviour of the power efficiency with load and coupling coefficient variations between the coupled coils, and the reason why this is done through the secondary voltage is because its maximising values correspond to a desirable load resistance satisfying the impedance matching condition [12]:

$$V_s = \frac{(\omega_0 L_m) R_L}{(R_L + R_s) (R_p R_s + (\omega_0 L_m)^2)} V_p \quad (2.40)$$

$$V_{smax} = \frac{\sqrt{\frac{R_s}{R_p}} \omega_0 k \sqrt{L_p L_s}}{\sqrt{R_p R_s} + \sqrt{R_p R_s + (\omega_0 k)^2 L_p L_s}} V_p \quad (2.41)$$

The implementation of control strategies commonly has two different variations for processing data for charging control. One option is to transfer battery data such as current, voltage and temperature from the Rx to the Tx, being the data processing completely done on the Tx side, whereas the other option is to make the processing on the Rx side and transmit the charging commands to the Tx side [32]. In order to transmit these and other information regarding situations such as no load, open circuit, short circuit, over voltage, under voltage, transient overshoot, inrush current and failure of communication, a reliable communication link between the Tx and the Rx must be ensured [2]. That being said, the communication link can be wirelessly realised using technologies such as optical communication, which is susceptible to water quality, acoustic communication, which has a small communication rate, and electromagnetic waves, which are capable of high communication rates at short distances [26]. However, even though there are some difficulties, inefficiencies and lack of robustness against abnormal conditions when doing it, the Tx and Rx could be controlled independently, without a communication link, and the data would ideally be estimated using the electrical quantities on the Tx side, making the systems more simple and of lower cost [2, 11]. For these data estimations, neural-network based machine learning could be employed to estimate the coupling coefficient [12]. To regulate the Rx power or voltage, the Tx must track the power at the load side, but when such information is not directly available

through a wireless communication link, one solution is to communicate using load modulation at the secondary side, which may be done with periodic load short-circuits, or by switching a resistance in parallel with the load, which relies on a change of the gain, or by switching a capacitor at the input pins of the rectifier, which provides a momentary deviation in the frequency response [11, 20]. When using load modulation, the data is sensed in the Tx side by monitoring the peak current through the primary coil, being the SS compensation structure more suitable to prevent the change of resonance frequency through data modulation [20].

Chapter 3

Wireless Power Transfer Coils Analysis

Before starting the study and development of the proposed solution, it is crucial that an analysis of the used coupled coils system shown in Fig. 3.1 is performed. These are 7 turn coils, with 2 twisted wires and a 16cm outer diameter in a structure resembling the periphery of a 20cm radius autonomous underwater vehicle. Both coils have the same number of turns, similar concavity, wire length, 14 AWG (American wire gauge) copper wire and other geometry parameters. The distance between coils is 4cm.

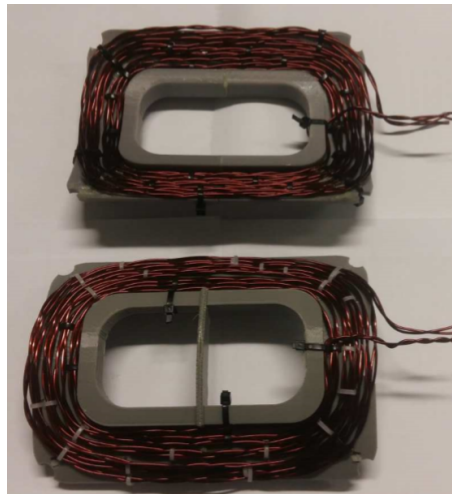


Figure 3.1: Used coils [33]

As such, in this chapter, a study of the coupled coils to be used in the experimental phase is performed from different points of view using previously collected characterisation data.

3.1 T-model

As presented before, one of the most commonly used models to study a system comprising two coupled coils is the T-model, as it can be seen in Fig. 3.2. Here, it can be seen that a mutual

resistance, R_m is added in series with the mutual inductance, L_m , to represent the resistive losses on the conductive medium.

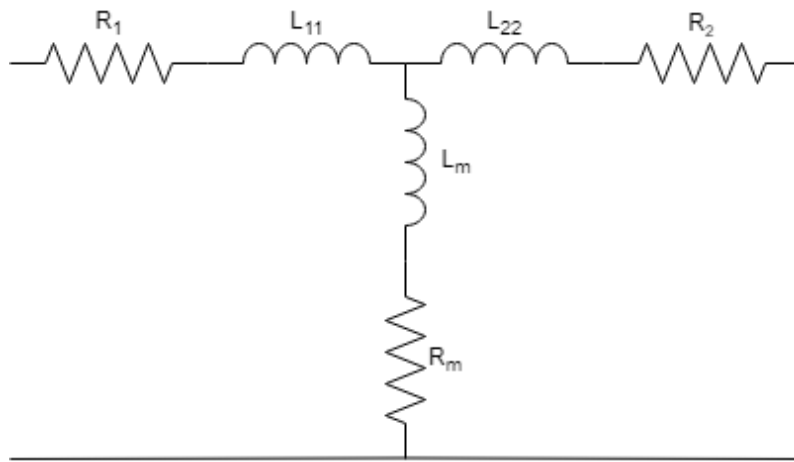


Figure 3.2: Used T-model circuit

In order to characterise the coupled coils system, an impedance meter must firstly be used to measure the impedance of the primary and secondary coils independently, with each coil in open circuit, imposing a null coupling coefficient. Then, to obtain the system input equivalent impedance, a measure from the primary side is performed, with the secondary side in short circuit, allowing to obtain the mutual inductance and resistance values and, consequently, the coupling coefficient value [22]. Such measurements were carried out prior to this thesis work in a salted water medium with a salt concentration of around 35 g/L by the investigation team responsible for this project, having been provided for the further presented development.

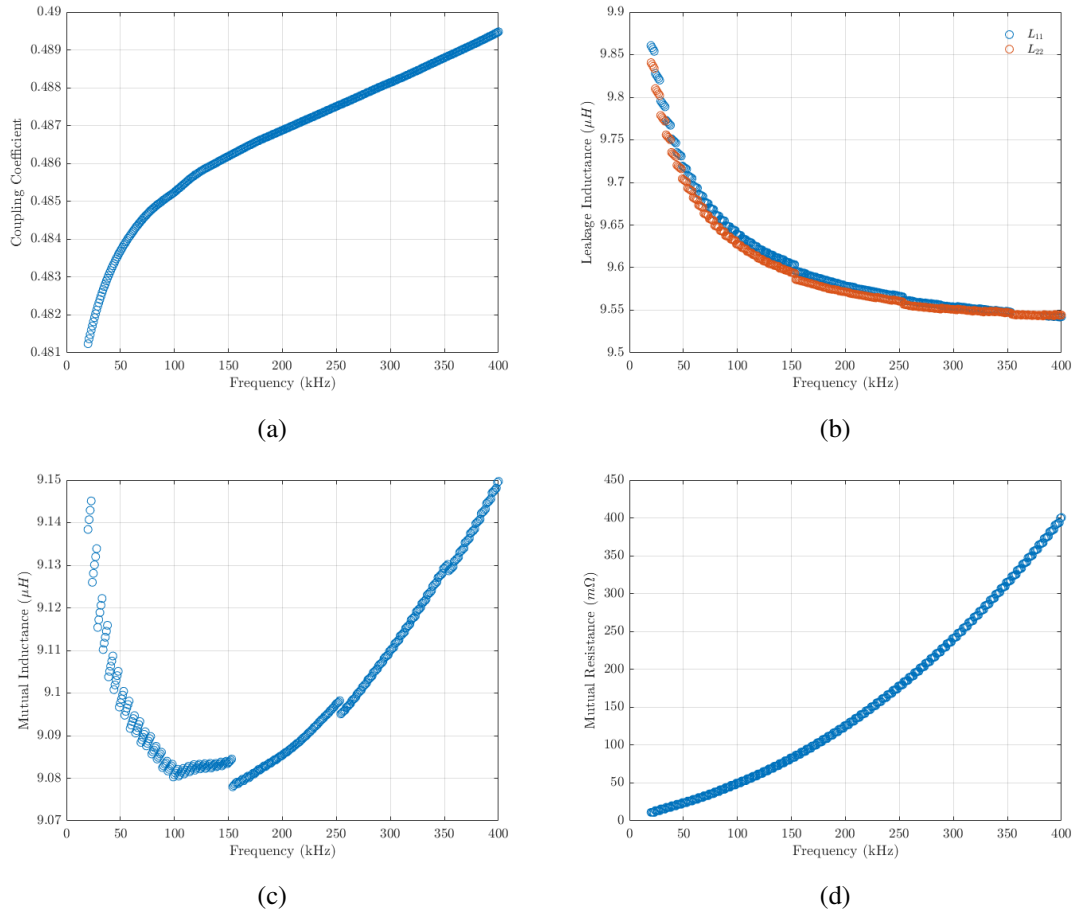


Figure 3.3: Variation with frequency of the (a) coupling coefficient , (b) leakage inductance, (c) mutual inductance, (d) mutual resistance

First, analysing the coupling coefficient variation with frequency in Fig. 3.3(a), it is possible to observe that it has its lower value for low frequencies, with a steep increase until around 100kHz and a slower increase for higher frequencies. The improvement of the coupling coefficient with increased frequencies was shown in the literature review and is now confirmed by the previously collected data, which means that higher frequencies allow that a higher amount of the magnetic flux from one coil is captured by the other. In Fig. 3.3(b) it is possible to observe the variation of the leakage inductance with frequency. As it is obvious, with higher coupling coefficients for higher frequencies, being the coupling stronger, the leakage magnetic flux should be smaller, which is confirmed in the plot. Further, being also dependent on the coupling coefficient in an opposite way from the leakage inductance, the mutual inductance increases with frequency as well, as shown in Fig. 3.3(c), with a higher amount of magnetic flux being shared. One should note that the scales of the two described plots are short, meaning that these variations are not so obvious as it might be seen from the graphs' shape. Additionally, in Fig. 3.3(d), it is observed that the mutual resistance representing the medium losses is very dependent on the frequency, as it was expected. A stronger varying magnetic field is created with higher frequencies, which, when

the surrounding medium is composed by a conductive material such as salted water, will induce higher Eddy currents on it.

Further, taking in account both coils series resistance variation with frequency in Fig- 3.4(a), it is possible to observe a steady increase to significant values from around 100mΩ below 50kHz up to around 700mΩ at 400kHz, which increases the conduction losses in the coil. Additionally, in Fig. 3.4(b) it can be seen that the inductance slightly decreases from low frequency values up to above 150kHz, after which it remains more or less constant.

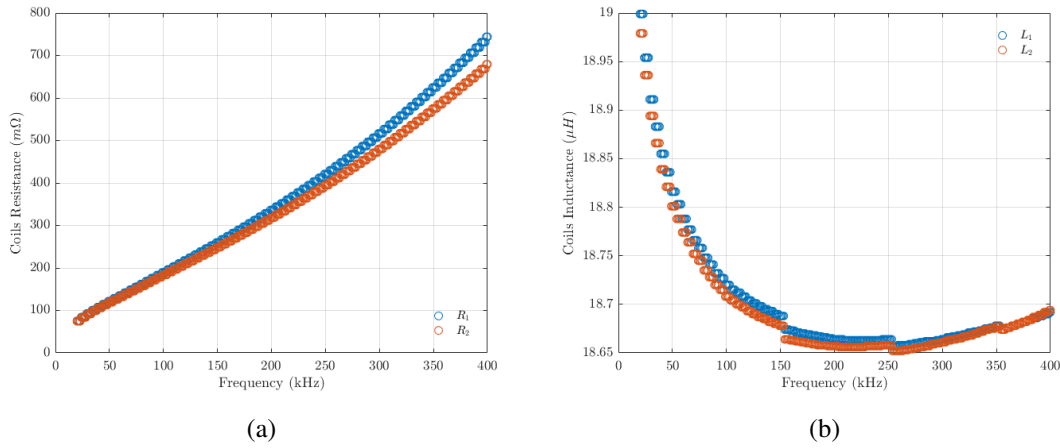


Figure 3.4: (a) Primary and secondary coils series resistance, (b) inductance variation with frequency

Finally, it is possible to draw both coils quality factor, variation with the operating frequency, $Q = \frac{\omega L}{R}$, as shown in Fig. 3.5(a). Here, it becomes obvious that there is a frequency that maximises this value which, in this case, is between 150kHz and 300kHz. With that in mind, one can now apply the maximum coupling efficiency using 2.4, presented in the Literature Review chapter, resulting in the plot from Fig. 3.5(b). Through these results it is possible to determine the optimum frequency region at which the coupled coils should be operated, meaning that to maintain the efficiency between 94% and 95%, the operating frequency should be between 130kHz and 400kHz.

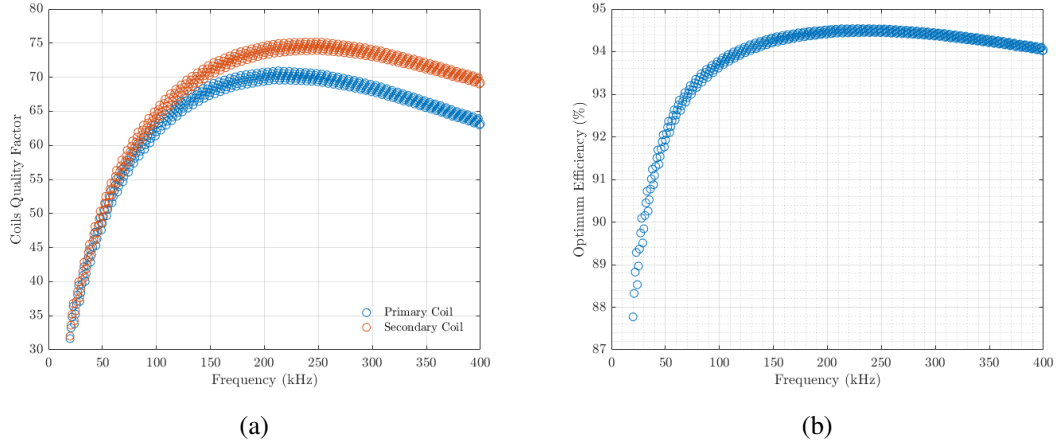


Figure 3.5: (a) Primary and secondary coils quality factor variation with frequency, (b) Coupling maximum efficiency variation with frequency

3.2 Simulation

In this section, some simulations based on analytical expressions from the coupled coils T-model and using the above described data are presented.

Now, taking the circuit from Fig. 3.2 as the starting point and adding two series resonance capacitors, C_1 and C_2 , a load resistance, R_L , and a voltage source, V_i , one can obtain the circuit from Fig. 3.6.

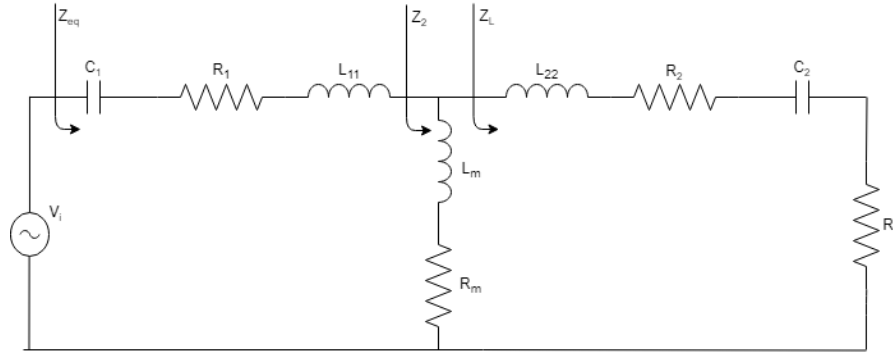


Figure 3.6: T-model with series resonance capacitors, load resistance and voltage source

From the above circuit it is then possible to get the impedance expressions from (3.1) to (3.3):

$$Z_L = j\omega L_{22} + \frac{1}{j\omega C_2} + R_2 + R_L \quad (3.1)$$

$$Z_2 = \frac{1}{\frac{1}{Z_L} + \frac{1}{j\omega L_m + R_m}} \quad (3.2)$$

$$Z_{eq} = \frac{1}{j\omega C_1} + j\omega L_{11} + R_1 + Z_2 \quad (3.3)$$

Being now possible to calculate the input power, the output power and consequent efficiency, as in (3.4), (3.5) and (3.6):

$$P_{in} = \frac{1}{2} \cdot \text{Re} \left\{ \frac{V_i^2}{Z_{eq}} \right\} \quad (3.4)$$

$$P_{out} = P_i \cdot \frac{\text{Re}\{Z_2\}}{R_1 + \text{Re}\{Z_2\}} \quad (3.5)$$

$$\eta = \frac{P_{out}}{P_{in}} \quad (3.6)$$

Using a MATLAB script containing these expressions and executing frequency and load sweeps, the following described plots were obtained.

First, a fixed series resonant capacitance value was chosen, being in this case $C_1 = C_2 = 68\text{nF}$ for a resonance frequency of around 140kHz. Then, with the previous expressions and the coils data, it was possible to analyse the efficiency and output power behaviour along the frequency for different R_L values as it is shown in Fig. 3.7(a) and Fig. 3.7(b), respectively. Regarding the efficiency, it can be seen that independently of the load value, the maximum always happens around the resonance frequency, and it is shown that higher loads are more capable of maintaining a higher efficiency at frequencies above resonance. Further, it is possible to compare the different efficiency waveforms with the previously calculated and drawn optimum frequency for the coupled coils. Regarding the output power distribution, the first thing that is observable is that the frequency splitting is not observed in the entire range of loads. Instead, as the load value increases, this splitting disappears and the maximum output power tends to happen in the resonance frequency, as it can better be observed for the maximum tested load of 13Ω . In opposite, it can be seen in the smaller loads that in the resonance frequency the output power suddenly decreases, creating a valley. Such behaviour is especially notorious in the smaller load of 5Ω .

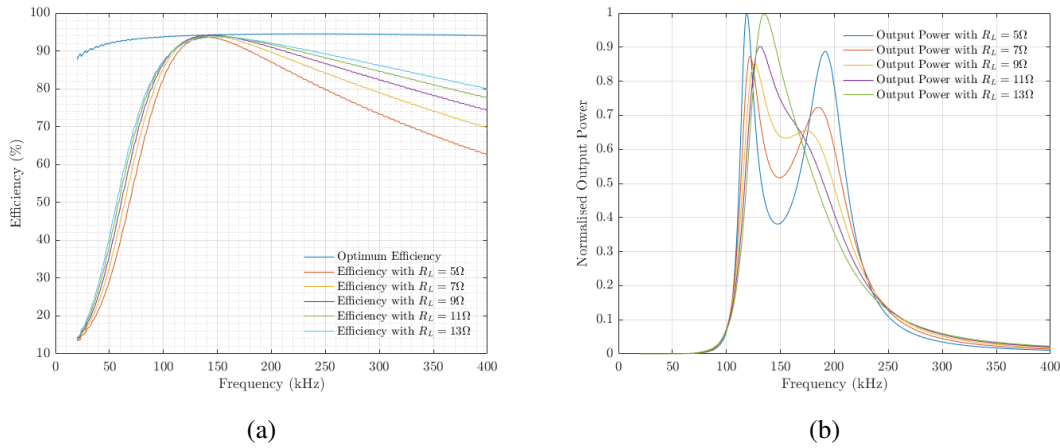


Figure 3.7: (a) Coupling efficiency with different loads and resonance capacitors of 68nF, (b) delivered power with different loads and resonance capacitors of 68nF

Now, performing a similar analysis but using a load sweep with different frequencies, it can be concluded by the plot in Fig. 3.8(a) that there is always a load value that maximises the efficiency, being this load value dependent on the operating frequency. It is also observable that the curve that reaches a higher efficiency is the curve of the resonance frequency of 140kHz. Moving on to the output power results in Fig. 3.8(b), one can conclude that the resonance frequency curve presents a different behaviour compared to the others, being constantly increased as the load increases. On the other hand, different frequencies will present maximum output power at specific load values.

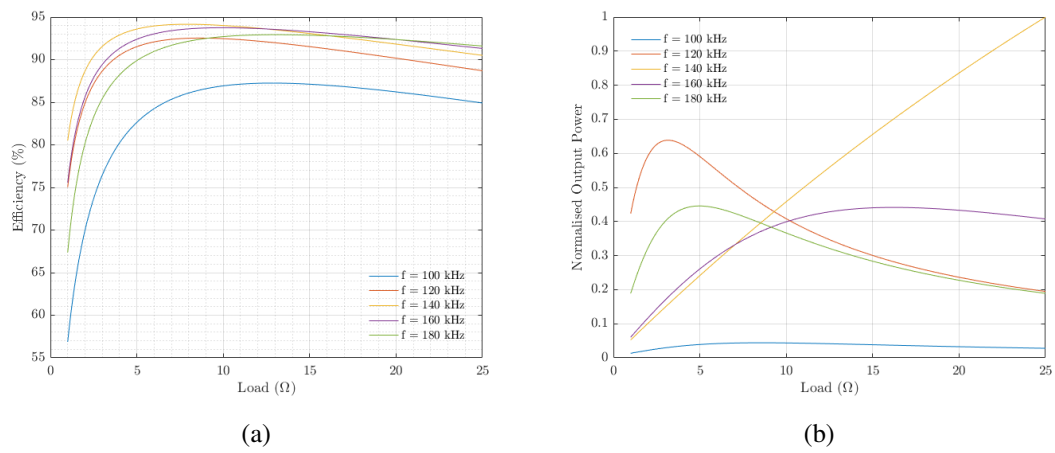


Figure 3.8: (a) Coupling efficiency at different frequencies and resonance capacitors of 68nF, (b) delivered power at different frequencies and resonance capacitors of 68nF

After the previous analysis have been conducted for fixed C_1 and C_2 values and a set of loads, it is now time to make a similar analysis but using more specific values instead. In the next analysis, frequency sweeps are still performed but, for each frequency step, the series resonance capacitors and load values are calculated to maximise the efficiency, obtaining the values from Fig. 3.9. Here, it is possible to see that the load value should grow linearly with the frequency for an optimum efficiency, whereas the capacitance value that optimises the efficiency for each frequency value is the one that allows resonance at that frequency.

Using now these series capacitances and load values, the resulting efficiency is as displayed in Fig. 3.9(b), which is completely overlapped with the previously calculated optimum efficiency of the coupled coils from Fig. 3.5. The resulting output power when operating in such conditions is depicted in Fig. 3.9(c) and presents an abrupt decrease at low frequencies until around 150kHz, adopting then a more steady decrease as the frequency keeps increasing.

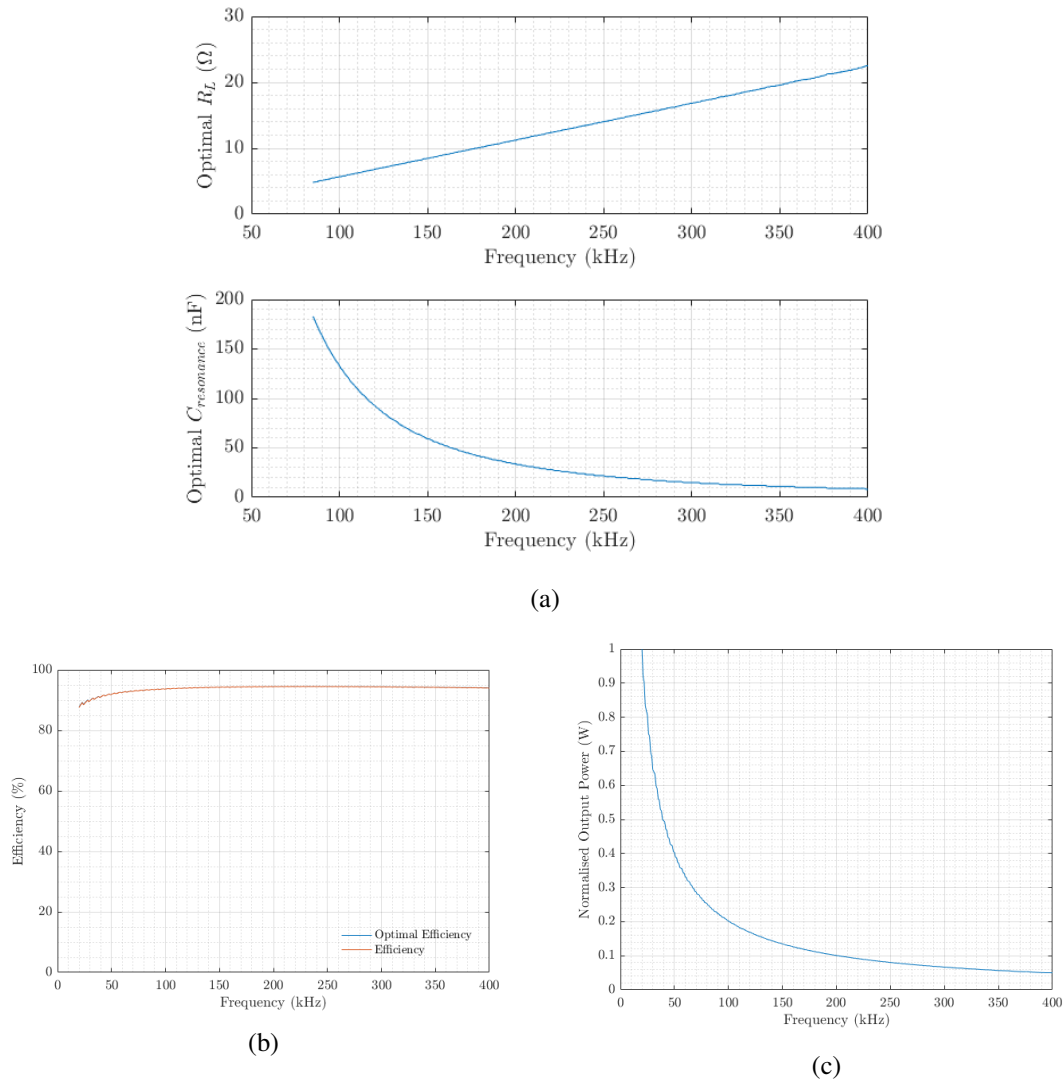


Figure 3.9: (a) Resonant capacitors and load resistance values for optimum efficiency, (b) efficiency with optimised resonant capacitors and load resistance, (c) output power with optimised resonant capacitors and load resistance

3.3 Summary

In this chapter, an analysis of the coupled coils to be used in the experimental phase was performed from different points of view using the previously collected characterisation data. The main goal of the analysis was to evaluate the coupled coils efficiency and output power behaviour when connected to a resistive load either by varying the operating frequency and by varying the load value.

Such analysis have shown the compliance of the results with what was expected after the literature review.

Chapter 4

Proposed Inverter

In this chapter, a theoretical approach to the scope of this thesis is presented. In a wireless power transfer system, the whole system efficiency is determined by the efficiency of each block of the system. The main blocks are the boost converter, that steps up the input battery voltage to a more appropriate level for power transfer, the power inverter that converts the DC power into AC power, the coupled coils for inductive coupling, the rectifier to convert the AC power into DC power and the load. Hence, to increase the overall system efficiency, one should increase each block efficiency individually. That being said, the scope of this work is to increase the efficiency of the power inverter on the transmitter taking in account the present state of the rest of the blocks between it and the load.

As such, first a literature review regarding some power inverter topologies is presented, followed by a theoretical analysis of the class-DE topology and respective simulation. Then, a new theoretical analysis of the class-DE topology applied to the wireless power transfer system is performed, followed by new simulations.

4.1 Class DE Introduction and Literature Review

The Class D is a power inverter topology dating from 1959 [34] in which the switches conduct on alternate half cycles, each with a duty factor approaching 50%. A series resonant network transforms the generated square voltage waveform into a sinusoidal load current [34]. Theoretically, in this topology, the switch current is zero while the switch voltage is high and the voltage is zero when the switch is closed and the current flows. Additionally, its switching transitions are assumed to occur instantly, in zero time, meaning that ideally its efficiency is 100% [35]. However, the class-D power amplifier is known to have a reduced efficiency caused by the parasitic capacitances across the switch device under high-frequency operation because of the nonzero-voltage turning on [35, 36], presenting switching losses due to the charging and discharging of these device capacitances in each period, which may dissipate a significant amount of energy, especially at higher frequencies [34].

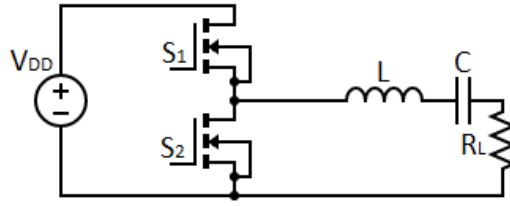


Figure 4.1: Class-D circuit

On the other hand, the Class E is a power inverter topology dating from 1975 [34] that can operate at high frequencies up to tens of MHz and power levels up to kW, while maintaining a high efficiency [37]. With a proper choice of circuit parameters this power amplifier topology is operated under zero voltage switching (ZVS) and zero voltage derivative switching (ZVDS) [37], eliminating the switching losses caused by the parasitic capacitance, which is realised through a shunt capacitor in parallel with the switch and allowing a dead-time between the switch on times [35, 36]. This happens by turning on the switch after its shunt capacitance has been discharged to the load [34]. The transistor S from Fig. 4.2 is usually switched with a duty factor of 50%. While the transistor is off, its current is null and the voltage increases up to a maximum of $3.6 \cdot V_{DD}$ and decreases back to zero, which is when the transistor is switched on again and its current increases up to a maximum of 2.9 times de input DC current, before the transistor is switched off again and the current drops to zero before the transistor voltage starts to increase from zero [37].

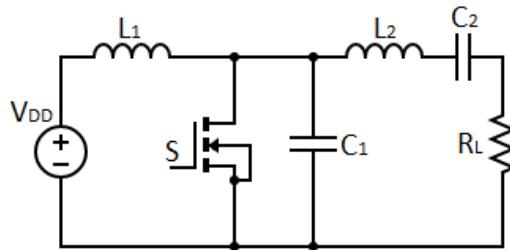


Figure 4.2: Class-E circuit

While class E and class D power inverter topologies switches have similar peak currents, the biggest disadvantage of the class-E topology is the peak voltage stress under which the switch is exposed when compared to the class-D topology, which is equal to $3.6V_{DD}$ in class E and only V_{DD} in class D [34].

That being said, a hybrid power inverter topology combining the best features of the both previously presented topologies [34, 38], known as class-DE power amplifier, will now be approached. Basically, it consists on the voltage mode class-D topology with switching operation similar to the class E. The key idea is to allow the shunt capacitances to charge/discharge to the next desired voltage level, so that when the switching operation occurs the shunt capacitances are

completely discharged. This ZVS and ZVDS condition minimises the switching losses, which may be dominant whenever the DC voltage is large or the operation frequency is high.

In this circuit topology, the switches turn on and off alternately with a dead time in between, in which the sinusoidal output current (caused by a high quality factor) charges one shunt capacitor and discharge the other, swinging the the load voltage between V_{DD} and 0 or $-V_{DD}$, depending on whether it is a half or a full bridge, during a dead time when both switches are open, with a null slope at the end of the dead time [34, 39, 40].

In conclusion, the class-DE power inverter achieves a high frequency efficiency equivalent to the one of class D under low frequency operation, having the same switch voltage stress [36], making it a good choice for employment on this system.

4.2 Class DE Analysis

In this section, a deep theoretical analysis of the class-DE power inverter will be presented. Fig. 4.3 depicts a full-bridge class-DE circuit, which is basically identical to class D with some minor modifications. An additional inductor is required in the load network as well as additional linear switch shunt capacitances to absorb the nonlinear behaviour of the transistors junction capacitances, both taking a fundamental role in the class DE realisation.

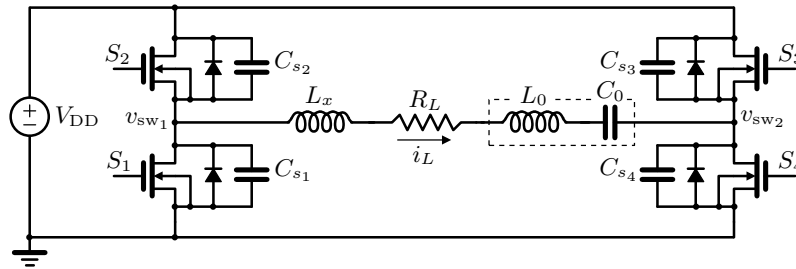


Figure 4.3: Class DE circuit [33]

The load network should be designed in a way that imposes that the voltage value at the v_{sw1} node will be null when the transistor turns on, maintaining the null voltage until it turns back off. Following the last transition, the v_{sw1} node will be charged to V_{DD} with energy coming back from the load network. The transistor is then turned on again in the instant that the v_{sw1} node voltage reaches V_{DD} again, in order to avoid the shunt capacitance to discharge. The second leg of the full bridge will have a similar behaviour on the v_{sw2} node, with the difference of doing it on alternated times with the first leg. Further, in order to satisfy the ZVS criteria, the current at the load network must be zero at the turning on instant of either transistors, otherwise the transition would be made without the shunt capacitances being fully charged or discharged, depending on if it is referring to the upper or the lower transistor, meaning that current spikes would flow through the switches. The previous means that the turning on transitions will occur at the load network current zero crossings, which means that the time between the transitions of the two transistors on each bridge leg is half period, or π on a 2π period, in order to be coincident with the sinusoidal current zero

crossings. This way, the class DE minimises the switching losses that are common in class D, improving its efficiency [33].

Assuming that the circuit achieved the steady-state regime, Fig. 4.4 shows the circuit waveforms. The further analysis of this power inverter topology will be based upon the assumptions:

1. the MOSFETs (S_1 to S_4) operate as ideal switches, i.e. with null conduction resistance and instantaneous switching transitions;
2. S_1 and S_3 form a pair of switches; S_2 and S_4 form another one. Each transistor from a pair is driven the same way;
3. the shunt capacitance (C_{s1} to C_{s4}) of each MOSFET is considered linear (and all equal to C_s);
4. the load network has a sufficiently high quality factor such as one can guarantee a pure sinusoidal current flowing in the load network.

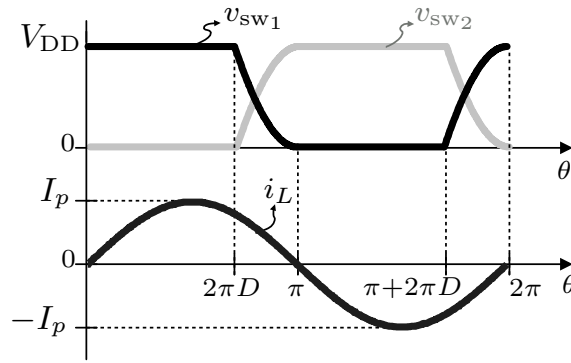


Figure 4.4: Class DE waveforms [33]

The load network comprises an LC tank formed by L_0 and C_0 , resonant at the switching frequency. An additional inductance L_x is included in the network, to make it inductive at the operating frequency. The load network current $i_L(\theta)$ is given by:

$$i_L(\theta) = I_p \sin(\theta) \quad (4.1)$$

where I_p is the peak value of the current signal, being possible to identify the off-on transitions θ_s for both switches on each leg by imposing:

$$i_L(\theta_{s1}) = 0 \quad (4.2)$$

$$i_L(\theta_{s2}) = 0 \quad (4.3)$$

$$i_L(\theta_{s3}) = 0 \quad (4.4)$$

$$i_L(\theta_{s4}) = 0 \quad (4.5)$$

with $\theta_{s1} = \theta_{s3}$ and $\theta_{s2} = \theta_{s4}$. Taking in account that all the charge from the shunt capacitors is shared with the load, then it is possible to set the following equation

$$V_{DD} = \frac{1}{2C_s} \int_{\pi}^{2\pi D} -\frac{i_L(\theta)}{\omega} d\omega = \frac{I_p}{2\omega C_s} \int_{2\pi D}^{\pi} \sin(\theta) d\theta \quad (4.6)$$

Resulting in

$$I_p = \frac{2\omega C_s V_{DD}}{1 + \cos(2\pi D)} \quad (4.7)$$

Observing (4.7), it is possible to conclude that as D tends to 50% (equivalent to class D), the peak current tends to infinity. That happens because, with a null dead time, the switching nodes voltage approaches to a square waveform, meaning that the capacitors must charge and discharge instantaneously from and to the load, if any switch current must be avoided. This means that the load value should also tend to zero.

Taking into account the voltage at one side of the full bridge, i.e. the switching node $v_{sw1}(\theta)$ that will be decomposed in time as follows:

$$v_{sw1}(\theta) = \begin{cases} V_{DD} \Leftarrow & 0 \leq \theta \leq 2\pi D \\ v_{sw1}^{\diamond}(\theta) \Leftarrow & 2\pi D \leq \theta \leq \pi \\ 0 \Leftarrow & \pi \leq \theta \leq \pi(1+2D) \\ v_{sw1}^{\star}(\theta) \Leftarrow & \pi(1+2D) \leq \theta \leq 2\pi \end{cases} \quad (4.8)$$

where

$$v_{sw1}^{\diamond}(\theta) = \frac{1}{2C_s} \int_{2\pi D}^{\theta} -\frac{i_L(\theta)}{\omega} d\omega + V_{DD} \quad (4.9)$$

$$= \frac{I_p}{2C_s} \left(\cos(\theta) - \cos(2\pi D) \right) + V_{DD} \quad (4.10)$$

$$= V_{DD} \cdot \left(1 + \frac{\cos(\theta) - \cos(2\pi D)}{1 + \cos(2\pi D)} \right) \quad (4.11)$$

and, similarly,

$$v_{sw1}^{\star}(\theta) = \frac{1}{2C_s} \int_{\pi(1+2D)}^{\theta} -\frac{i_L(\theta)}{\omega} d\omega \quad (4.12)$$

$$= V_{DD} \cdot \frac{\cos(\theta) + \cos(2\pi D)}{1 + \cos(2\pi D)} \quad (4.13)$$

The fundamental voltage between the switching nodes, which is the same at the load due to LC series filtering, can be determined by the complex Fourier coefficient:

$$\mathbf{V}_1 = \frac{a_1 + jb_1}{2} = \frac{1}{2\pi} \int_{\Theta} 2v_{sw1}(\theta) [\cos(\theta) - j\sin(\theta)] d\theta \quad (4.14)$$

$$= \frac{V_{DD}}{2\pi} \cdot \frac{2\pi(1-2D) + \sin(4\pi D) - j2\sin^2(2\pi D)}{1 + \cos(2\pi D)} \quad (4.15)$$

The complex Fourier coefficient for the load current can also be determined, as follows:

$$\mathbf{I}_1 = -j \frac{I_p}{2} = -j \frac{\omega C_s V_{DD}}{1 + \cos(2\pi D)} \quad (4.16)$$

The output power at the fundamental frequency can be derived from (4.15) and (4.16)

$$P_{out} = \frac{1}{2} \mathbf{Re}\{2\mathbf{V}_1 \cdot 2\mathbf{I}_1^*\} = \frac{2\omega C_s V_{DD}^2}{\pi} \cdot \left(\frac{\sin(2\pi D)}{1 + \cos(2\pi D)} \right)^2 \quad (4.17)$$

The result in (4.17) shows that, for a fixed C_s , the output power is linearly dependent in frequency. This is because the fundamental current depends on ω , whereas the output voltage only depends on D and V_{DD} . However, not all duty factor values are possible for each C_s if the circuit is to be operated in class DE. Additionally, it is also evident that as D tends to 50%, the system turns into a class D and its output power tends to infinity, which is not true. For these reasons, the value of C_s should be expressed in terms of duty factor D , which can be derived from the impedance of the load network as further described.

The class DE load impedance can be obtained as follows:

$$\mathbf{Z}_L = \mathbf{V}_1 / \mathbf{I}_1 = R_L + jX_L = \frac{1}{2\pi\omega C_s} \cdot \left\{ 2\sin^2(2\pi D) + j[2\pi(1-2D) + \sin(4\pi D)] \right\} \quad (4.18)$$

The load resistance R_L is directly determined from (4.18) as follows

$$R_L = \mathbf{Re}\{\mathbf{Z}_L\} = \frac{\sin^2(2\pi D)}{\pi\omega C_s} \quad (4.19)$$

which is maximum for $D = 0.25$ for a fixed ωC_s . Now from (4.17) and (4.19), one can get

$$P_{out} = \frac{2}{\pi^2} \cdot \frac{V_{DD}^2}{R_L} \cdot \left(\frac{\sin^2(2\pi D)}{1 + \cos(2\pi D)} \right)^2 \quad (4.20)$$

Assuming the operating frequency defined by the resonance

$$\omega^2 = 1/(L_0 C_0) \quad (4.21)$$

the additional reactance L_x can be derived from (4.18) as follows

$$L_x = \mathbf{Im}\{\mathbf{Z}_L / \omega\} = \frac{L_0 C_0}{2\pi C_s} \cdot \left[2\pi(1-2D) + \sin(4\pi D) \right] \quad (4.22)$$

which can be related with R_L , or Q_L as

$$L_x = \frac{\alpha(D)}{2} \cdot \frac{R_L}{\omega} = \frac{\alpha(D)}{1 - \alpha(D)} \cdot \frac{L_0}{Q_L} \quad (4.23)$$

where Q_L is the loaded quality factor of the RLC series of the output network described by

$$Q_L = \frac{\omega(L_0 + L_x)}{R_L} \quad (4.24)$$

and $\alpha(D)$ is a duty-cycle dependent function given by

$$\alpha(D) = \frac{2\pi(1 - 2D) + \sin(4\pi D)}{\sin^2(2\pi D)} \quad (4.25)$$

A complete mapping of all parameter is depicted in Fig. 4.5, being useful for design reference.

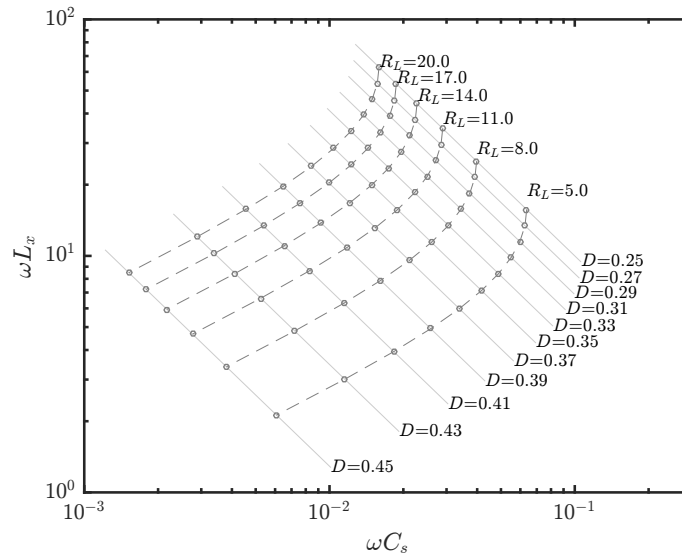


Figure 4.5: Mapping of class-DE domain with parameters ωL_x vs ωC_s for possible different pairs D and R_L .

4.3 Class DE Simulation

In order to verify the previously described theoretical formulations, a set of simulations in Cadence Virtuoso was performed, using the PSS analysis (periodic steady state analysis), since the initial transients were not taken in account in this study. As such, the circuit from Fig. 4.6 was implemented and the values of the variables were defined, with some of them (C_s , L_x , C_0) being dependent on others with the previously presented expressions, namely 4.19, 4.23 and 4.21, in order to allow a sweep of any of the independent variables (resonant ω , D , R_L , L_0) while maintaining a class-DE operation, validating it.

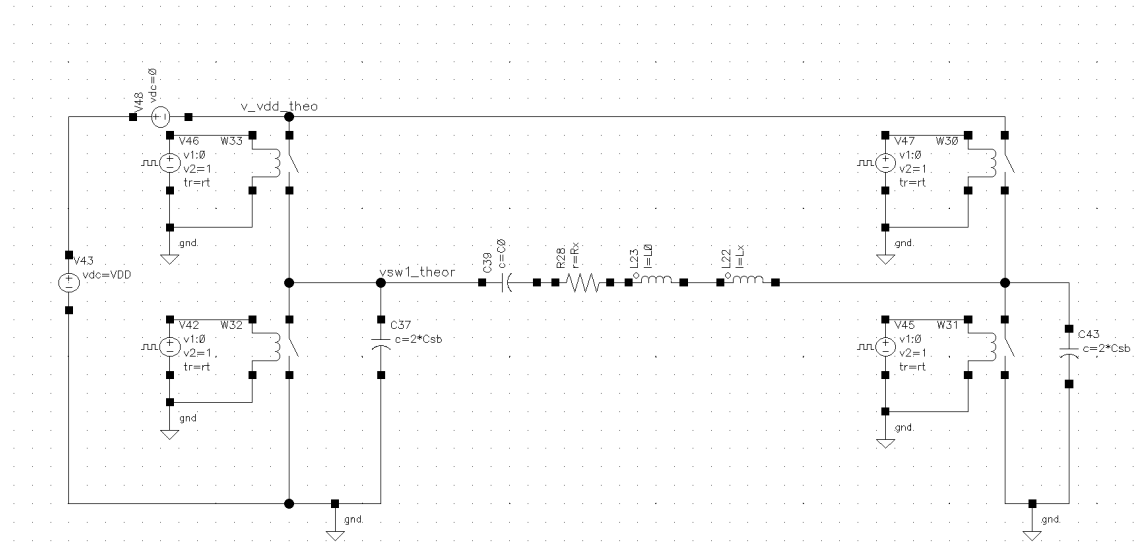


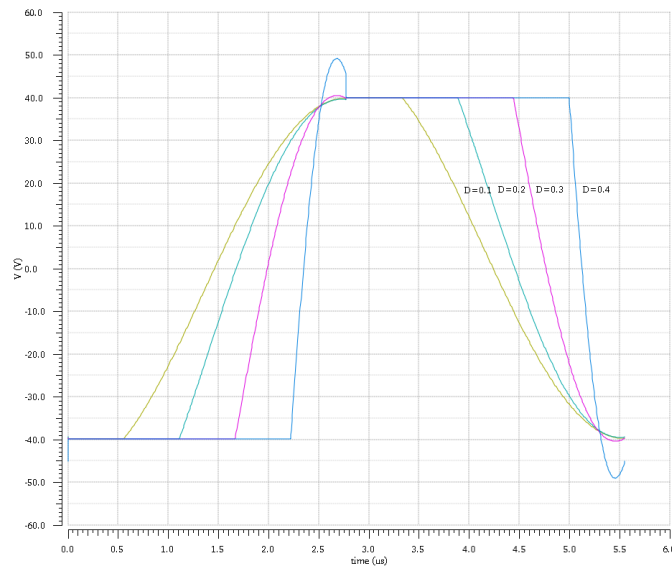
Figure 4.6: Simulation schematic

In the simulation schematic of Fig. 4.6, the top switch of the left leg of the inverter and the bottom switch of the right leg of the inverter are driven with the same command signal, which is a pulse width modulation (PWM) signal with frequency f and pulse width as a function of the duty factor D . In opposite, the bottom switch of the left leg of the inverter and the top switch of the right leg are driven by a different command signal, which is delayed by 50% of the period and has the same frequency and pulse width as the other previously referred command signal. Additionally, the rise and fall times of all switches were set with the same small value in order to reduce its influence on the circuit behaviour. The same principle applies to the conducting series resistance of each switch R_{on} , that was made small for a reduced influence on the circuit. Such simplification was created in order to clearly evaluate the previously stated theory, which assumed the MOSFETs to operate as ideal switches, i.e. with null conduction resistance and instantaneous switching transitions. Following, the variables in Cadence Virtuoso simulation were set like it is shown in Table 4.1. Finally, one should mention that the shunt capacitances C_s in each switch were reduced to one single shunt capacitance (which in the circuit is represented by C_{sb}) in each of the bottom switches with its value doubled relative to the original one, which results in a valid equivalent circuit, since the voltage and current on the top capacitance is equal to the ones in the bottom capacitance.

Table 4.1: Theoretical class DE simulation variables

f	180 kHz
ω	$2\pi f$
L_0	$5 \mu\text{H}$
C_0	$\frac{1}{\omega^2 L_0}$
D	10% to 40%
R_x	2 to 20 Ω
V_{DD}	40 V
T	$\frac{1}{f}$
α	$\frac{2\pi(1-2D)+\sin(4\pi D)}{\sin^2(2\pi D)}$
L_x	$\frac{\alpha R_x}{2\omega}$
C_{sb}	$\frac{\sin^2(2\pi D)}{\pi\omega R_x}$
pulse width	$D \cdot T$
delay time	$\frac{T}{2}$
rise time	$0.000001 \cdot T$
fall time	$0.000001 \cdot T$
R_{on}	$10\mu\Omega$

Following the definition the circuit variables, different variable sweeps were executed in order to evaluate the validity of the theoretical class-DE model. As such, to observe the influence of the duty factor in the circuit, different values of D were set and the waveforms of Fig. 4.7 were obtained for the differential switching voltage. Through this graph, it is possible to conclude that the waveform gets distorted relative to a theoretical class-DE as the duty factor value approaches 50%. Such behaviour will further be explained by analysing other electrical quantities.

Figure 4.7: Differential switching voltage for $R_x = 10\Omega$ and various D

After observing the resulting differential switching voltage, one should now evaluate the load

current waveform. As it was stated above in section 4.2, it is assumed the load current to be purely sinusoidal for the theoretical formulations to be valid. However, as it may be observed in Fig. 4.8, this is not the case, especially for higher duty factor values. Instead, the resulting load currents are a distorted approximation to a pure sinusoid, which may be the reason why the differential switching voltage is so different from a class DE for D values closer to 50%.

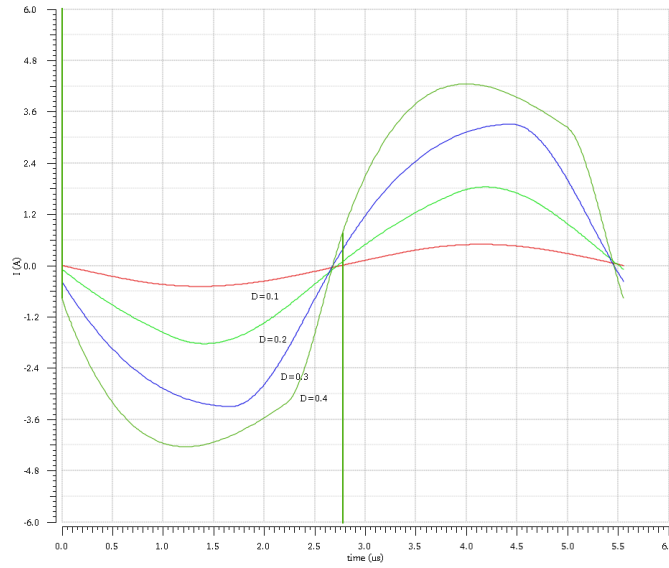
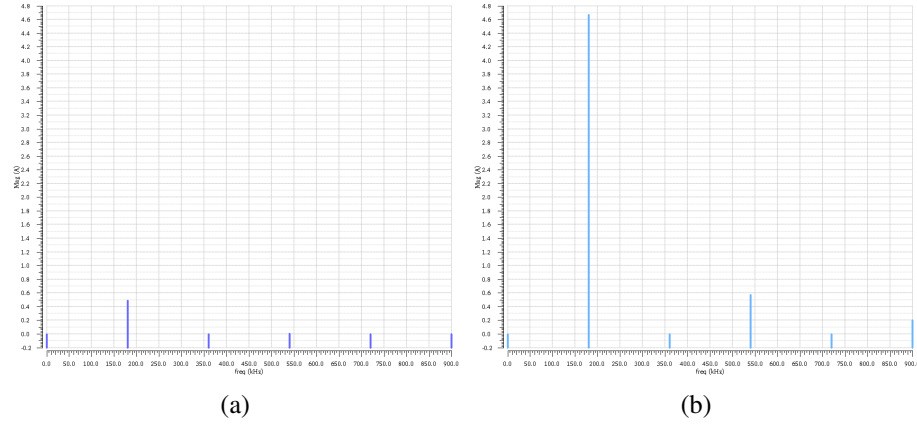


Figure 4.8: Load current for $R_x = 10\Omega$ and various D

In fact, one may confirm this difference from a pure sinusoid between higher and lower duty factor values by observing the harmonic distribution of the load current, which is depicted in Fig. 4.9(a) and Fig. 4.9(b), for a small duty factor value, $D = 0.1$ and a high duty factor value, $D = 0.4$. From the images, the presence of higher values of current in harmonics different from the fundamental frequency with higher duty factors becomes obvious. Such behaviour is explained by the small quality factor present in the load network, which is affected by the frequency, the inductance and the resistance. As stated previously, the load network should guarantee a sufficiently high quality factor to assume the absence of higher power harmonics and, thus, a pure sinusoidal load current. For this reason, the circuit behaviour was somewhat different to as class DE with higher duty factors. Specifically, in this simulation, the quality factor was $Q = \frac{2\pi \cdot 180000 \cdot 5 \cdot 10^{-6}}{10} = 0.5655$, which is fairly low.

Figure 4.9: Current harmonics for $R_x = 10\Omega$ and (a) $D = 0.1$, (b) $D = 0.4$

Now, having identified the reason for an operation that is not close to a pure class-DE, the circuit values may be changed in order to define a higher quality factor, reducing the load current harmonics outside the fundamental frequency. Such may be achieved by increasing the values of L_0 and frequency f , knowing that $Q = \frac{\omega}{L_0 R}$. Having that said, the variables in Cadence Virtuoso simulation were set like it is shown in Table 4.2.

Table 4.2: Theoretical class-DE simulation variables for high quality factor

f	350 kHz
ω	$2\pi f$
L_0	100 μH
C_0	$\frac{1}{\omega^2 L_0}$
D	10% to 40%
R_x	2 to 20 Ω
V_{DD}	40 V
T	$\frac{1}{f}$
α	$\frac{2\pi(1-2D)+\sin(4\pi D)}{\sin^2(2\pi D)}$
L_x	$\frac{\alpha R_x}{2\omega}$
C_{sb}	$\frac{\sin^2(2\pi D)}{\pi\omega R_x}$
pulse width	$D \cdot T$
delay time	$\frac{T}{2}$
rise time	$0.000001 \cdot T$
fall time	$0.000001 \cdot T$
R_{on}	10 $\mu\Omega$

Now that the quality factor is $Q = \frac{2\pi \cdot 350000 \cdot 100 \cdot 10^{-6}}{10} = 21.9911$ for a 10 Ω load resistance, as in the previous case, it is possible to observe in Fig. 4.11 that, even though not exactly, the differential switching voltage became closer to a theoretical class-DE for high duty factor values.

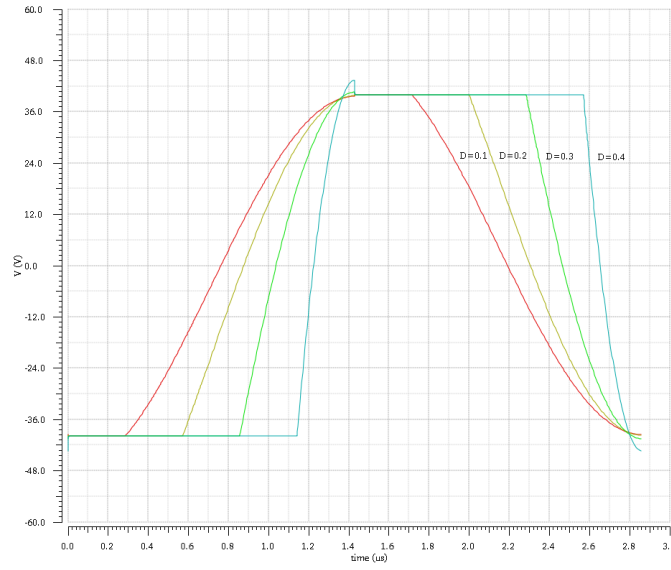


Figure 4.10: Differential switching voltage for $R_x = 10\Omega$ and various D

Actually, as it can be seen in Fig. 4.11 the load current waveforms are now much closer to a pure sinusoid, as it is required in the class-DE theoretical analysis.

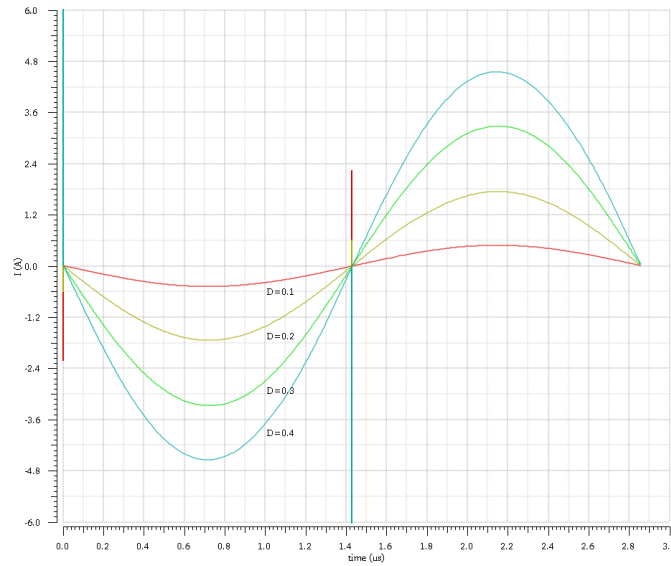


Figure 4.11: Load current for $R_x = 10\Omega$ and various D

Further, just as an instrument for comparison, it is now easy to observe in Fig. 4.12(a) and Fig. 4.12(b) that the current harmonics for both low and high duty factors is close to zero and the difference is obvious compared to the results obtained with a low quality factor.

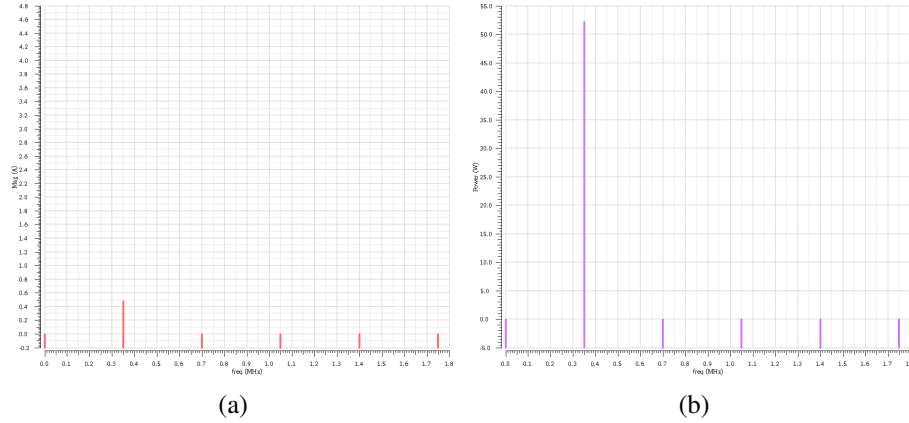


Figure 4.12: Current harmonics for $R_x = 10\Omega$ and (a) $D = 0.1$, (b) $D = 0.4$

From Fig. 4.12 it is also easy to observe a very significant difference between the supplied current with a high and a low duty factor. As it was predicted in section 4.2, the delivered power is continuously increased as the duty factor approaches 50%, like it is depicted in Fig. 4.13.

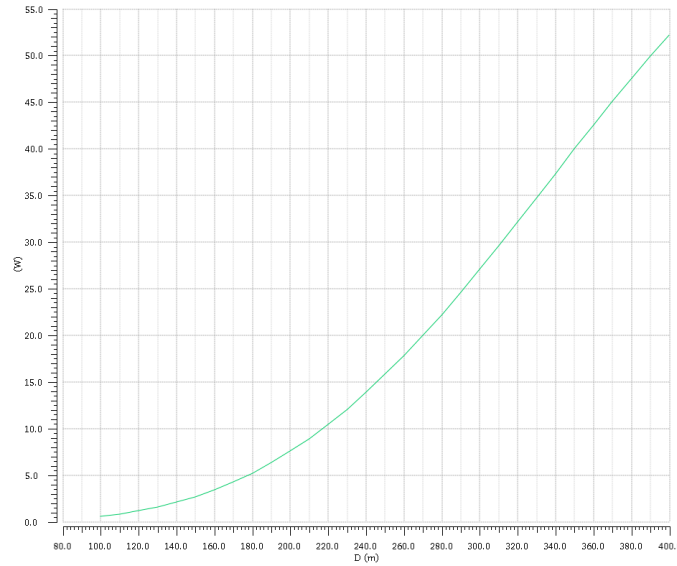


Figure 4.13: Delivered power with a variable D for $R_x = 10\Omega$

Having analysed and simulated the traditional scheme of a class-DE power inverter and having validated its theoretical formulations, such process will now be adapted and applied to a wireless power transfer system to achieve the same operation states.

4.4 Class DE in WPT Analysis

The circuit model shown in Fig. 4.14 represents an equivalent circuit for the WPT coupling coils, with the primary and secondary series resonances included (C_1 and C_2).

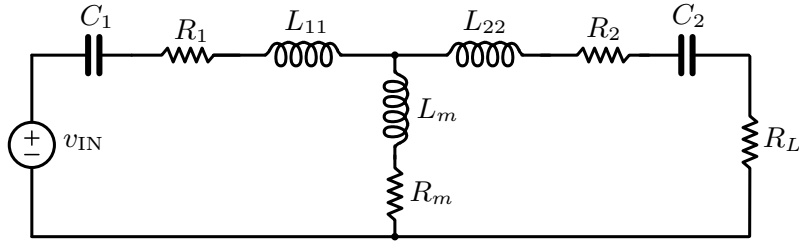


Figure 4.14: Coupled coils equivalent model [33]

For simplicity, the inductances L_{11} and L_{22} are considered as equal:

$$L_{11} = L_{22} = (1 - k)L \quad (4.26)$$

where k is the inductive coupling coefficient. Therefore, the mutual inductance is:

$$L_m = kL \quad (4.27)$$

and

$$C_1 = C_2 = C \quad (4.28)$$

Now, an operating frequency ω above resonance ω_0 is imposed:

$$\omega > \omega_0 = \frac{1}{\sqrt{LC}} \quad (4.29)$$

As a result, at the operating frequency the primary and secondary sides will be operated under an inductive regime, which may be represented by the circuit of Fig. 4.15.

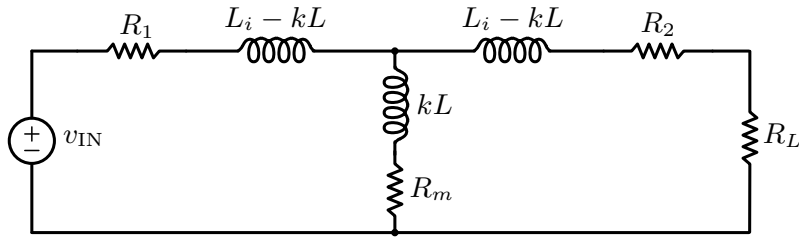


Figure 4.15: Equivalent circuit valid at frequencies above resonance [33]

The equivalent inductance L_i satisfies the following:

$$j\omega L_i = j\left(\omega L - \frac{1}{\omega C}\right) \quad (4.30)$$

Hence,

$$L_i = L - 1/(\omega^2 C) \quad (4.31)$$

In such circumstances, an equivalent circuit may be determined for the complete coupling system, resulting in a series of an inductor L_{eq} and a resistor R_{eq} , as shown in Fig. 4.16.

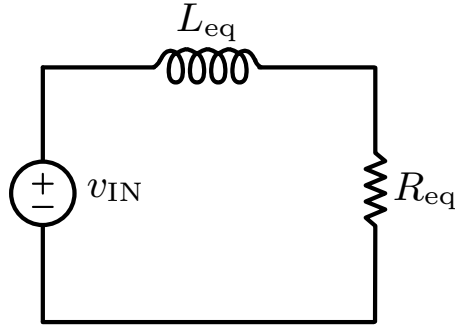


Figure 4.16: Input impedance seen from primary [33]

Such input inductance and resistance are given in (4.32) and (4.33), respectively.

$$L_{eq} = \frac{[R_m(L_i - kL) + kL(R_2 + R_L)](R_2 + R_L + R_m) - L_i R_m(R_2 + R_L) + \omega^2 L_i kL(L_i - kL)}{\omega^2 L_i^2 + (R_2 + R_L + R_m)^2} + L_i - kL \quad (4.32)$$

$$R_{eq} = \frac{\omega^2 L_i [R_m(L_i - kL) + kL(R_2 + R_L)] + R_m(R_2 + R_L)(R_2 + R_L + R_m) - \omega^2 kL(L_i - kL)(R_2 + R_L + R_m)}{\omega^2 L_i^2 + (R_2 + R_L + R_m)^2} + R_1 \quad (4.33)$$

Following the previous analysis, it is now possible to apply the results from section 4.2 into the wireless power transfer system model improving its transmitter efficiency, with the advantage of avoiding the addition of an extra coil to realise the class DE due to the operation above the resonant frequency.

Comparing the circuits in Fig. 4.3 and Fig. 4.16, it becomes evident that both circuit loads (R_L and R_{eq}) are the same. As such, it is possible to use R_{eq} of the WPT system to meet the class-DE conditions. Further, L_{eq} also needs to be equalised for class-DE operation, i.e. it must be equivalent to the earlier derived L_x , given in (4.23). The value chosen for C_s must absorb also the output capacitance of the active devices, although nonlinear.

Now, knowing that for a class DE on a fixed coupled coils T-model the manipulable variables are ω , D , C_s and R_L and that the model's equivalent R_{eq} and L_{eq} are both dependent on frequency and on R_L , it is possible to fix one of the four manipulable variables and define a trajectory with the remaining three. In a practical application, the best candidate to be fixed would be the shunt capacitance C_s , being the only physical/material element apart from R_L , which should be assumed to vary independently throughout the process. That being said, it is possible to reuse the mapping from Fig. 4.5 to trace a route with ω and D as a function of the load, R_L . The procedure to obtain such trajectory is further explained.

Having fixed the desired C_s value, an initial R_L should be set as starting point. Now, from that C_s and R_L , a D sweep should be performed and, on each D value, frequencies above resonance should be swept as well, until a match is found. This match happens when, for a certain D and ω , both the L_{eq} and R_{eq} value are equalised to the aforementioned L_x and R_L from the class-DE load

network in section 4.2, knowing that, for each C_s and R_L there is only one possible solution, one pair of D and frequency. The entire process is then repeated with the remaining load values, R_L , until each R_L has a defined frequency and D of operation as in the example shown in Fig. 4.17. The following example results were obtained using previously measured T-model frequency dependent dimensions on two real coupled coils, later used in the experimental phase.

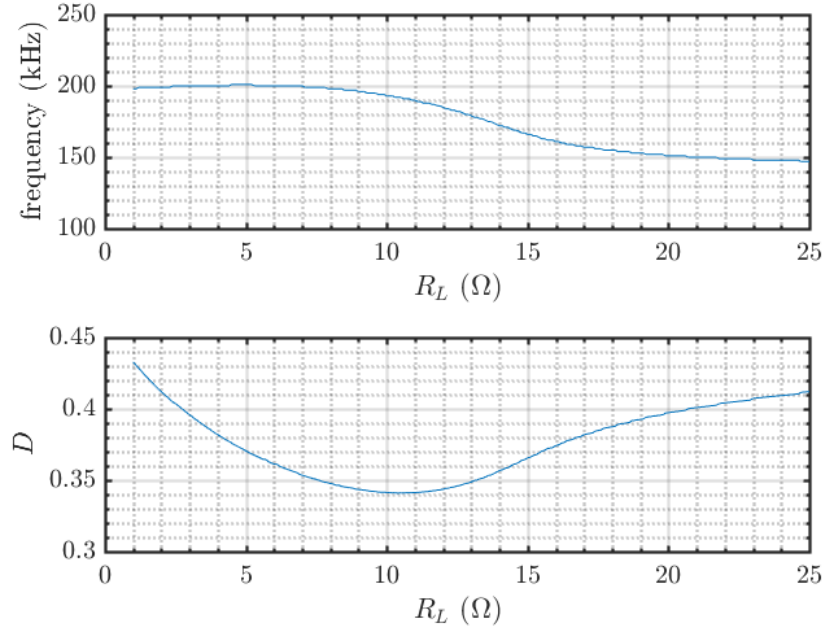


Figure 4.17: Resulting D and frequency values to achieve class DE on a coupled coils T-model for an example where $C_s = 31\text{nF}$

From that same example where $C_s = 31\text{nF}$, with the previous results it is possible to obtain the resulting R_{eq} and L_{eq} values for each R_L value with the corresponding frequency, as illustrated in Fig. 4.18 (a) and Fig. 4.18 (b), respectively.

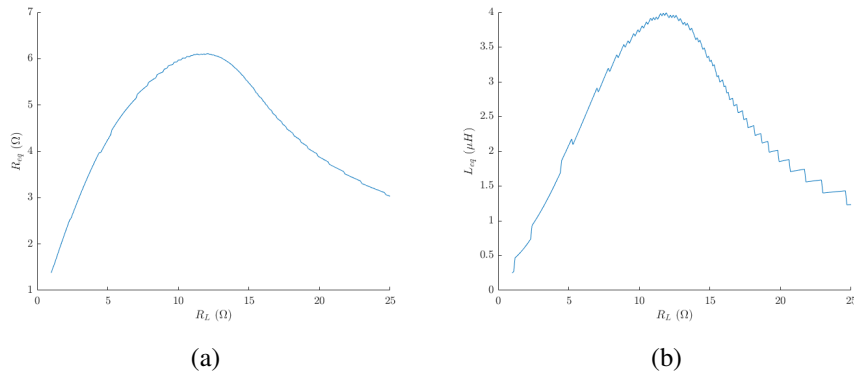


Figure 4.18: (a) - Equivalent resistance R_{eq} for an example where $C_s = 31\text{nF}$, (b) - Equivalent inductance L_{eq} for an example where $C_s = 31\text{nF}$

Finally, it is possible to overlap the obtained calculated results with the general mapping from Fig. 4.5 in section 4.2, resulting in the different trajectories depicted in Fig. 4.19 for different C_s values.

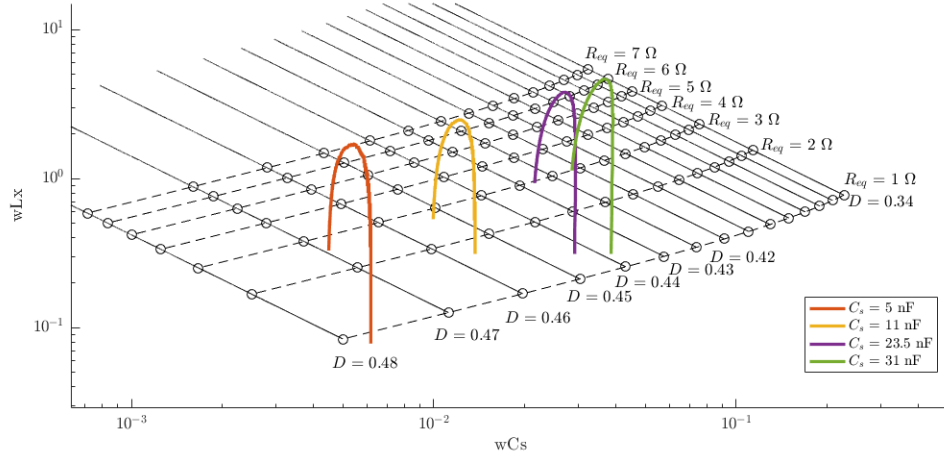


Figure 4.19: Mapping example for a various C_s values

4.5 Class DE in WPT Simulation

Having the analysis from section 4.4 been performed, the next step was to evaluate the validity of the class-DE theory applied to a wireless power transfer circuit model, the T-model as shown in Fig. 4.20, through simulations in Cadence Virtuoso, using the PSS analysis (periodic steady state analysis), since the initial transients were not taken in account in this study.

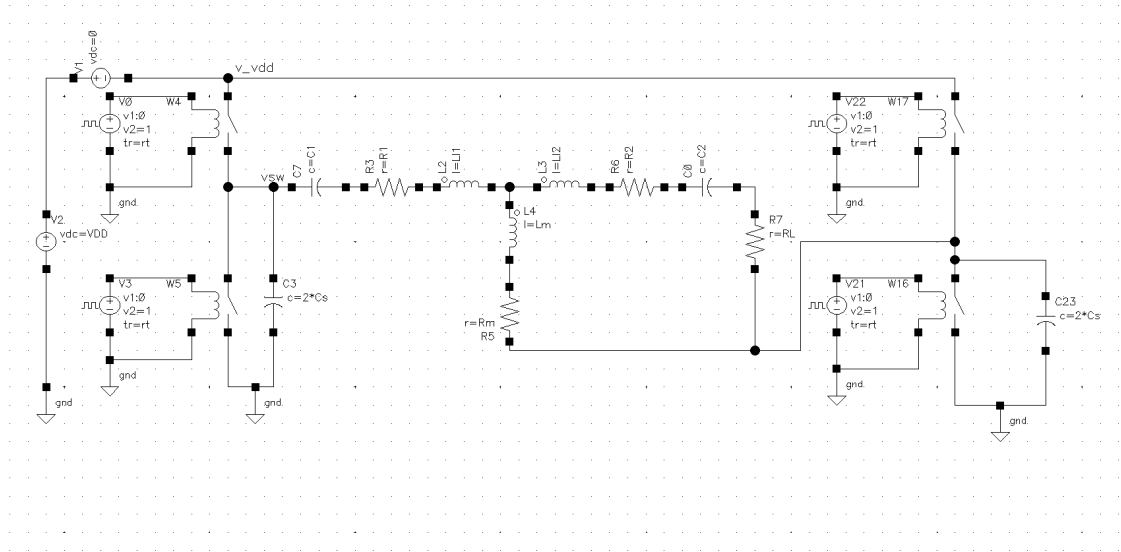


Figure 4.20: Simulation schematic for the wireless power transfer T-model

Such validation of theoretical results in the wireless power transfer model was conducted through simulation by applying the obtained frequency and duty-factor values from Fig. 4.17 to a T-model circuit where R_L has the corresponding value to each pair of those variables, and observing the resulting waveforms. Once again, as in the previous simulations, the top switch of the left leg of the inverter and the bottom switch of the right leg of the inverter are driven with the same command signal, which is a PWM signal with frequency f and pulse width as a function of the duty factor D . The bottom switch of the left leg of the inverter and the top switch of the right leg are driven by a different command signal, which is delayed by 50% of the period and has the same frequency and pulse width as the first command signal. The rise and fall times of all switches were set with the same small value in order to reduce its influence on the circuit behaviour. The conducting series resistance of each switch R_{on} was made small as well, for a reduced influence on the circuit. Further, as explained before, these simplifications were made in order to clearly evaluate the previously stated theory, which assumed the MOSFETs to operate as ideal switches, i.e. with null conduction resistance and instantaneous switching transitions. The variables in Cadence Virtuoso simulation were set like it is shown in Table 4.3. The shunt capacitances C_s in each switch were reduced to one single shunt capacitance in each of the bottom switches with its value doubled relative to the original one, which results in a valid equivalent circuit, since the voltage and current on the top capacitance is equal to the ones on the bottom capacitance.

Table 4.3: Wireless power transfer class-DE simulation variables for high quality factor

C_s	31 nF
R_L	1 to 25 Ω
f	variable with R_L (143 to 900 kHz)
D	variable with R_L
C_1	68 nF
C_2	68 nF
R_m	variable with f (0.07636 to 1.9884 Ω)
R_1	variable with f (0.16937 to 0.78521 Ω)
R_2	R_1
L_m	$k \cdot L_1$
L_1	variable with f (18.693 to 19.097 μH)
L_2	L_1
L_{11}	$L_1 \cdot (1 - k)$
L_{22}	L_{11}
k	variable with f (0.4877 to 0.4995)
V_{DD}	40 V
T	$\frac{1}{f}$
pulse width	$D \cdot T$
delay time	$\frac{T}{2}$
rise time	$0.000001 \cdot T$
fall time	$0.000001 \cdot T$
R_{on}	10 $\mu\Omega$

That being said, some load resistance R_L values were chosen to evaluate the validity of the conducted theoretical analysis and consequent results for a fixed C_s value of 31 nF, specifically the frequency and duty factor pairs for each R_L taken from Fig. 4.17 being the used values presented in Table 4.4.

Table 4.4: Corresponding frequency and duty factor pairs for each R_L

R_L	Frequency	D
5 Ω	201kHz	37.06%
10 Ω	193kHz	34.14%
15 Ω	166kHz	36.57%
20 Ω	151kHz	39.74%
25 Ω	147kHz	41.21%

In the plotted simulation results of Fig. 4.21 it is possible to observe the differential switching voltage for the R_L values of 5 Ω , 10 Ω , 15 Ω , 20 Ω and 25 Ω .

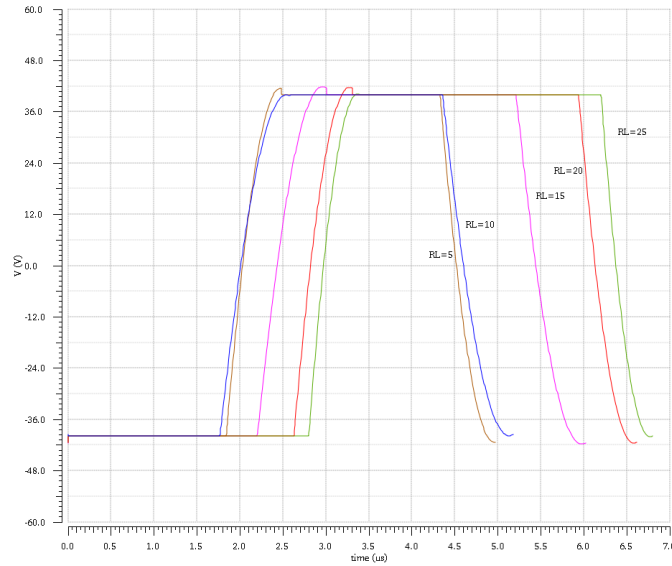


Figure 4.21: Differential switching voltage for different R_L values with the corresponding frequency and D

Even though all waveforms are fairly close to class-DE operation, it is possible to identify some R_L values at which the waveform had some distortion, namely at 5 Ω , 15 Ω and 20 Ω . At first sight, one could claim that such behaviour is caused by the presence of higher current harmonics, as it was the case in the simulations of section 4.3. However, observing the respective load current waveforms from Fig. 4.22(a), all currents appear to be close to a pure sinusoid and there are no observable differences between each of them in terms of signal purity. Such can be confirmed by observing and comparing the spectral distribution of one of the closer to class-DE waveforms with one of the further from class DE, say $R_L = 10\Omega$ and $R_L = 15\Omega$ for example, as depicted in Fig. 4.22(b) and Fig. 4.22(c).

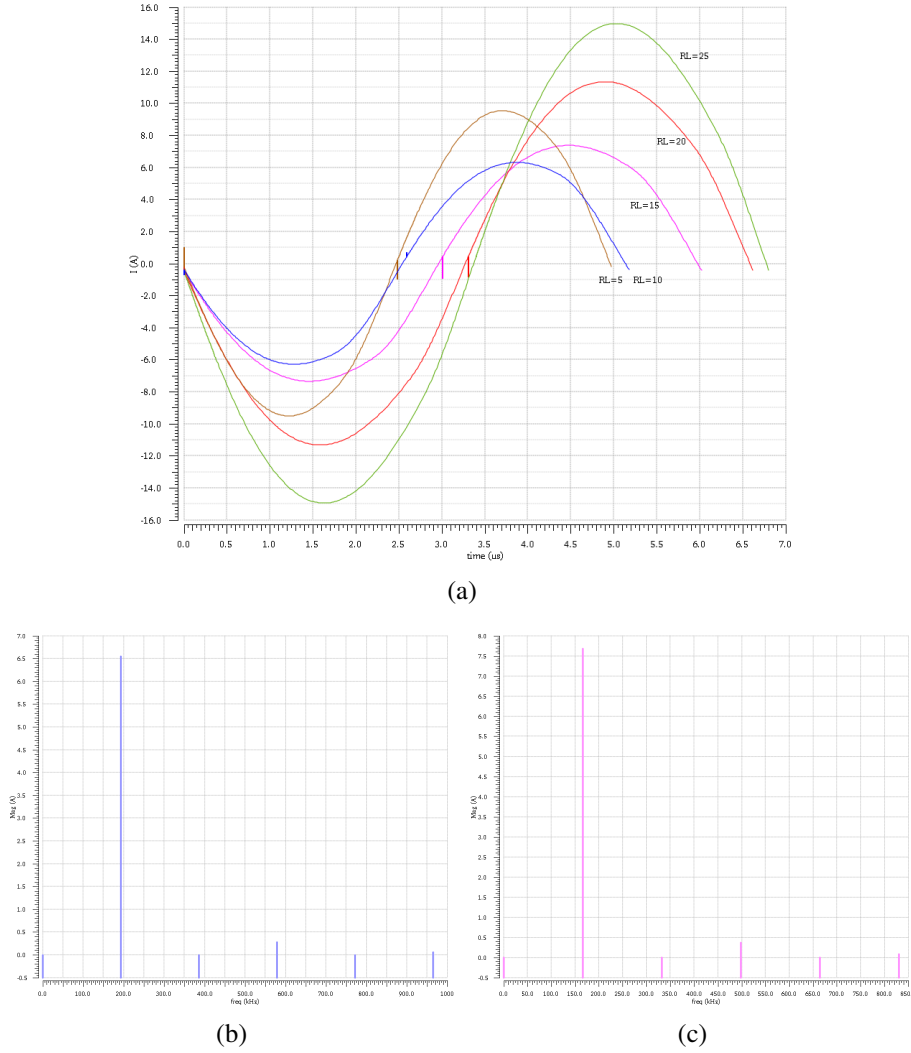


Figure 4.22: (a) - Load current for different R_L values with the corresponding frequency and D , (b) - Current harmonics for $R_L = 10\Omega$, (c) - Current harmonics for $R_L = 15\Omega$

That being said, it becomes evident that the cause for such differences is not related with the quality factor and, consequently, the harmonics. That way, one other plausible explanation would be the lack of resolution in the process of determining the frequency and duty factor pairs for each R_L value in the T-model, since it is a numerical process that requires approximations and is dependent on the resolution of the external data and variables. With that in mind, a parametric analysis was conducted in order to evaluate the behaviour of the less satisfying differential switching voltage waveforms with different frequency and duty factor values from the ones obtained in the numerical process. With such parametric analysis, it was possible to identify the frequency and duty factor value at which the differential switching voltage was as close to a class DE as possible. After this analysis, it was concluded that it was possible to correct the waveforms just by making slight corrections in frequency (less than 2kHz) and in duty factor (less than 1%), depending on the specific waveform. For example, as it can be seen in Fig. 4.23, the differential switching voltage

waveform for $R_L = 15\Omega$ becomes nearly perfect if the frequency gets from 166 kHz to 164 kHz and the duty factor gets from 36.57% to 37.5%.

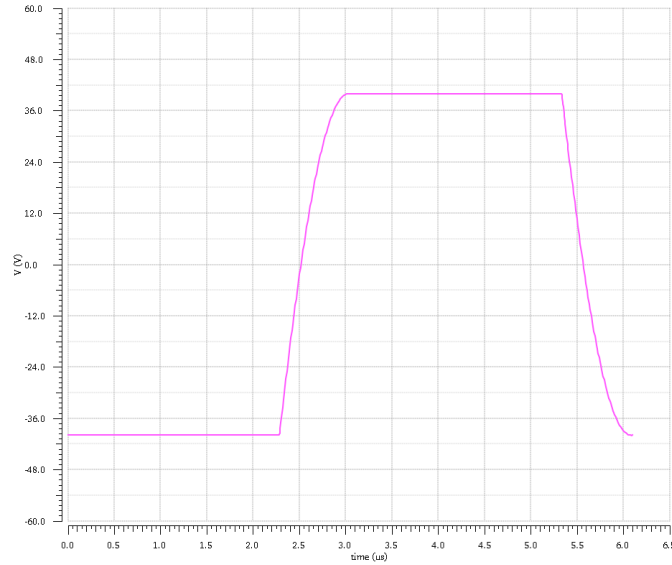


Figure 4.23: Differential switching voltage wave form for $R_L = 15\Omega$ with corrected frequency and duty factor

4.6 Summary

In this chapter, a theoretical approach to the power inverter to be developed was conducted, with its validity being confirmed through simulations. It was concluded that a class-DE operation is feasible in a wireless power transfer system, with the advantage of avoiding the addition of any extra coil for its realisation by operating at specific frequencies above resonance. Such results reveal to be promising for the intended increase in the whole system efficiency, as it will further be presented.

Chapter 5

System Realisation

In this chapter, the previously described theoretical analysis and simulations applied to wireless power transfer are put into practice. As such, the system design and implementation are explained, followed by a description of the carried tests and the presentation of the results. Finally, with the obtained results, an use case is presented.

5.1 Class DE in WPT Design and Implementation

In order to test and experiment the previous theoretical formulations, the system design and implementation were performed as will further be explained, being divided into Hardware and Software.

5.1.1 Hardware

Being a fundamental part of this thesis, the test board should be presented and explained. This board may be divided into different parts, namely the microcontroller and the measurement module, the low-dropout regulator (LDO) and protection circuit, the boost converter and the full bridge inverter. The schematics and complete system specifications may be found in the appendix. Additionally, the hardware also includes the coupled coils, the full bridge rectifier and the DC load.

5.1.1.1 Microcontroller and Measurements Module

The used microcontroller is the PIC32MZ1024EFE064, as can be seen in Fig. 5.1, which has the responsibility of controlling the full bridge rectifier through PWM signals generation, controlling the boost converter output voltage through a digital potentiometer, to acquire the voltages and currents measured by the measurements module, to communicate via universal asynchronous receiver transmitter (UART) with another device through a universal serial bus (USB) to serial UART interface chip (FT230XS) and to receive orders through the buttons connected to it. The implemented software will be described further in another subsection. For the full bridge control, four independent ports of the microcontroller are used, being each one connected to an operational amplifier (OPAMP) whose output is connected to a gate driver (UCC27714DR) that will turn on

and off the corresponding MOSFET. The gate drivers are powered at 10V, while the OPAMPs are powered at 5V and the microcontroller is powered at 3.3V. Further, regarding the measurements module, five different quantities are measured, namely the voltage and current supplied by the battery to power the system, the voltage and current outputted by the boost converter and the current outputted by the full bridge inverter. The voltage measurements are performed through resistive voltage dividers followed by OPAMPs (TSH82IYDT, presenting a large bandwidth, low distortion and good supply voltage rejection) that capture the signal and output it between 0V and 3.3V to one of the microcontroller ADCs. On the other hand, the current measurements are performed by Hall sensors (CQ-3303, chosen due to its high current range from -20A to 20A and to its large bandwidth, presenting a high accuracy for high frequency currents) that senses a magnetic field and outputs a voltage proportional to the current that creates it, being this voltage between 0V and 2.5V for negative currents and between 2.5V and 5V for positive currents. This voltage is then converted to a range of 0V to 3.3V with the medium value of 1.65V corresponding to 0A. The Hall sensors are powered at 5V and have a sensibility of 0.06V/A. Finally, regarding the digital potentiometer (AD5235) for the boost converter voltage control, its actuation is done through the serial peripheral interface (SPI) protocol by setting the desired resistance on it inside a range that goes to a maximum of 25k Ω .

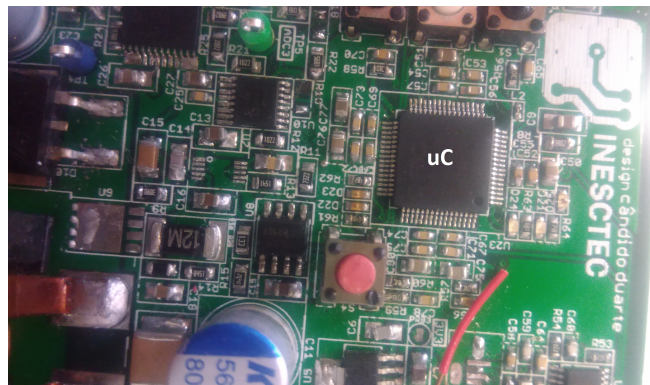


Figure 5.1: Microcontroller in the board

5.1.1.2 LDO and Protection Circuit

Moving on to the LDO regulation and protection circuit that can be seen in Fig. 5.2, its function is to transform the battery supplied voltage to 10V (UA7810CKTTR), 5V (14940D2T5-TR) and 3.3V (TPS79633DCQR), being these voltages used in different parts of the board, and to provide protection against overvoltage, undervoltage and reverse supply protection through a chip (LTC4365) that makes this control by actuating in a dual switch (SQJB60EP-T1_GE3) at the input provided from the battery.

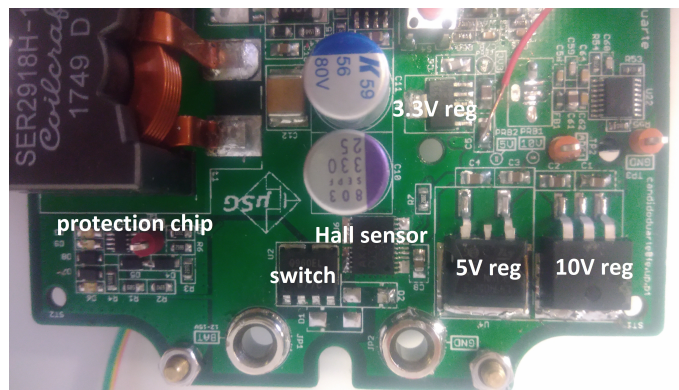


Figure 5.2: LDO and protection circuit in the board

5.1.1.3 Boost Converter

Regarding the boost converter, that can be seen in Fig. 5.3, its function is to step up the battery input voltage to an appropriate value for a significant power transfer level. It is comprised by a coil (SER2918H-153KL), a MOSFET (BSC057N08NS3GATMA1), a diode (STPS20M100SG-TR), a high voltage low-side MOSFET controller (LM5022QDGSTQ1) to control the converter operation and keep its output voltage in a defined reference and a low pass filter in the output to reduce the ripple caused by the commutation of the switch. Such reference is defined through a resistive voltage divider that transforms the output voltage value in a smaller value to be read by the feedback pin of the controller. The controller will then actuate in the output voltage in order to maintain the feedback pin voltage equal to a 1.25V reference. This way, by setting a specific resistive voltage divider, knowing that the feedback pin will tend to 1.25V, it is possible to calculate the what the output voltage value will be. With that in mind, the resistive voltage divider must be defined to obtain the desired output voltage.

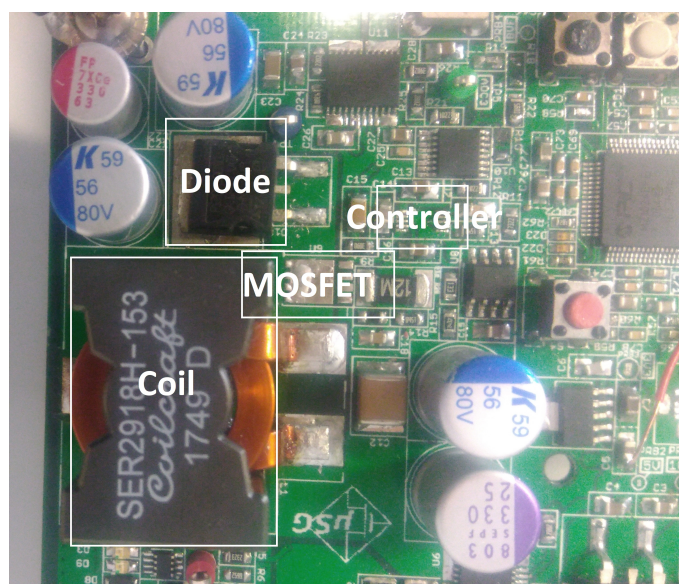


Figure 5.3: Boost converter in the board

5.1.1.4 Power Inverter

Now, concerning the power inverter that can be seen in Fig. 5.4, which is the main scope of this work, it is a regular full bridge inverter comprised by four MOSFETs (CSD18536KTT) with the only difference from a traditional full bridge inverter being two shunt capacitors added in parallel with the bottom switches of each leg of the inverter to realise the class-DE operation. As explained previously, it is equivalent to reduce the two capacitances of each inverter leg to one single shunt capacitance (which from now on be referred as C'_s or as $2C_s$) in each of the bottom switches with its value doubled relative to the original one, since the voltage and current on the top capacitance is equal to the ones on the bottom capacitance. The function of the power inverter is to receive a DC voltage and to transform it in an AC voltage by turning on pairs of diagonally opposed switches while keeping the other pair off.

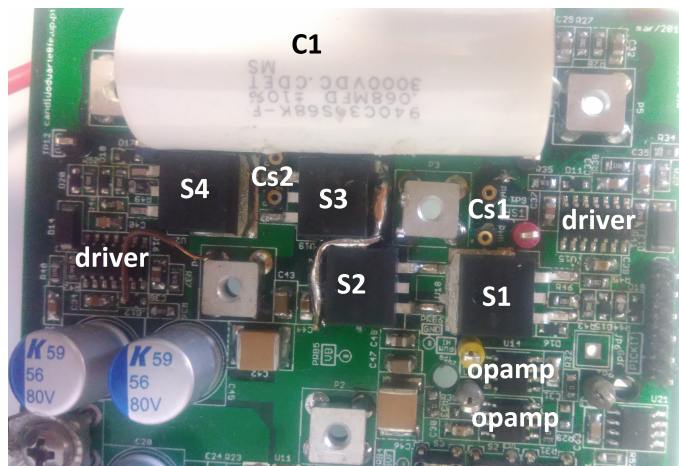


Figure 5.4: Power inverter in the board

5.1.1.5 Coupled Coils

After the inverter there are two magnetically coupled coils, with a capacitor in series on each side to allow the existence of resonance. The receiver and transmitter coils and capacitors are identical in their construction and characteristics. As it can be seen in Fig. 5.5, the coils are shaped to resemble an AUV surface and the docking station. The coils characterisation was previously presented in chapter 3.

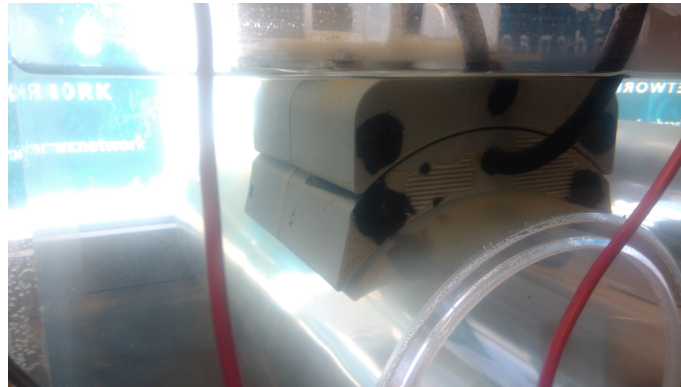


Figure 5.5: Encapsulated coils in the water tank

5.1.1.6 Rectifier

On the secondary side, following the secondary coil and series capacitor, there is a full bridge rectifier comprised by four diodes (FSV10100V) as can be seen in Fig. 5.6, whose function is to convert the AC power signal from the coil into a DC power signal to the load. A low pass filter is placed after the rectifier in order to reduce the ripple of the signal, making it closer to a pure DC voltage.

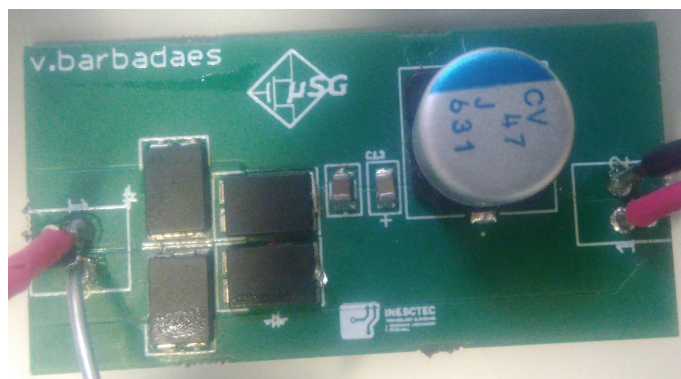


Figure 5.6: Rectifier board

5.1.1.7 Load

Finally, at the end of the circuit, there is a resistive load whose function is to consume the transferred power. In this case, the load is represented by a rheostat (BFA522R) that converts the electrical power into heat, and has the advantage of being variable on a 22Ω range, which allows further load sweeps in the tests procedures.

5.1.2 Software

In order to have the microcontroller performing the tasks described before, a software application had to be implemented for the specified hardware.

5.1.2.1 Initialisation and Clock Configuration

First of all, beyond the microcontroller specific default configurations, some of them are set differently. Those mainly include the configurations that regard the system oscillator and clock frequency, as depicted in Fig. 5.7. First, a system oscillator has to be selected, which in this case was the system phase locked loop (SPLL). The frequency of the SPLL would then by default be divided in half of its value to originate the peripheral bus clock, peripheral bus clock (PBCLK). Then, the SPLL needs to be defined. As such, the internal Fast RC Oscillator is selected, having a frequency of 8MHz, being then the PLL Input Divider set as 1, the PLL Multiplier set as 128, the PLL Range set as 8MHz to 16MHz input and the PLL Output Divider set as 4. That configuration will then result in a SPLL frequency of 256MHz and a PBCLK frequency of 128MHz. A high PBCLK frequency value was required in this application due to the high frequencies of the inverter (reaching values around 200kHz) and due to the need of high accuracy when setting the duty factor of the control signals, since differences of only 1% in D are a cause for different the behaviours.

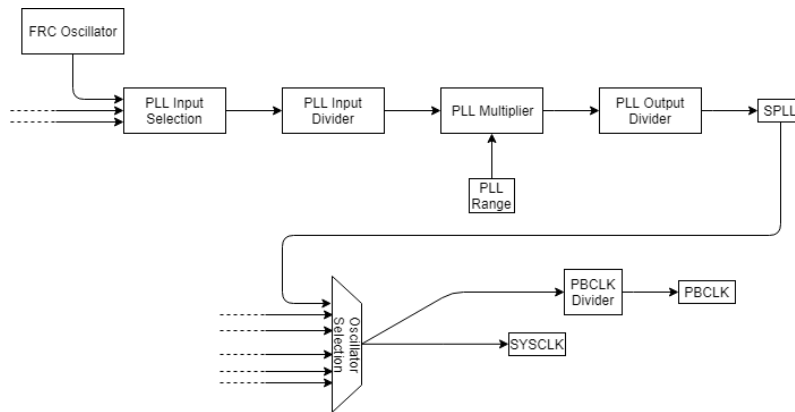


Figure 5.7: Clock configuration diagram

Moving on to the system operation software, being a software designed for testing different states, it can be said that it is an interactive open loop software since it only requires inputs from the user and does not present any kind of control feedback. Once the microcontroller is turned on, the software makes the basic initialisations, such as the previously mentioned configurations as well as the ports setting (inputs or outputs and digital or analog) and the UART communication initialisation. Then, the software waits for the clicking of an external button actuated by the user that allows it to initialise the rest of its modules, such as the SPI, the timers, the ADCs, the interrupts and the PWM generation, and to start operating at a predefined state.

5.1.2.2 Terminal

Once all the initialisations are performed and the system starts working in the default state, it becomes possible for the user to interact with the system through a terminal. In this terminal,

the user has the possibility of send several commands and to receive information. The list of commands is presented in table 5.1.

Table 5.1: List of commands

Command	Description
help	Prints to the terminal a list with the existing commands with utilisation examples.
pwm xxxxxx xxx	Allows the user to set the PWM configuration at any time, namely its frequency and duty factor, as well as to turn it off without turning off the rest of the system. This command is sent by writing into the terminal the word "pwm" followed by the frequency in Hz and the duty factor in per mille.
adc xxxxx	Makes the system return the measured DC current and voltage in the supply from the battery or in the boost output, also returning the power calculated with these measures, depending on whether the word "adc" is followed by "bat" or "boost".

5.1.2.3 Implemented Functions

Now, moving on to a detailed description of the most relevant implemented functions, the hardware initialisations will first be addressed.

UART

Regarding the UART configuration after remapping its receiver and transmitter pins to the corresponding pins used in the schematic, which is only possible if the control register is unlocked and locked again in the end, the Baud Rate generator register U5BRG is set using $U5BRG = \frac{PBCLK1}{16 \cdot \text{BaudRate}} - 1$, being the used Baud Rate 9600 bps. Then, the receiver interrupt is enabled to be used in the terminal with the interrupt on first byte in FIFO (first in first out), being the transmission and reception enabled and the UART turned on. In order to print data on the terminal, that data must be converted in a char vector and single characters are sent to the transmit buffer until the end of the vector.

ADCs Initialisation

Now, concerning the five ADCs initialisation, after configuring the respective ports, the following steps are executed for each ADC module:

- The ADC calibration data is copied through the DEVADCx register and all control registers features disabled;
- The warm up time register is initialised and set to 32 clocks to warm up the ADC;
- The Peripheral Bus Clock source is selected and the general ADC clock is set to be half of the input clock, being each ADC module clock multiplied back by two, which in this case results in an ADC clock frequency of 128MHz. The sampling time is set to be twice the ADC clock period, resulting in a sampling time of 15.625ns and the ADC resolution is set to be of 12 bits;

- The analog inputs are selected and configured to use unsigned data mode and single ended mode;
- The interrupts are enabled; and the ADC is configured to not to use early interrupts, scanning, digital comparators and oversampling filters;
- The ADC is set to use edge trigger and triggers from software and is turned on and waits until the reference voltage is ready, waits to determine if there is a fault with the reference voltage, enable clock to analog bias, wait until the respective ADC module is ready and enables that ADC module.

PWM

Looking now at the PWM configuration and initialisation function, which has as inputs the frequency, the duty factor and the deadtime, after remapping its receiver and transmitter pins to the the corresponding pins used in the schematic, which is only possible if the control register is unlocked and locked again in the end, the comparators are disabled and Timer2, which will drive all four comparators, is configured using the frequency received as input. Such is done by setting the counting period $PR2 = \frac{PBCLK1}{frequency} + 1$ and using a 1 : 1 prescale. The following steps are then executed for each of the independent PWM generators:

- The control registers features are disabled and the comparator is configured for continuous output pulses;
- The comparators rising and falling edges are configured as shown in table 5.2;
- Timer2 and the comparators are enabled.

Table 5.2: Comparators configuration

Comparator	Rising Edge (OCxR)	Falling Edge (OCxRS)	Controlled Switch
OC9	$OC9R = 0$	$OC9RS = PR2 \cdot dutyfactor + 1$	top left
OC6	$OC6R = OC9RS + PR2 \cdot deadtime$	$OC6RS = OC6R + PR2 \cdot dutyfactor + 1$	bottom left
OC1	$OC1R = 0$	$OC1RS = PR2 \cdot dutyfactor + 1$	bottom right
OC3	$OC3R = OC1RS + PR2 \cdot deadtime$	$OC3RS = OC3R + PR2 \cdot duty_{cycle} + 1$	top right

SPI

For the communication with the digital potentiometer, the corresponding SPI module had to be configured. As such, after remapping its receiver and transmitter pins to the corresponding pins used in the schematic, which once again is only possible if the control register is unlocked and locked again in the end, the following steps are executed:

- The SPI module control register and receive buffer are cleared, the clock frequency is configured;
- The input data is configured to be sampled at the end of data output time;
- The master mode is selected, the clock signal is set to be active high, the 32 bit mode is set and the SPI module is turned on.

Digital Potentiometer

Now, to send data to the digital potentiometer via SPI, there is a function that receives as an input the desired resistance. Then, using the address defined by the manufacturer to set the resistance of one of the potentiometer variable resistors (there are two), the received resistance value is sent using the conversion $SentValue = \frac{MaxResistance - Resistance - MinResistance}{\frac{MaxResistance - MinResistance}{1024}}$, being $MaxResistance$ and $MinResistance$ measured values obtained by setting the resistance to its maximum and to its minimum of 24840Ω and 98.9Ω , respectively. The digital potentiometer resolution is 10 bits, having 1024 resistance positions. After sending the resistance value with the respective address, this value is stored in nonvolatile memory using another specific address. When a request for a change in the resistance value is received by the function, instead of directly setting its value to the desired final value, it does so with 100Ω steps in order to avoid abrupt changes.

ADCs Parsing

Finally, besides other minor functions such as timers definitions and ports initialisations, there are the ADCs parsing functions. As described before, the ADCs conversions are triggered by software. Such is done using a timer that will periodically cause an interrupt that in its routine creates a trigger for the ADCs. After the the conversion, an interrupt will happen for each ADC module, which fetches the result and turns on a flag indicating that new data is available. Once the application loop detects one of these flags, it calls the respective function which returns the corresponding current or voltage value. In this function, first, the received binary data value is converted into a decimal value equal to the voltage acquired on the ADC analog port using $VoltageValue = BinData \cdot \frac{3.274}{4096}$, where $3.274V$ is the measured value in the $3.3V$ and used in the conversion for better accuracy, and 4096 is used due to the 12 bit resolution. Then, the resulting measured voltage in the ADC input must be converted into the the real quantity to be measured, either a current from a Hall sensor or a voltage from a resistive voltage divider and/or and OPAMP. The different conversions are listed on table 5.3.

Table 5.3: ADCs conversion formulas

Battery Voltage	$VoltageValue \cdot \frac{66500 + 232000}{66500}$
Battery Current	$VoltageValue \cdot \frac{39000 + 20000}{39000} - 2.462$
Boost Voltage	$VoltageValue \cdot \frac{21600 + 470}{470} \cdot \frac{13000 + 1000}{13000 + 1000 + 24280}$
Boost Current	$VoltageValue \cdot \frac{39000 + 20000}{39000} - 2.462$

The battery voltage formula is defined simply by using the voltage divider resistances, which are $66.5k\Omega$ and $232k\Omega$. The battery current and boost converter current are defined using the

voltage divider with $39k\Omega$ and $20k\Omega$ and subtracting the measured 0A corresponding voltage value 2.462V to eliminate the offset, being all divided by the Hall sensor sensibility of 0.06 V/A. The boost converter voltage is defined first using the voltage divider with $21.6k\Omega$ and 470Ω , being then used the OPAMP non-inverting gain, as shown in Fig. 5.8.

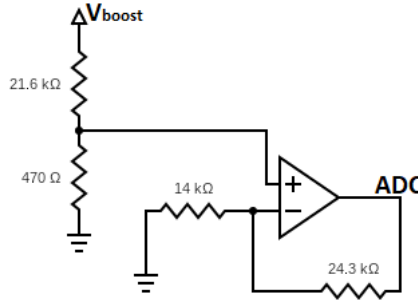


Figure 5.8: Boost voltage ADC configuration

5.2 Tests and Results

From the previously described implementation, a set of measurements was conducted in order to evaluate the system performance in terms of efficiency and transmitted power as well as to compare the experimental behaviour with the theoretical predictions obtained through the electrical model of the system.

5.2.1 Setup

The tests consisted in the measurement of the currents and voltages in four different points of the circuit in order to obtain the corresponding Power and impedance of each part. These were the DC input voltage and current, the differential switching voltage and primary coil current, the secondary coil voltage and current and the rectified voltage and current. For such measurements, a power analyser was used. Additionally, an oscilloscope was also used to observe the differential switching voltage waveform as well as the primary current using an amperometric clamp.

For the setup, which was designed as shown in Fig. 5.10, a small tank containing the encapsulated coils was filled with salted water, with a salt concentration of around 35 g/L, completely submerging both coils. Further, the necessary electrical connections were done, namely the connection of the inverter to a voltage source, the board to the primary coil, the secondary coil to the rectifier, the rectifier to the load, whose role was played by a rheostat, and the voltage and current measurements connection to the power analyser. **It should be mentioned that the original setup did not take in a**

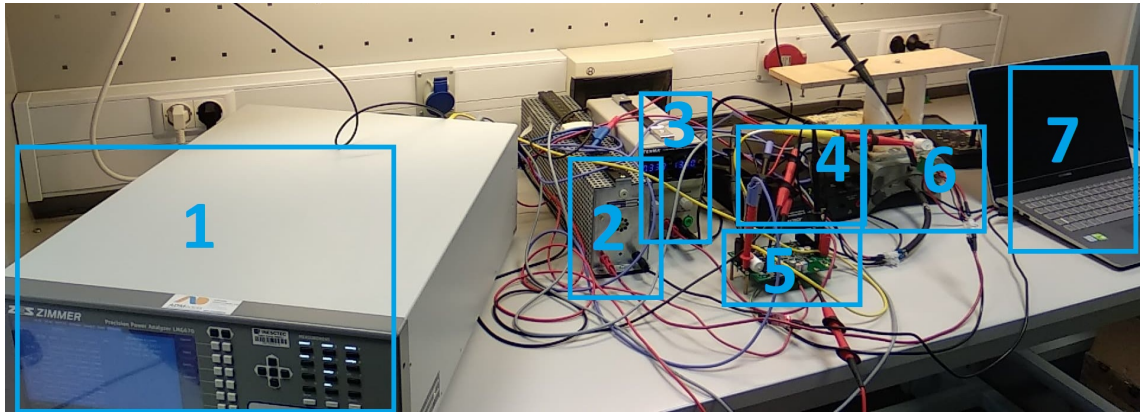


Figure 5.9: Experimental setup without the salted water tank. 1 - Power analyser, 2 - rheostat, 3 - DC voltage source 1, 4 - DC voltage source 2, 5 - transmitter board, 6 - coils, 7 - computer

The effect of the connecting cables length and the results were getting far from the expected, to a point where the resonant frequency was well below the original predicted one of around 143kHz. With that in mind, the setup was reconstructed with every cable that was carrying AC current, namely the AC current measurement cables and the coils connecting cables, being shortened as much as possible to reduce the strong inductive effects previously detected. With this measure, the new results shifted to a region where they were expected to be and the tests could be considered to be more reliable.

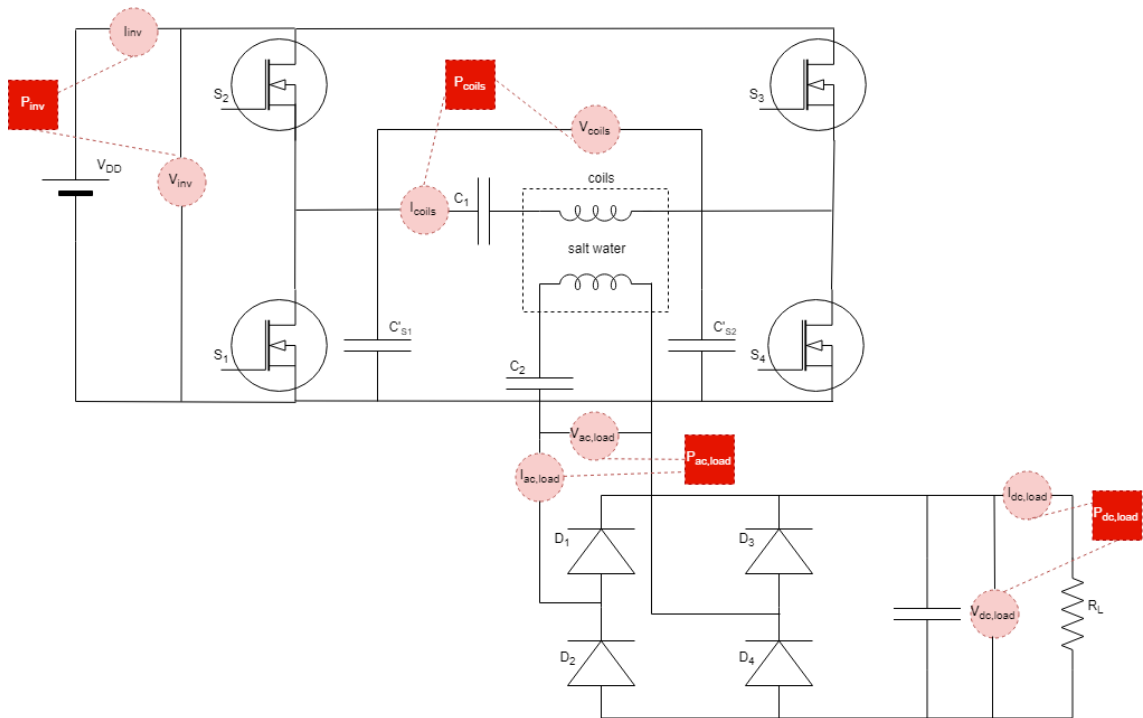


Figure 5.10: Circuit design for experimental validation [33]

Additionally, as described before, the tests were performed connecting a voltage source directly to the inverter instead of using the boost converter. This option was taken for practical, since the scope of these tests was to evaluate the behaviour and the performance of the power inverter, and its relation to what was in its output, namely the coils, rectifier and load, which eliminates the need of using a boost converter for the tests in question. That being said, two different voltage sources were connected to the board. One supplying 13V to the control circuit and other supplying a fixed value of 22V for the execution of the following tests.

Following, some components were measured to fully characterise the system, since the used theoretical model does not have in account these values. That being said, the added shunt capacitors $C'_s = 2C_s$ were measured at different frequencies, namely the minimum used frequency of 150kHz and the maximum used frequency of 200kHz, using an LCR meter, and the series capacitance and equivalent series resistance (ESR) were determined as shown in table 5.4.

Table 5.4: Measured parameters of added shunt capacitances at around minimum and maximum used frequencies

C'_s	Measured C'_s ($f = 150\text{kHz}$)	ESR ($f = 150\text{kHz}$)	Measured C'_s ($f = 200\text{kHz}$)	ESR ($f = 200\text{kHz}$)
10 nF	9.8 nF	1.9 Ω	9.67 nF	1.42 Ω
22 nF	19.99 nF	1.3 Ω	19.66 nF	0.881 Ω
47 nF	46.68 nF	0.028 Ω	46.89 nF	0.031 Ω
62 nF	61.8 nF	0.057 Ω	62.4 nF	0.048 Ω

Additionally, the same procedure was taken to measure the used resonance capacitors, being the results presented in table 5.5.

Table 5.5: Measured parameters of resonant capacitances at around minimum and maximum used frequencies

Resonant Capacitor	Measured C_{res} ($f = 150\text{kHz}$)	ESR ($f = 150\text{kHz}$)	Measured C_{res} ($f = 200\text{kHz}$)	ESR ($f = 200\text{kHz}$)
C_1	66.1 nF	13.6 m Ω	66.2 nF	13.0 m Ω
C_2	65.9 nF	13.2 m Ω	66.0 nF	13.0 m Ω

5.2.2 Test Procedure

The testing of the system was done and repeated for each $2C_s$ value through the following procedure: after selecting a specific $2C_s$ value, a range of D values had to be defined. To define such list, an initial measurement of a switching node voltage with the oscilloscope was made and the maximum and minimum D values at which the waveform seemed to be similar to a class DE (like in Fig. 5.11) with any R_L were defined as the maximum and minimum D values for the measurements list range. After this, the load was set to its maximum value and the first D value of the list was sent to the microcontroller, via UART, with the smallest frequency value at which was possible to observe a similar to class DE (like in Fig. 5.11) waveform with that maximum R_L . Having

found and recorded this frequency value with the power analyser, the D was maintained and the frequency was increased in multiples of 5kHz at each iteration. For each frequency iteration, the load was manually reduced until a waveform similar to class DE (like in Fig. 5.11) was observed in the oscilloscope and the state was recorded with the power analyser once again. The frequency was increased until it was no longer possible to observe a waveform similar to class DE (like in Fig. 5.11) by reducing the load value. At this moment, the D was increased by 1% and the whole process was repeated.

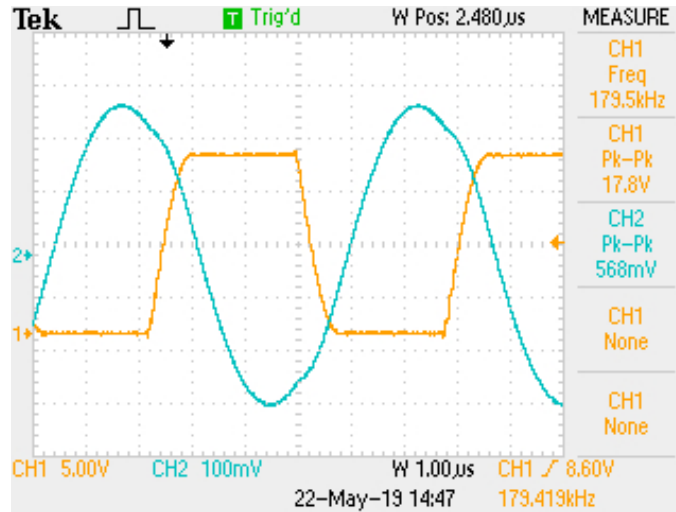


Figure 5.11: Oscilloscope print from one of the switching nodes and the current at the primary side with the system operating in class DE

Finally, for comparison, similar measurements were made with the inverter operating as a class D in the resonant frequency of around 143kHz. In this case, the only variable value was the load, R_L , and all the previous measurements were made once for each R_L .

5.2.3 Output Power

From the previously described tests, a large amount of points were collected for further data analysis, which could be conducted from different perspectives. That being said, a very relevant output of the previous tests is the power delivered to the DC load, after the rectifier, since a good power capability is an important requirement to be fulfilled in a system such as this one. Thus, knowing from the previous theoretical analysis that a class-DE operation will have its output power capability reduced when compared to a class D and even more for reduced duty factors, the following results are important to evaluate in which degree does that happen. First, just as a straightforward analysis of the duty factor influence, Fig. 5.12(a) shows a clear increase in the output power as the duty factor approaches 50%, with the higher values belonging to the class-D operation, which works practically on top of $D = 50\%$. Additionally, in Fig. 5.12(b) it is possible to observe that the output power tends to be higher for frequencies closer the resonant frequency and, after decreasing with increasing frequencies, it starts increasing again. Such behaviour is related to the

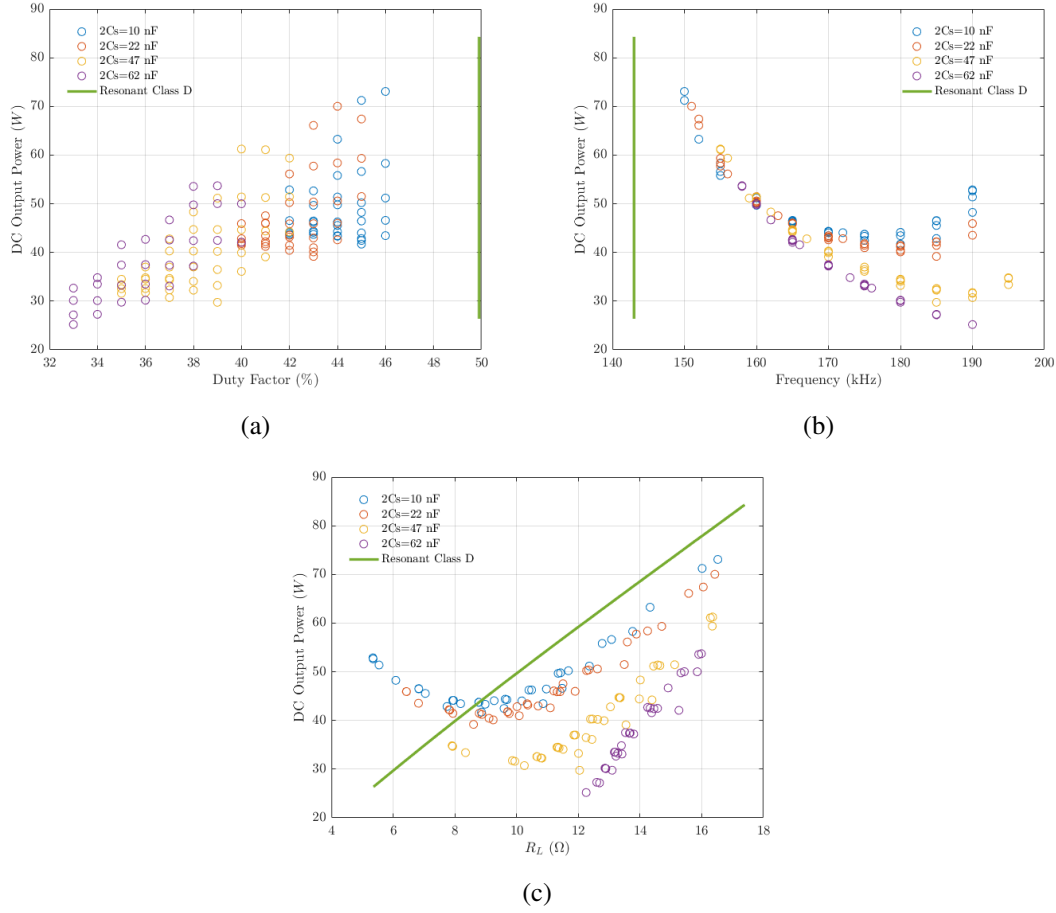


Figure 5.12: Output power for each tested C_s value in function of (a) - the duty factor, (b) - the frequency, (c) - the AC R_L .

results from Fig. 5.12(c), since the higher amount of power happens for higher load values, which correspond to smaller frequency values, and the same happens to smaller load values, oppositely from the class-D behaviour. It can be seen that for AC loads lower than 9Ω the class-DE power is higher than the class D, which was not expected. Such behaviour is explained by the fact that the class DE uses much higher operating frequencies in that load region, whereas the class D operates on top of the resonant frequency in the entire load range. For that reason, and as it is confirmed in Fig. 5.12(b), operating in higher frequencies allows the class DE to be in a frequency region where the coupled coils present a high power capability.

The obtained output power values, reaching around 75W were measured with a $V_{DD} = 22V$ and could easily be raised with small V_{DD} increases, since there is a quadratic relation between the two, as described in 4.17. For example, some tests have been done with $V_{DD} = 30V$, obtaining output power values above 120W. One should take in consideration that the used C_s values were dimensioned for relatively high duty factors, meaning that other C_s values could lead to a severe loss in output power capability.

5.2.4 Efficiency

The next output to be analysed is the system efficiency, being the goal of this thesis **was** to study and improve it. Inside this system, at least four different efficiencies may be accounted and analysed, namely the inverter efficiency, the coupled coils efficiency, the rectifier efficiency and the whole system efficiency, which is the product of all the previous ones, accounting for all losses and relates the input DC power with the output DC power. Being this work concerned with the inverter optimisation, it is obvious that the inverter efficiency is one of the most relevant too be studied. Further, since that the whole system efficiency relates the inverter efficiency with the rest of the system efficiency, it was considered to be of high relevance along with the inverter efficiency. That being said, the following presented results will account all four and will be analysed relative to different variables, such as the shunt capacitance, the duty factor, the frequency and the AC load. As it was shown before, the following results ere obtained with the delivered power being in a range of around 20W to 80W.

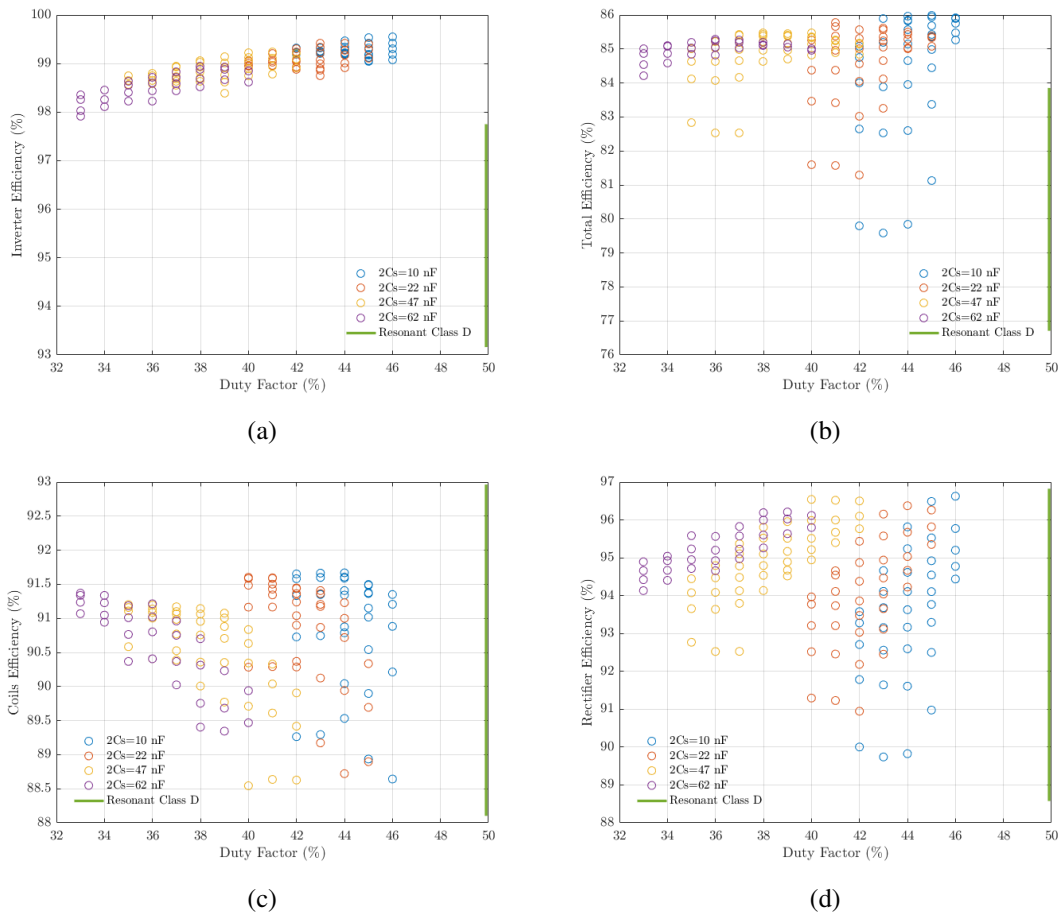


Figure 5.13: Efficiency for each tested C_s value. Different points for each D were obtained with different frequencies and loads. Class D points were obtained at the resonant frequency with different loads in the same range (a) - Inverter efficiency, (b) - Total system efficiency, (c) - Coupled coils efficiency, (d) - Rectifier Efficiency

Starting by analysing the inverter efficiency dependent on the duty factor plotted in Fig. 5.13(a), it is possible to observe a slight increase in efficiency as D increases, even though all values are fairly high, between 98% and almost 100%. Such behaviour may be explained as a result of the losses caused by the current flowing in the shunt capacitances internal resistance, which are higher when their conducting time is higher, i.e. the deadtime is higher and the duty factor is lower. This plot also shows that all class-DE inverter efficiency points are higher than the class-D ones (represented by the green line), not allowing additional conclusions related to the class DE and class D comparison though. Looking now to Fig. 5.13(b), there is still an increase in efficiency as D increases, not being so clear as before however, and being the maximum efficiency values for each D between 85% and 86%. This increase is less evident here because since all efficiencies are accounted, the inverter efficiency will have less relevance, being the differences attenuated. It is also possible to observe that not all whole system efficiencies are higher in class DE, with no further conclusions obtainable here. Additionally, Fig. 5.13(c) and Fig. 5.13(d) show how the coupled coils efficiency and the rectifier efficiency vary with different duty factor, being rather inconclusive here, being necessary to observe them in other perspectives, as will be presented further.

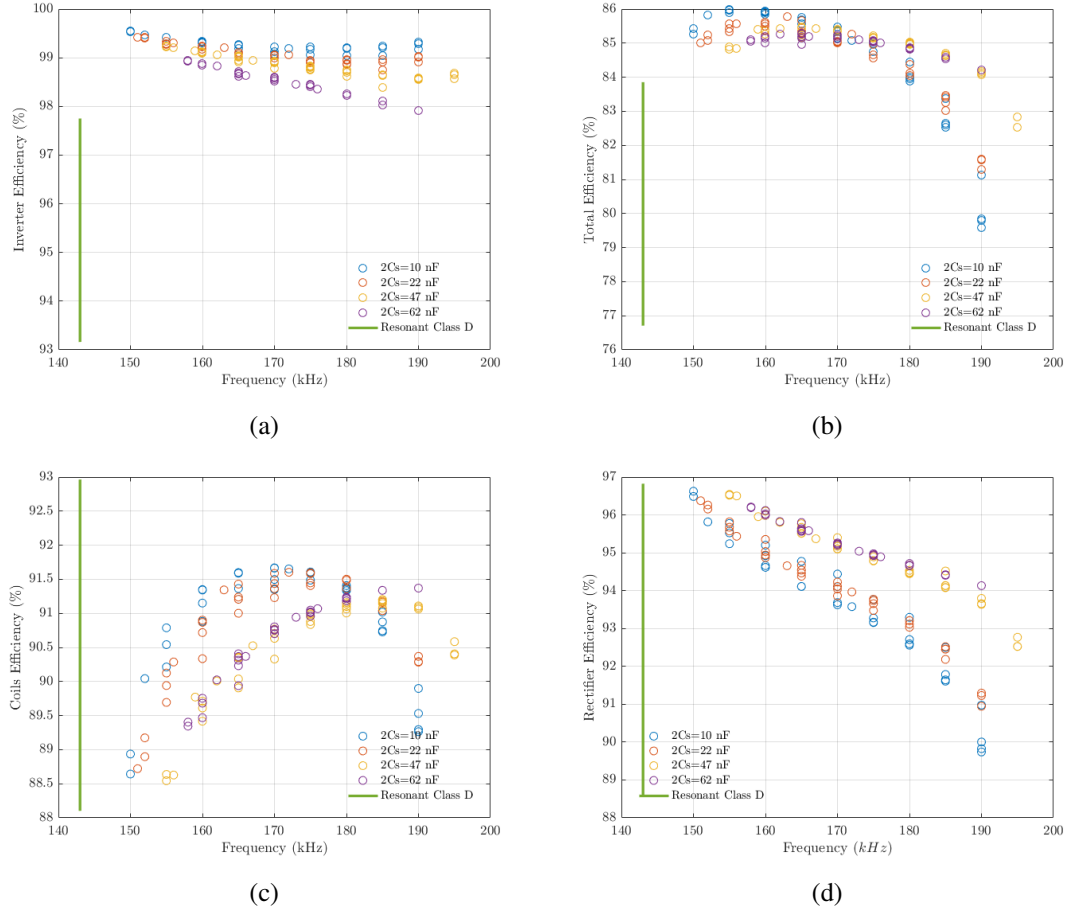


Figure 5.14: Efficiency for each tested C_s value. Different points for each frequency were obtained with different duty factors and loads. Class D points were obtained at the resonant frequency with different loads in the same range (a) - Inverter efficiency, (b) - Total system efficiency, (c) - Coupled coils efficiency, (d) - Rectifier Efficiency

Now, observing the results from Fig. 5.14(a), one can observe the dominance of the smaller C_s values in the entire range of frequencies, with slightly higher inverter efficiency at lower frequencies. Such dominance is related with the previous plots from Fig. 5.13, since the smaller shunt capacitances correspond to the higher duty factors. Further, analysing the whole system efficiency from Fig. 5.14(b), the dominance of smaller C_s values becomes less obvious and does not happen in the entire range of frequencies. Additionally, it is possible to see that the frequency dependence is bigger, with a maximum between 155kHz and 160kHz. Such maximum region is related to the results plotted in Fig. 5.14(c) and Fig. 5.14(d), relative to the coupled coils efficiency and the rectifier efficiency, respectively. In terms of coupled coils efficiency, it is very evident that there is a maximum area, around 170kHz, whereas the rectifier efficiency grows continuously as the frequency is decreased. Such combination of all efficiencies result in the aforementioned maximum area in the whole system efficiency.

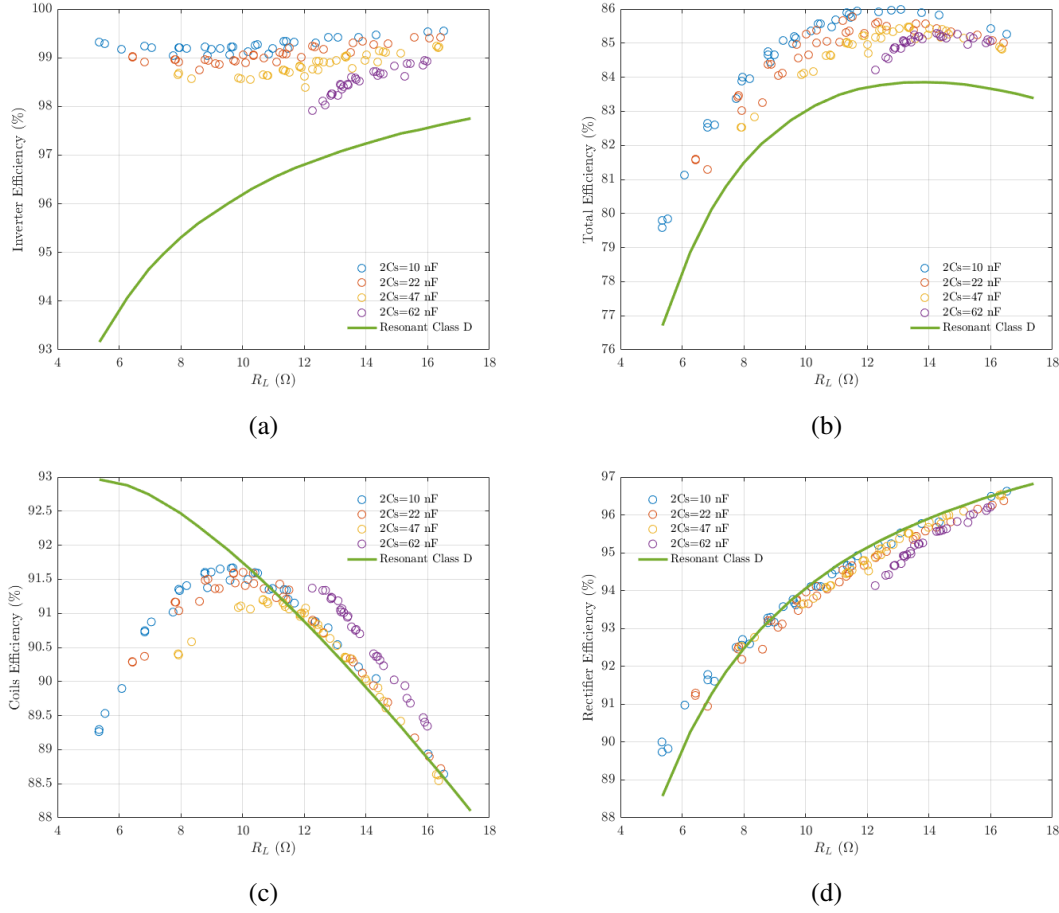


Figure 5.15: Efficiency for each tested C_s value in function of the AC R_L . Class D points were obtained at the resonant frequency with different loads in the same range (a) - Inverter efficiency, (b) - Total system efficiency, (c) - Coupled coils efficiency, (d) - Rectifier Efficiency

Taking now in consideration the variation of the inverter efficiency with the AC load R_L in Fig. 5.15(a), besides observing that the smaller shunt capacitances remain consistently dominant from the previous plots, it is also possible to compare the class-DE performance with the class D in a fair way, with the class DE having an obvious superiority over the entire range of loads. Such superiority is confirmed in Fig. 5.15(b) by the whole system efficiency results. In the end, one can say that this system with class DE can reach total efficiencies up to 86%, while with class D it reaches a maximum of lower than 84%. The next two plots in Fig. 5.15(c) and Fig. 5.15(d) are now rather interesting. Looking at the coupled coils efficiency behaviour, one would say that the best region to operate would be around 9 Ω or 10 Ω , as it was previously predicted when analysing the T-model parameters in chapter 3.1. However, observing the results in the rectifier efficiency, it is possible to understand why the total system efficiency was shifted to higher loads. In this rectifier, the load differences had a too high influence on its efficiency and, with lower loads corresponding to lower voltage in the secondary, the forward voltage drop in the diodes becomes more significant compared to the transmitted voltage and the dissipated power is higher when compared to the transmitted power.

Finally, using a thermal camera, it was possible to analyse the different heat production on the switches when operating in class DE and class D. Such heat is caused by the losses on the switches, and it can be seen in Fig. 5.16(a) that it is almost nonexistent in class DE, while it is fairly high in Fig. 5.16(b), referent to a class-D operation. It is however observable that in class DE again in Fig. 5.16(a), a significant amount of heat is dissipated by the shunt capacitors, meaning that they also might represent some relevant losses.

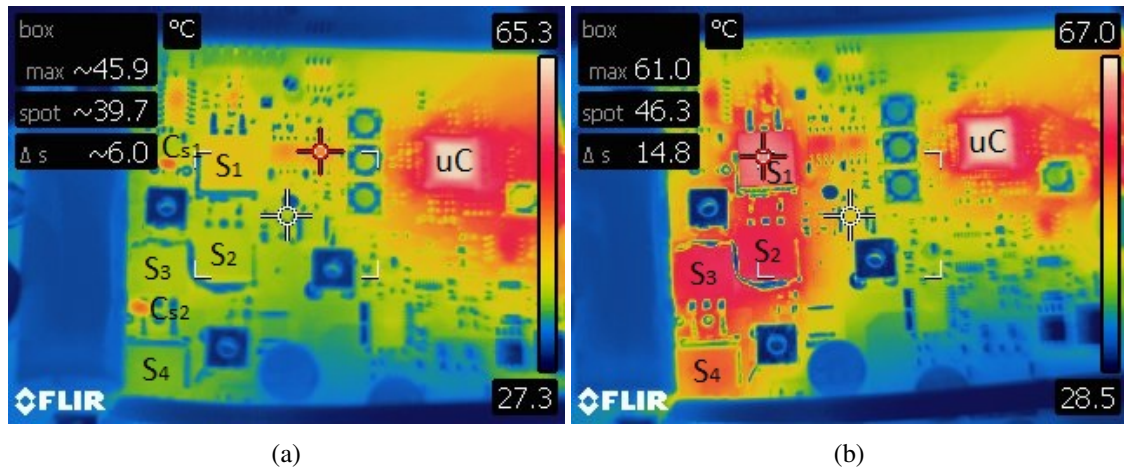


Figure 5.16: Thermal pictures of the inverter (a) - in class DE, (b) - in resonant class D

5.2.5 Experimental Results vs. Theoretical Predictions

Finally, even though some discrepancies are expected between the predicted theoretical class-DE working points and the experimentally measured ones due to the many non ideal component parameters like the non linear MOSFETs capacitances, their non instantly rising and falling times, their conducting resistance R_{on} , the added capacitors equivalent series resistance, the non purely sinusoidal load current, just to list a few, and even the T-model parameters possible inaccuracy, it is still interesting to evaluate how similar both results are.

One should also note that the class-DE test procedure does not comply with the fact that theoretically there is only one frequency and duty factor pair for a given C_s and R_L . Actually, as described before, the tests for each C_s were made by sweeping the load with a fixed D and changing frequencies, and repeated for many D values in the same load range. This means that for each R_L there were various frequency and duty factor pairs. Such was the case because there was no measurable metric to evaluate if the system was in theoretically pure class-DE or not, being that evaluation done through the looks of the differential switching voltage waveform, as described before. That being said, the different frequency and duty factor pairs for the same load may not be pure class-DE, because there can only be one, but were very close approximations instead.

Now, looking at the comparison in Fig. 5.17(a) showing the predicted frequency compared to the real used one for two different C_s values, one can affirm that the overlapping was very satisfactory, surpassing expectations.

Further, relative to the comparison between duty factors in Fig. 5.17(b), the experimental points distribution is as expected, with some values being further away from the prediction, but with the average being more or less coincident with it.

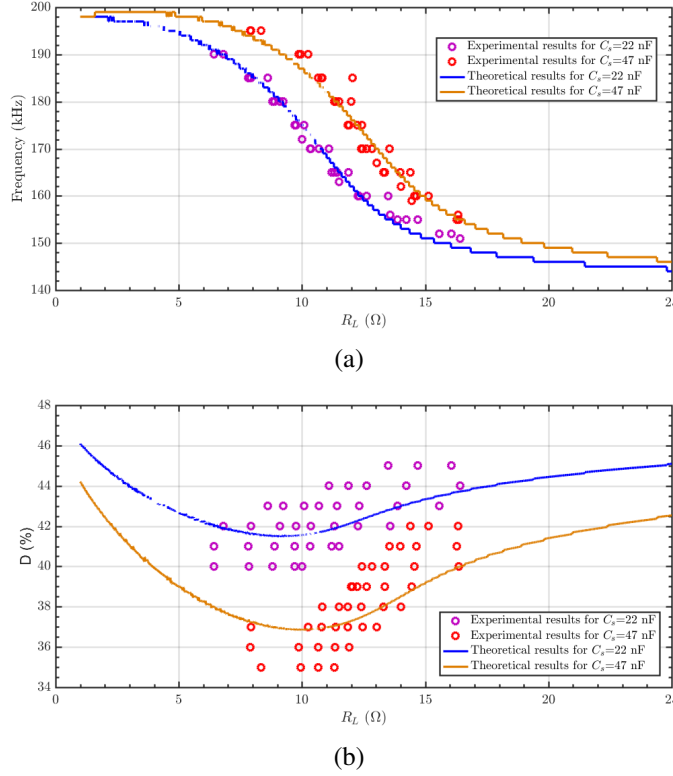


Figure 5.17: Comparison between the calculated and the experimental (a) - frequency and (b) - duty factor pairs for $2C_s = 22$ nF and $2C_s = 47$ nF

The reason why the frequency predictions were better satisfied than the duty factor ones may be explained by the fact that the duty factor has a smaller impact on the class-DE behaviour and its alteration was not so easily observable in the waveforms as it was with the frequency, especially since the used duty factors for each C_s are defined in such a small range like, for example for $2C_s = 22$ nF, the theoretical variation in the entire load range is only of about 5%,

5.3 Use Case

Through the results analysis it is now possible to design use cases with a good degree of accuracy. As an example, knowing now that the whole system efficiency is maximised for AC loads around 12Ω , an use case where a power value of 80W is required to be delivered to the DC load, with an AC load around equal to 12Ω will be presented basing on the diagram in Fig. 5.18. The chosen $2C_s$ value for this example is 22nF.

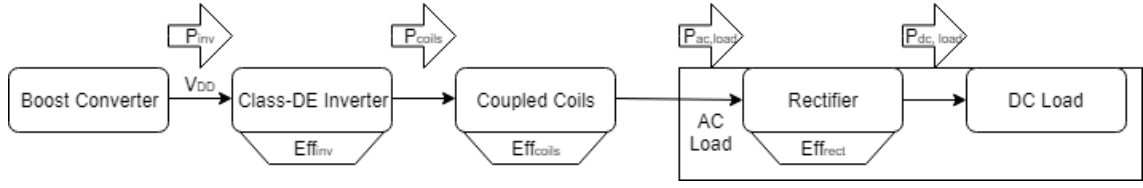


Figure 5.18: Use case diagram

Knowing that the DC load is required to receive 80W, it is possible to calculate the anterior power values using the efficiencies obtained in the results for these specific conditions. Looking at Fig. 5.15(a), (c) and (d) these are: $\eta_{inv} = 99\%$, $\eta_{coils} = 91\%$ and $\eta_{rect} = 95\%$, respectively, and result in the following powers calculated from $P_{dc,load} = 80W$: $P_{ac,load} = 84.2W$, $P_{coils} = 92.5W$, $P_{inv} = 93.5W$.

Having the AC load and $2C_s$ values been defined, it is possible to obtain the operating frequency and duty factor from Fig. 5.17, which results in a frequency of around 162kHz and a duty factor of around 42%. Now, using the obtained frequency and the predefined AC load, it is possible to obtain the equivalent resistance seen from the inverter, as in (4.33). Then, using the obtained D and equivalent resistance seen from the inverter, and having determined the desired output power from the inverter, $P_{coils} = 92.5W$, it is possible to define the V_{DD} value to be outputted from the boost converter using (4.20), which results in approximately $V_{DD} = 28.7V$. The previously presented use case is summarised in table 5.6, table 5.7 and table 5.8.

Table 5.6: Use case summary – Class-DE parameters

AC R_L	$2C_s$	Frequency	D
12 Ω	22nF	162kHz	42%

Table 5.7: Use case summary – Efficiencies

η_{rect}	η_{coils}	η_{inv}
95%	91%	99%

Table 5.8: Use case summary – Powers and V_{DD}

$P_{dc,load}$	$P_{ac,load}$	P_{coils}	P_{inv}	V_{DD}
80W	84.2W	92.5W	93.5W	28.7V

5.4 Summary

After the design and implementation of the system has been described followed by a presentation of the conducted tests and the created setup, the results were analysed with detail. From these results, which mainly regarded the delivered power and efficiency variation by changing different system parameters, it was very obvious to conclude that the very first objective of a higher system

efficiency using the new circuit of a class DE applied to wireless power transfer when compared to the traditionally used class D was achieved, even though the power capability slightly decreases when using the new circuit topology, which was not problematic since the needed amount of power is still reachable. Further, the obtained experimental results were compared to the predicted ones, showing a satisfying proximity.

Finally, using the obtained results, a use case example was created and explained for an implementation in a high efficiency operating point with the required power delivered to the DC load.

coils must always be the same, and the transmission medium must always be salted water. Only that way this architecture is valid.

Having this mapping be done once the controller is initialised, it is required that it knows the value of the load resistance to conveniently adjust its duty factor and frequency. However, it is assumed that there is no communication link between the transmitter and the receiver. That way, the controller needs to indirectly determine the load value, since it can only measure the electrical quantities existent in the primary side. That means that the controller is able to measure the equivalent impedance seen from the inverter, R_{eq} . With that in mind, the controller uses the previously presented expression (4.33) to accurately determine the load value, R_L . One should once again note that such strategy will only be valid if the T-model parameters remain equal to the known ones, since they are needed here again. Once the R_L value is obtained, it is possible to set the convenient duty factor and frequency using the mapping obtained in the initialisation.

Concerning the boost converter control, it may be achieved through a traditional proportional integral (PI) loop. The reference power value is compared to the one measured using the voltage and current on the power inverter output and an error is obtained. This error and its integral will then be multiplied with the corresponding proportional and integral gains, k_p and k_i , resulting in a control signal that may actuate in whatever is controlling the boost converter. It may be, for example, directly on the PWM generator that drives the boost converter MOSFET.

A more practical implementation in a simulation is described in the next section.

6.2 Control Simulation

In order to implement the previously described architecture, the Simulink scheme from Fig. 6.2 was created.

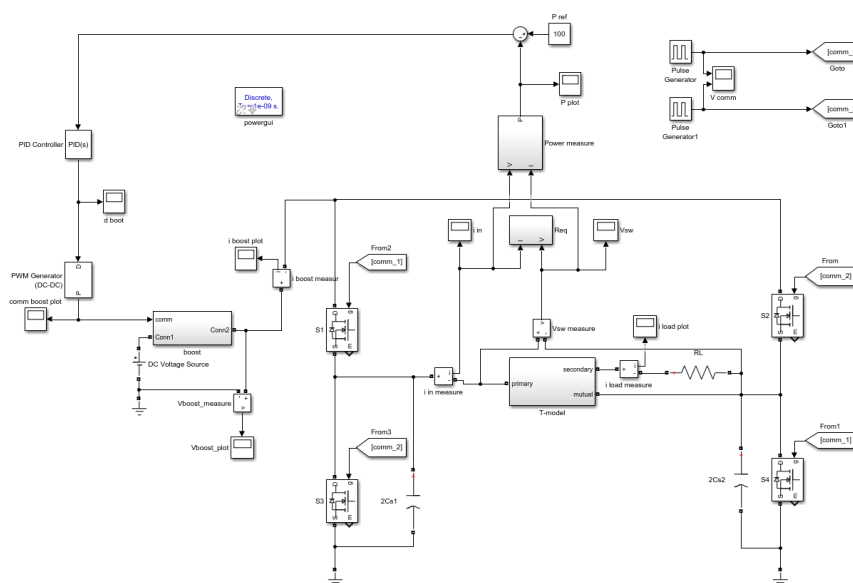


Figure 6.2: Simulink control scheme

As it can be seen, the system is comprised by a DC voltage source of 12V followed by a boost converter to step up the voltage. This converter was created in a sub block called "boost", which is shown in Fig. 6.3(a), and is connected to a power inverter with two shunt capacitances on each bottom switch to allow the realisation of the class DE as explained before in this document. The inverter then works as an AC source for a T-model circuit created inside the sub block called "T-model" as depicted in Fig. 6.3(b). The T-model parameters were set like previously shown in chapter 3, being variable with frequency.

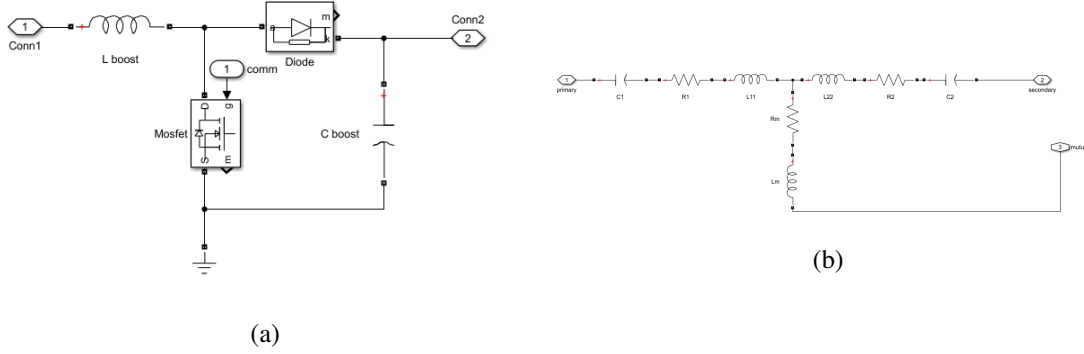


Figure 6.3: (a) - Simulink boost converter block, (b) - Simulink T-model block

Further, for the control realisation, two measurements are performed, namely the differential switching voltage and the inverter output current. These measurements have two different purposes, which are the determination of the resistance seen from the inverter and the delivered power.

First, regarding the resistance seen from the inverter determination, the sub block is implemented as seen in Fig. 6.4(a). As such, considering the higher current harmonics despicable, a Fourier transform for the fundamental frequency is performed for both the differential switching voltage and the inverter output current. Then, using the modules of both resulting phasors, the impedance module is determined. The real part of that impedance, corresponding to R_{eq} is then obtained using the phase difference between the two phasors, as shown in the diagram. The resulting R_{eq} is sent to the MATLAB workspace so that the load resistance value R_L is determined using (4.33) and the needed duty factor and frequency are obtained using the previously created mapping. Following, these duty factor and frequency values are sent back to Simulink and the inverter starts operating in class DE, as will be seen further in the results.

Concerning the delivered power determination, the sub block is implemented as depicted in Fig. 6.4(b). Once again, considering the higher current harmonics despicable, a Fourier transform for the fundamental frequency is performed for both the differential switching voltage and the inverter output current. Then, using the modules of both resulting phasors, the complex power phasor module is determined. The real part of that complex power is then obtained using the phase difference between the two phasors, as shown in the diagram. The resulting measured delivered power is then compared to a predefined power reference and the resulting error is processed by a PI controller. The PI controller was manually tuned using a heuristic approach for a null steady

state error and relatively fast transient response with low oscillation. The PI controller output is then used as a PWM signal generation reference, which will directly actuate on the boost converter MOSFET commutation. This control loop can be seen in Fig. 6.2.

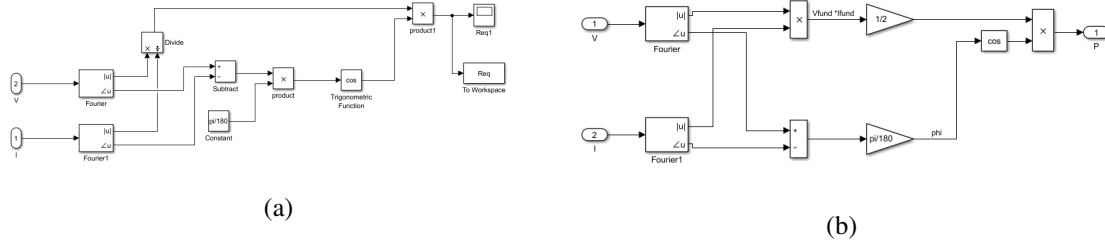


Figure 6.4: (a) - Simulink R_{eq} measurement block, (b) - Simulink power measurement block

6.3 Simulation Results

Following the previous described implementation, the corresponding results will now be presented. In this example, the two used shunt capacitances C'_s are of 62nF and the load R_L is set to 10 Ω . First, analysing the resulting differential switching voltage from Fig. 6.5(a), it can be seen that in any period, like in Fig. 6.5, the control system appears to be able to create class-DE conditions through the measurements described before. Similarly satisfying results were obtained for tests in a range of R_L values from 2 Ω to 25 Ω , showing a good performance of the designed control architecture in the class-DE maintenance part.

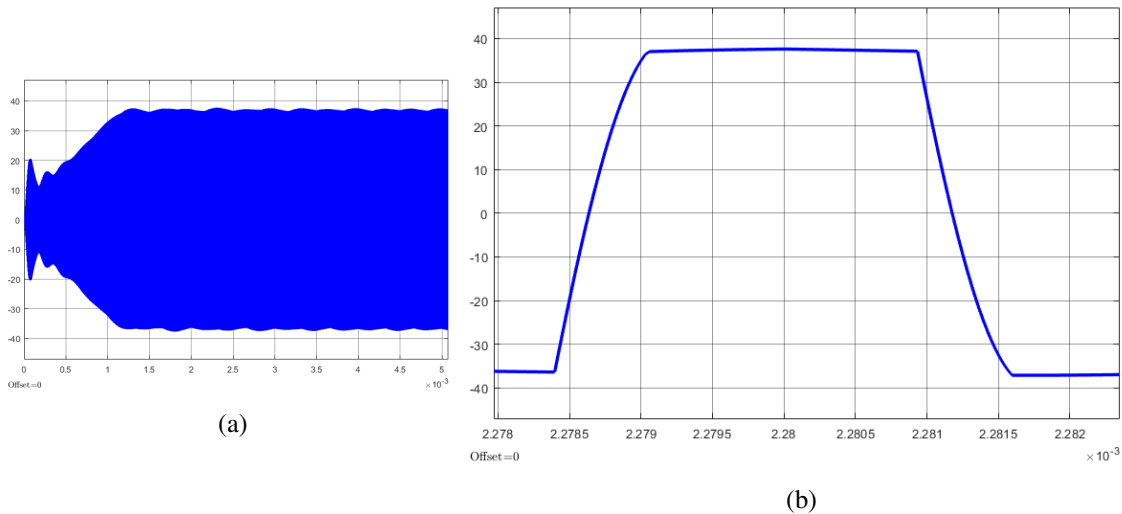


Figure 6.5: (a) - Simulation differential switching voltage, (b) - One period of the simulation differential switching voltage

Further, analysing the capacity of the system to follow a power reference, the first graph to be observed should be the one in Fig. 6.6(a). The predefined power reference was set to 100W, and it can be seen that the controller is able to follow this value with a good precision in steady

state, establishing in less than 1.5ms and with some minor oscillations. Observing now the boost converter output voltage, which is how the controller changes the delivered power value, it can be seen that it reaches a steady state at around 36V and has small oscillations.

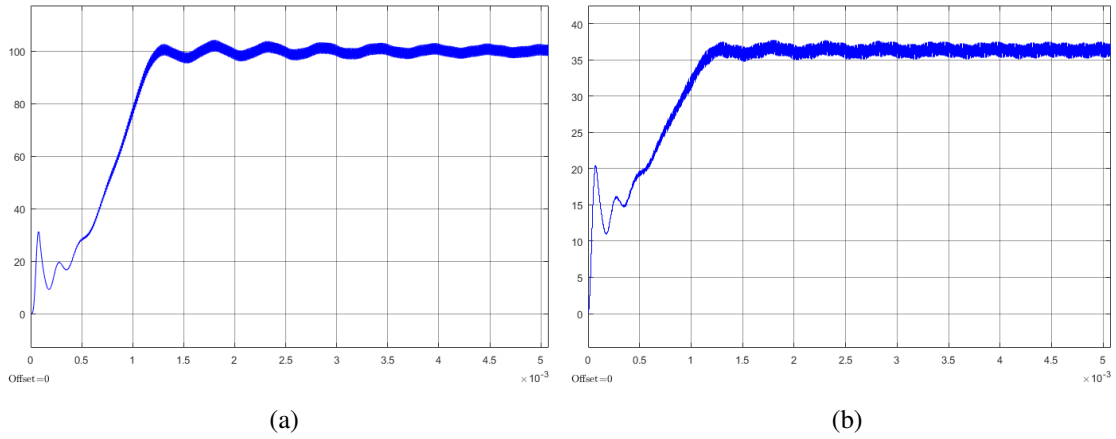


Figure 6.6: (a) - Simulation delivered power, (b) - Simulation boost converter voltage

Now, it would be interesting to observe the inverter output current behaviour. As such, it can be seen in Fig. 6.7(a) that when the 100W power is reached, the current maximum value is almost 6A. Further, Fig. 6.7(b) shows only two periods of the same current, confirming the previous assumption that the higher current harmonics are despicable.

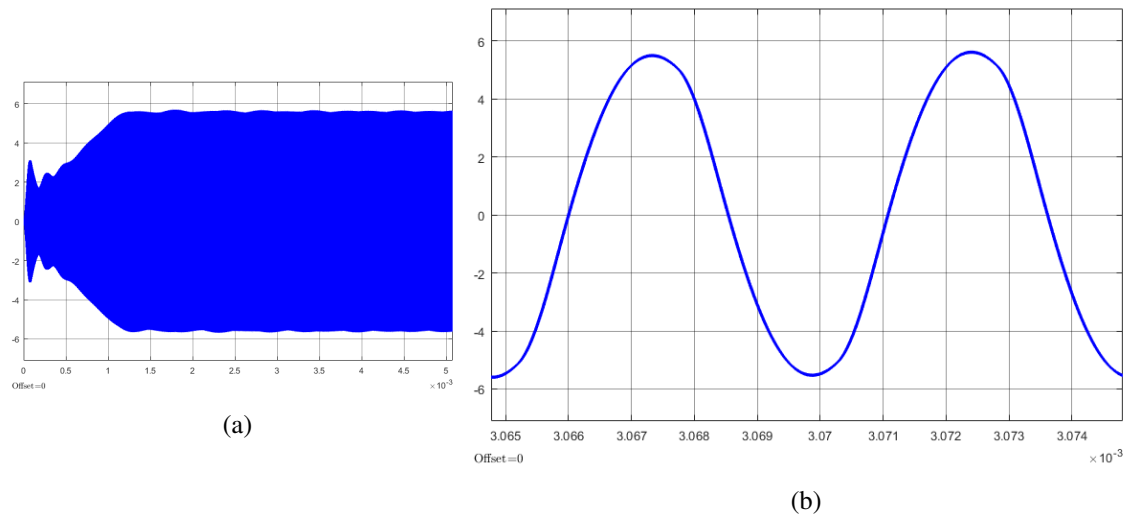


Figure 6.7: (a) - Simulation inverter output current, (b) - Two periods of the simulation inverter output current

The previously presented results prove that the designed control architecture has a promising potential for further development and implementation on the real tested system, allowing an accurate delivered power control while the proven high efficiency of the inverter is maintained by being able to operate in class DE at different load conditions.

6.4 Summary

In this chapter, a control approach to keep the system operating in class DE while controlling the delivered power was presented and tested through simulations of the system in Simulink. The results have shown a satisfying performance of the proposed architecture, making it possible to be implemented in the real system as a future work.

Chapter 7

Conclusions and Future Work

In this chapter, a review of the accomplishments of this thesis is conducted. Further, additional future work to improve this project is presented.

7.1 Satisfaction of Objectives

Now that the developed work and respective results were analysed, it is opportune to evaluate its quality. As such, regardless of the demonstrated high performance of the developed transmitter added to its relevant degree of innovation in the area of underwater wireless power transfer, this work quality must be evaluated by its conformity with the proposed requirements, here in the form of the defined objectives and goals.

As mentioned before, the first objective was the realisation of the control of a transmitter system capable of monitor all system parameters, such as current sensing, voltage sensing, switching waveform monitoring and overload system protection. This was the first part of the developed work and involved the greatest part of the microcontroller programming, to control the basic operation of the inverter. This part, as mentioned before, it was chosen to be implemented for an interactive operation, i.e. with required commands from an user, with the goal of facilitating the test of different parameters. That being said, this objective was satisfied with success with the exception of the switching waveform monitoring, due to the nonexistence of a voltage sensing module in this point of the board, and the overload system protection.

The following objective was to design and develop a class-DE full bridge power inverter to improve its efficiency and, consequently, the whole system efficiency. As presented before, this part of the thesis took the biggest parcel of the work, with existing theory being analysed and additional new theory being formulated for a wireless power transfer appliance, with further simulations, implementation and experimental testing. In this part, one could have used the traditional class-DE theory to realise the system. However, using as an advantage the system high inductive capability, it was shown that an operation slightly above resonance would be sufficient to realise the class-DE inverter, being only necessary to add two capacitors. With that in mind, a mapping of the class-DE operation was created, containing all system variables. Further, after confirming

the theoretical formulations through simulations with the coupled coils model, a test setup was built and the consequent tests proven the theory to be correct and shown a relevant increase in the transmitter performance in terms of efficiency, while still holding a satisfying power capability. That being said, it is possible to affirm that this objective was fully satisfied.

The next objective was to design of a suitable control approach of the system to keep the power transfer efficiency at a maximum. Even though a control architecture has been designed and simulated, this objective was not completely fulfilled since this architecture was not implemented and tested in the real system due to time constraints.

Finally, regarding the last objective of complete system integration and validation in a practical scenario, it was described before how it was approached. This was the final part of the work and consisted in the creation of an environment similar to the real application, which was the ocean. As such, a water tank containing the same saline concentration that can be found in the sea water was prepared, allowing the testing with a high degree of proximity to the real application. Further, relative to the complete integration, as shown before, the developed system block was completely integrated with the rest of the system blocks and the results were taken and evaluated not only for the developed block, but also for each of the other blocks and their junction. Additionally, like it was mentioned before, the developed power inverter took in account the interaction with the other blocks and was designed so that it had operation regions in the other blocks high performing operating regions. In conclusion, it can be said that the last objective was successfully satisfied.

7.2 Future Work

Having this project so many variables and being composed of so many different blocks, the future work may take several different directions, with many parts having room for relevant improvements.

As mentioned in the previous section, the design of a suitable control approach of the system to keep the power transfer efficiency at a maximum and was not implemented and tested on the real system due to time constraints, making it the first step in the future work. Additionally, a different and more adaptive control strategy could be adopted, eliminating the need of knowing the system parameters like it happens in the previously presented architecture. Besides the actuation over the boost converter to control the delivered power, which would remain the same, the controller would receive the acquired DC current and voltage in the boost converter output and the AC current and voltage in the the power inverter output. With these measures, the controller would be able to calculate the power inverter efficiency and, using a maximum tracking algorithm, it will actuate over the duty factor, D , and the frequency of the inverter to ensure it operates on a maximum efficiency region, corresponding to a practical class-DE. Practical because it may theoretically be a pure class-DE, as explained before, but still maintain the efficiency values of a pure class-DE.

As it was mentioned, there are several ways to improve the overall system. First, in order to have a more complete characterisation of the developed system, the conducted tests could be complemented. The presented results were all obtained with almost no distance between the coils

enclosure and with almost perfect alignment. As such, if more tests are made with different distances and different alignments, further conclusions would be taken regarding the behaviour of the system for different values of the coupling coefficient, which would result in a more complete system characterisation.

Additionally, as an improvement to the system itself, the coils could be redesigned to obtain the maximum possible quality factor, i.e., the maximum inductance to resistance ratio at a certain frequency. Such could be obtained through electromagnetic simulations using tools such as the Ansys Maxwell software, which allows to perform parametric coupled coils simulations to find the optimum parameters. Further, when redesigning the coils, different materials could be used, namely Litz wire to reduce the AC resistance, as described in the literature review and, finally, a ferrite core could be added to improve the coils inductance and consequent quality factor, improving the coupling coefficient as well.

One other system improvement could be made by fully characterise the boost converter that is placed before the power inverter. By doing so, its optimum efficiency regions could be determined and its operation could be limited to it, improving the whole system efficiency.

Finally, having observed that a significant amount of losses happen in the rectifier at the secondary side, especially for smaller loads, such block could be redesigned in a way that its efficiency is improved. As an example, a synchronous rectifier composed by MOSFETs instead of diodes would reduce the conduction losses, since the diodes always present a forward voltage drop, which may represent some significant losses.

Appendix A

Hardware

A.1 Test Material List

Table A.1: Testing material

Power Analyser	ZES ZIMMER, LMG671
Oscilloscope	Tektronix, TDS 2022B
Amperometric Clamp	ELDITEST, CP6990
Multimeter	Protek, 505
DC Voltage Source 1	Tenma, 72-2690
DC Voltage Source 2	Extech Instruments, 382276
Rheostat	TE Connectivity, BFA522R
Thermal Camera	FLIR E5
LCR Meter	Keysight, E4980AL

Appendix B

Software

B.1 Microcontroller Code

```
#pragma config FMIIEN = ON
#pragma config FETHIO = ON
#pragma config PGL1WAY = ON
#pragma config PMDL1WAY = ON
#pragma config IOL1WAY = ON
#pragma config FUSBIDIO = ON
#pragma config FPLLIDIV = DIV_1
#pragma config FPLL RNG = RANGE_8_16_MHZ
#pragma config FPLLCLK = PLL_FRC
#pragma config FPLLMULT = MUL_128
#pragma config FPLLODIV = DIV_4
#pragma config UPLLFSEL = FREQ_24MHZ
#pragma config FNOSC = SPL
#pragma config DMTINTV = WIN_127_128
#pragma config FSOSCEN = ON
#pragma config IESO = ON
#pragma config POSCMOD = OFF
#pragma config OSCIOFNC = OFF
#pragma config FCKSM = CSECME
#pragma config WDTPS = PS1048576
#pragma config WDTSPGM = STOP
#pragma config WINDIS = NORMAL
#pragma config FWDTEN = OFF
#pragma config FWDTWINSZ = WINSZ_25
#pragma config DMTCNT = DMT31
#pragma config FDMTEN = OFF
#pragma config DEBUG = OFF
#pragma config JTAGEN = OFF
#pragma config ICESEL = ICS_PGx1
#pragma config TRCEN = ON
#pragma config BOOTISA = MIPS32
#pragma config FECCCON = OFF_UNLOCKED
#pragma config FSLEEP = OFF
#pragma config DBGPER = PG_ALL
#pragma config SMCLR = MCLR_NORM
#pragma config SOSCGAIN = GAIN_2X
#pragma config SOSCOBOOST = ON
```

```

#pragma config POSCGAIN = GAIN_2X
#pragma config POSCBOOST = ON
#pragma config EJTAGBEN = NORMAL
#pragma config CP = OFF

#include<stdio.h>
#include<stdlib.h>
#include<xc.h>
#include<stdint.h>
#include"config.h"
#include<string.h>
#include<sys/attribs.h>
#define PBCLK1 128000000
char data_rx_uart[256], n_rx_uart;
char rx_buffer;
int freq_pwm = 0, duty_cycle;
float dead_time, set_duty_cycle;
float data;
float result0,result1,result2,result3,result4;
uint16_t adc_0[1024], adc_1[1024], adc_2[1024], adc_3[1024], adc_4[1024];
uint16_t stop_flag = 0;
int adc_index;
float bat_voltage, bat_current, boost_voltage, boost_current;
uint8_t sample0 = 0;
uint8_t sample1 = 0;
uint8_t sample2 = 0;
uint8_t sample3 = 0;
uint8_t sample4 = 0;
uint8_t trigger = 1;
uint8_t newF4 = 0;
uint8_t newD4 = 0;
uint8_t oldF4 = 0;
uint8_t oldD4 = 0;
uint16_t count = 0;
uint8_t start_t3 = 0;
uint8_t increase = 0;
uint32_t aux_resistance=15000;
uint32_t resistance=15000;

init_uart(){
    TRISFbits.TRISF5 = 0;
    TRISDbits.TRISD10 = 1;
    CFGCONbits.IOLOCK = 0;
    RPF5R = 0b0011;
    U5RXR = 0b0011;
    CFGCONbits.IOLOCK = 1;
    uint32_t baud=9600;
    U5MODE = 0;
    uint32_t baudrate = PBCLK1/(16*baud)-1;
    U5BRG = baudrate;
    U5STA = 0;
    U5STAbits.URXISEL0 = 0;
    U5STAbits.URXISEL1 = 0;
    IPC45bits.U5RXIP = 7;
    IFS5bits.U5RXIF = 0;
    IEC5bits.U5EIE = 0;
    IEC5bits.U5TXIE = 0;

```

```

    IEC5bits.U5RXIE = 1;
    U5MODEbits.LPBACK = 0;
    U5STAbits.UTXEN = 1;
    U5STAbits.URXEN = 1;
    U5MODEbits.ON = 1;
}

int serial_transmit(char* buffer){
    unsigned int size = strlen(buffer);
    while(size){
        while(U5STAbits.UTXBF);
        U5TXREG = *buffer;
        buffer++;
        size--;
    }
    while(!U5STAbits.TRMT);
    return 1;
}

init_adc(){
    TRISBbits.TRISB0 = 1;
    ANSELBbits.ANSB0 =1;
    TRISBbits.TRISB1 = 1;
    ANSELBbits.ANSB1 =1;
    TRISBbits.TRISB2 = 1;
    ANSELBbits.ANSB2 =1;
    TRISBbits.TRISB3 = 1;
    ANSELBbits.ANSB3 =1;
    TRISBbits.TRISB4 = 1;
    ANSELBbits.ANSB4 =1;
    ADC0CFG = DEVADC0;
    ADC1CFG = DEVADC1;
    ADC2CFG = DEVADC2;
    ADC3CFG = DEVADC3;
    ADC4CFG = DEVADC4;
    ADC7CFG = DEVADC7;
    ADCCON1 = 0;
    ADCCON2 = 0;
    ADCANCON = 0;
    ADCANCONbits.WKUPCLKCNT = 5;
    ADCCON3 = 0;
    ADCCON3bits.ADCSEL = 0;
    ADCCON3bits.CONCLKDIV = 1;
    ADCCON3bits.VREFSEL = 0;
    ADC0TIMEbits.ADCDIV = 0b0000001;
    ADC0TIMEbits.SAMC = 0;
    ADC0TIMEbits.SELRES = 3;
    ADC1TIMEbits.ADCDIV = 0b0000001;
    ADC1TIMEbits.SAMC = 0;
    ADC1TIMEbits.SELRES = 3;
    ADC2TIMEbits.ADCDIV = 0b0000001;
    ADC2TIMEbits.SAMC = 0;
    ADC2TIMEbits.SELRES = 3;
    ADC3TIMEbits.ADCDIV = 0b0000001;
    ADC3TIMEbits.SAMC = 0;
    ADC3TIMEbits.SELRES = 3;
    ADC4TIMEbits.ADCDIV = 0b0000001;

```

```

ADC4TIMEbits.SAMC = 0;
ADC4TIMEbits.SELRES = 3;
ADCTRGMODEbits.SH0ALT = 0;
ADCIMCON1bits.SIGN0 = 0;
ADCIMCON1bits.DIFF0 = 0;
ADCTRGMODEbits.SH1ALT = 0;
ADCIMCON1bits.SIGN1 = 0;
ADCIMCON1bits.DIFF1 = 0;
ADCTRGMODEbits.SH2ALT = 0;
ADCIMCON1bits.SIGN2 = 0;
ADCIMCON1bits.DIFF2 = 0;
ADCTRGMODEbits.SH3ALT = 0;
ADCIMCON1bits.SIGN3 = 0;
ADCIMCON1bits.DIFF3 = 0;
ADCTRGMODEbits.SH4ALT = 0;
ADCIMCON1bits.SIGN4 = 0;
ADCIMCON1bits.DIFF4 = 0;
ADCGIRQEN1bits.AGIEN0 = 1;
ADCGIRQEN1bits.AGIEN1 = 1;
ADCGIRQEN1bits.AGIEN2 = 1;
ADCGIRQEN1bits.AGIEN3 = 1;
ADCGIRQEN1bits.AGIEN4 = 1;
ADCGIRQEN2 = 0;
IPC14bits.ADCD0IP = 0b100;
IFS1bits.ADCD0IF = 0;
IEC1bits.ADCD0IE = 1;
IPC15bits.ADCD1IP = 0b110;
IFS1bits.ADCD1IF = 0;
IEC1bits.ADCD1IE = 1;
IPC15bits.ADCD2IP = 0b101;
IFS1bits.ADCD2IF = 0;
IEC1bits.ADCD2IE = 1;
IPC15bits.ADCD3IP = 0b111;
IFS1bits.ADCD3IF = 0;
IEC1bits.ADCD3IE = 1;
IPC15bits.ADCD4IP = 0b011;
IFS1bits.ADCD4IF = 0;
IEC1bits.ADCD4IE = 1;
asm volatile("ei");
INTCONbits.MVEC = 1;
ADCCSS1 = 0;
ADCCSS2 = 0;
ADCCMPCON1 = 0;
ADCCMPCON2 = 0;
ADCCMPCON3 = 0;
ADCCMPCON4 = 0;
ADCCMPCON5 = 0;
ADCCMPCON6 = 0;
ADCFLTR1 = 0;
ADCFLTR2 = 0;
ADCFLTR3 = 0;
ADCFLTR4 = 0;
ADCFLTR5 = 0;
ADCFLTR6 = 0;
ADCTRGSNSbits.LVL2 = 0;
ADCTRG1bits.TRGSRC0 = 1;
ADCTRG1bits.TRGSRC1 = 1;

```

```

    ADCTRG1bits.TRGSRC2 = 1;
    ADCTRG1bits.TRGSRC3 = 1;
    ADCTRG2bits.TRGSRC4 = 1;
    ADCEIEN1 = 0;
    ADCEIEN2 = 0;
    ADCCON1bits.ON = 1;
    while(!ADCCON2bits.BGVRRDY);
    while(ADCCON2bits.REFFLT);
    ADCANCONbits.ANEN0 = 1;
    ADCANCONbits.ANEN1 = 1;
    ADCANCONbits.ANEN2 = 1;
    ADCANCONbits.ANEN3 = 1;
    ADCANCONbits.ANEN4 = 1;
    while(!ADCANCONbits.WKRDY0);
    while(!ADCANCONbits.WKRDY1);
    while(!ADCANCONbits.WKRDY2);
    while(!ADCANCONbits.WKRDY3);
    while(!ADCANCONbits.WKRDY4);
    ADCCON3bits.DIGEN0 = 1;
    ADCCON3bits.DIGEN1 = 1;
    ADCCON3bits.DIGEN2 = 1;
    ADCCON3bits.DIGEN3 = 1;
}

void init_timer1(){
    T1CON = 0;
    TMR1 = 0;
    T1CONbits.TCKPS = 0b00;
    PR1 = 0xffff;
    IPC1bits.T1IP = 0b011;
    IFS0bits.T1IF = 0;
    IEC0bits.T1IE = 1;
    asm volatile("ei");
    T1CONbits.ON = 1;
    INTCONbits.MVEC = 1;
}

void init_timer3(){
    T3CON = 0;
    TMR3 = 0;
    T3CONbits.TCKPS = 0b01;
    PR3 = 0x04E2;
    IPC3bits.T3IP = 0b011;
    IFS0bits.T3IF = 0;
    IEC0bits.T3IE = 1;
    asm volatile("ei");
    T3CONbits.ON = 1;
    INTCONbits.MVEC = 1;
}

init_pwm(int freq, float duty_cycle, float dead_time){
    PORTGbits.RG9 = 0;
    PORTGbits.RG8 = 0;
    PORTDbits.RD1 = 0;
    PORTDbits.RD2 = 0;
    OC1CON = 0;
    OC9CON = 0;

```

```

OC3CON = 0;
OC6CON = 0;
T2CON = 0;
TMR2 = 0;
T2CONbits.TCKPS = 0b000;
PR2 = PBCLK1/freq+1;
IPC2bits.T2IP = 0b001;
IFS0bits.T2IF = 0;
IEC0bits.T2IE = 1;
OC9CON = 0x0000;
OC9R = 0;
OC9RS = PR2*duty_cycle+1;
OC9CONbits.OCM = 5;
OC6CON = 0x0000;
OC6R = OC9RS + PR2 *dead_time;
OC6RS = OC6R + PR2 *duty_cycle+1;
OC6CONbits.OCM = 5;
OC1CON = 0x0000;
OC1R = 0;
OC1RS = PR2*duty_cycle+1;
OC1CONbits.OCM = 5;
OC3CON = 0x0000;
OC3R = OC1RS + PR2 *dead_time;
OC3RS = OC3R + PR2 *duty_cycle+1;
OC3CONbits.OCM = 5;
asm volatile("ei");
INTCONbits.MVEC = 1;
T2CONbits.ON = 1;
OC1CON |= 0x8000;
OC9CON |= 0x8000;
OC3CON |= 0x8000;
OC6CON |= 0x8000;
}

init_spi(){
    CFGCONbits.IOLOCK = 0;
    RPF1R = 0b0111;
    SDI3R = 0b0101;
    CVRCONbits.CVROE = 0;
    CVRCONbits.ON = 0 ;
    CFGCONbits.IOLOCK = 1;

    int rData;
    IEC4&=0x10111111;
    SPI3CON = 0;
    SPI3CON2 = 0;
    rData=SPI3BUF;
    IPC39bits.SPI3TXIP = 0b111;
    IFS4bits.SPI3TXIF = 0;
    IEC4bits.SPI3TXIE = 0;
    SPI3BRG=0xff;
    SPI3STATCLR=0x40;
    SPI3CON2bits.AUDEN=0;
    SPI3CONbits.SMP=1;
    SPI3CONbits.MSTEN=1;
    SPI3CONbits.MSSEN=0;
    SPI3CONbits.CKP=0;

```

```

    SPI3CONbits.MODE32=1;
    SPI3CONbits.MODE16=0;
    SPI3CONbits.ON=1;
}

void init(){
    TRISGbits.TRISG9 = 0; //setting as output
    PORTGbits.RG9 = 0;
    TRISGbits.TRISG8 = 0;
    PORTGbits.RG8 = 0;
    TRISDbits.TRISD1 = 0;
    PORTDbits.RD1 = 0;
    TRISDbits.TRISD2 = 0;
    PORTDbits.RD2 = 0;
    CFGCONbits.IOLOCK = 0;
    RPG9R = 0b1100;
    RPG8R = 0b1011;
    RPD1R = 0b1101;
    RPD2R = 0b1100;
    CFGCONbits.IOLOCK = 1;
    TRISFbits.TRISF0 = 0;
    TRISBbits.TRISB15 = 0;
    ANSELBbits.ANSB15 = 0;
    TRISBbits.TRISB8 = 1;
    ANSELBbits.ANSB8 = 0;
    TRISBbits.TRISB9 = 1;
    ANSELBbits.ANSB9 = 0;
    TRISFbits.TRISF1 = 0;
    TRISDbits.TRISD11 = 1;
}

float get_bat_voltage(){
    char buf[5];
    float bat_voltage_sum=0;
    float result_dec = result0*(3.274/4096);
    data=result_dec;
    float bat_voltage = data*(66500+232000)/66500;
    bat_voltage_sum = bat_voltage_sum+bat_voltage;
    float avg_bat_voltage = bat_voltage_sum;
    bat_voltage_sum = 0;
    sample0 = 0;
    return avg_bat_voltage;
}

float get_bat_current(){
    char buf[5];
    float bat_current_sum=0;
    float result_dec = result1*(3.274/4096);
    data=result_dec;
    float bat_current = 0 + (data*59/39 - 2.462) / 0.06;
    bat_current_sum = bat_current_sum+bat_current;
    float avg_bat_current = bat_current_sum;
    bat_current_sum = 0;
    sample1 = 0;
    return avg_bat_current;
}

```

```

float get_boost_voltage(){
    char buf[5];
    float boost_voltage_sum=0;
    float result_dec = result2*(3.274/4096);
    data=result_dec;
    float boost_voltage = data*16.9862;
    boost_voltage_sum = boost_voltage_sum+boost_voltage;
    float avg_boost_voltage = boost_voltage_sum;
    boost_voltage_sum = 0;
    sample2 = 0;
    return avg_boost_voltage;
}

float get_boost_current(){
    char buf[5];
    uint8_t n=10;
    int i = 0;
    float boost_current_sum=0;
    float result_dec = result3*(3.274/4096);
    data=result_dec;
    float boost_current = 0 + (data*59/39 - 2.462) / 0.06;
    boost_current_sum = boost_current_sum+boost_current;
    i=0;
    float avg_boost_current = boost_current_sum;
    boost_current_sum = 0;
    sample3 = 0;
    return avg_boost_current;
}

float get_inv_current(){
    char buf[5];
    uint8_t n=10;
    int i = 0;
    float inv_current_sum=0;
    float result_dec = result4*(3.278/4096);
    data=result_dec;
    float inv_current = 0 + (data*59/39 - 2.455) / 0.06;
    inv_current_sum = inv_current_sum+inv_current;
    i=0;
    float avg_inv_current = inv_current_sum;
    inv_current_sum = 0;
    snprintf(buf, sizeof(buf), "%f", avg_inv_current);
    serial_transmit("inv current: ");
    serial_transmit(&buf);
    serial_transmit("\r");
    sample4 = 0;
    return avg_inv_current;
}

void set_pot(uint32_t resistancel){
    uint32_t max = 0x0003ff;
    uint32_t min = 0x000000;
    uint32_t address1 = 0xB00000;
    uint32_t address2 = 0xB10000;
    if (resistancel>24000){
        resistancel = 24000;
    }
}

```



```

if (resistance1<100){
    resistance1 = 100;
}
trigger = 0;
if (aux_resistance==resistance1){
    uint32_t pot1 = 24840-aux_resistance;
    uint32_t value1 = (pot1-98.9)/((24840-98.9)/1024);
    if (value1>max){
        value1=max;
    }
    if (value1<min){
        value1=min;
    }
    uint32_t code1 = address1+value1;
    uint32_t code2 = address2+value1;
    int i = 0;
    while(i<10){
        PORTBbits.RB15 = 0;
        SPI3BUF = code1;
        while (!SPI3STATbits.SPITBE);
        PORTBbits.RB15 = 1;
        PORTBbits.RB15 = 0;
        SPI3BUF = 0x200000;
        while (!SPI3STATbits.SPITBE);
        PORTBbits.RB15 = 1;
        i++;
    }
    i=0;
    while(i<10){
        PORTBbits.RB15 = 0;
        SPI3BUF = code2;
        while (!SPI3STATbits.SPITBE);
        PORTBbits.RB15 = 1;
        PORTBbits.RB15 = 0;
        SPI3BUF = 0x210000;
        while (!SPI3STATbits.SPITBE);
        PORTBbits.RB15 = 1;
        i++;
    }
    i = 0;
}
else if (aux_resistance<resistance1){
    while(aux_resistance<resistance1){
        start_t3 = 1;
        while(increase==0);
        increase = 0;
        aux_resistance=aux_resistance+100;
        uint32_t pot1 = 24840-aux_resistance;
        uint32_t value1 = (pot1-98.9)/((24840-98.9)/1024);
        if (value1>max){
            value1=max;
        }
        if (value1<min){
            value1=min;
        }
        uint32_t code1 = address1+value1;
        uint32_t code2 = address2+value1;
    }
}

```

```

int i = 0;
while(i<2){
    PORTBbits.RB15 = 0;
    SPI3BUF = code1;
    while (!SPI3STATbits.SPITBE);
    PORTBbits.RB15 = 1;
    PORTBbits.RB15 = 0;
    SPI3BUF = 0x200000;
    while (!SPI3STATbits.SPITBE);
    PORTBbits.RB15 = 1;
    i++;
}
i=0;
while(i<2){
    PORTBbits.RB15 = 0;
    SPI3BUF = code2;
    while (!SPI3STATbits.SPITBE);
    PORTBbits.RB15 = 1;
    PORTBbits.RB15 = 0;
    SPI3BUF = 0x210000;
    while (!SPI3STATbits.SPITBE);
    PORTBbits.RB15 = 1;
    i++;
}
i = 0;
}
}
else if (aux_resistance>resistancel){
    while(aux_resistance>resistancel){
        start_t3 = 1;
        while(increase==0);
        increase = 0;
        aux_resistance=aux_resistance-100;
        uint32_t pot1 = 24840-aux_resistance;
        uint32_t value1 = (pot1-98.9)/((24840-98.9)/1024);
        if (value1>max){
            value1=max;
        }
        if (value1<min){
            value1=min;
        }
        uint32_t code1 = address1+value1;
        uint32_t code2 = address2+value1;
        int i = 0;
        while(i<2){
            PORTBbits.RB15 = 0;
            SPI3BUF = code1;
            while (!SPI3STATbits.SPITBE);
            PORTBbits.RB15 = 1;
            PORTBbits.RB15 = 0;
            SPI3BUF = 0x200000;
            while (!SPI3STATbits.SPITBE);
            PORTBbits.RB15 = 1;
            i++;
        }
        i=0;
        while(i<2){

```

```

        PORTBbits.RB15 = 0;
        SPI3BUF = code2;
        while (!SPI3STATbits.SPITBE);
        PORTBbits.RB15 = 1;
        PORTBbits.RB15 = 0;
        SPI3BUF = 0x210000;
        while (!SPI3STATbits.SPITBE);
        PORTBbits.RB15 = 1;
        i++;
    }
    i = 0;
}
}
trigger = 1;
}

int main(){
    init();
    init_uart();
    serial_transmit(">>>> WPT - PIC CONTROLLER <<<< - help for list of
commands\n\r");
    while(PORTDbits.RD11!=0);
    init_spi();
    init_adc();
    init_timer1();
    init_timer3();
    set_duty_cycle = 0.37;
    dead_time = 0.5 - set_duty_cycle;
    freq_pwm = 180000;
    init_pwm(freq_pwm, set_duty_cycle, dead_time);
    uint32_t resistance = 15000;
    set_pot(resistance);
    while (1){
        terminal_console();
        if (sample0 == 1){
            bat_voltage = get_bat_voltage();
        }
        if (sample1 == 1){
            bat_current = get_bat_current();
        }
        if (sample2 == 1){
            boost_voltage = get_boost_voltage();
        }
        if (sample3 == 1){
            boost_current = get_boost_current();
        }
    }
}

void __ISR(_ADC_DATA0_VECTOR, ip14auto) ADC0DataIntHandler(void){
    result0 = ADCDATA0;
    sample0 = 1;
    IFS1bits.ADCD0IF = 0;
}

void __ISR(_ADC_DATA1_VECTOR, ip16auto) ADC1DataIntHandler(void){
    result1 = ADCDATA1;

```

```

    sample1 = 1;
    IFS1bits.ADCD1IF = 0;
}

void __ISR(_ADC_DATA2_VECTOR, ipl5auto) ADC2DataIntHandler(void){
    result2 = ADCDATA2;
    sample2 = 1;
    IFS1bits.ADCD2IF = 0;
}

void __ISR(_ADC_DATA3_VECTOR, ipl7auto) ADC3DataIntHandler(void){
    result3 = ADCDATA3;
    sample3 = 1;
    IFS1bits.ADCD3IF = 0;
}

void __ISR(_ADC_DATA4_VECTOR, ipl3auto) ADC4DataIntHandler(void){
    result4 = ADCDATA4;
    sample4 = 1;
    IFS1bits.ADCD4IF = 0;
}

void __ISR(_TIMER_1_VECTOR, ipl3auto) Timer1IntHandler(void){
    IFS0bits.T1IF = 0;
    if(trigger==1){
        ADCCON3bits.GSWTRG = 1;
    }
}

void __ISR(_TIMER_2_VECTOR, ipl1auto) T2_IntHandler (void){
    IFS0bits.T2IF = 0; // Clearing Timer2 interrupt flag
}

void __ISR(_TIMER_3_VECTOR, ipl3auto) Timer3IntHandler(void){
    IFS0bits.T3IF = 0;
    if (start_t3 == 1){
        count++;
        if (count > 1){
            start_t3=0;
            count=0;
            increase=1;
        }
    }
}

void __ISR(_TIMER_4_VECTOR, ipl1auto) T4_IntHandler (void){
    IFS0bits.T4IF = 0;
}

void __ISR(_TIMER_5_VECTOR, ipl1auto) T5_IntHandler (void){
    IFS0bits.T5IF = 0;
}

void __ISR(_TIMER_6_VECTOR, ipl1auto) T6_IntHandler (void){
    IFS0bits.T6IF = 0;
}

```



```

        serial_transmit(data_rx_uart);
        serial_transmit("\r\n");
        snprintf(buf, sizeof(buf), "%f", bat_voltage);
        serial_transmit("battery voltage: ");
        serial_transmit(&buf);
        serial_transmit(" V");
        serial_transmit("\r\n");
        snprintf(buf, sizeof(buf), "%f", bat_current);
        serial_transmit("battery current: ");
        serial_transmit(&buf);
        serial_transmit(" A");
        serial_transmit("\r\n");
        snprintf(buf, sizeof(buf), "%f", bat_voltage*bat_current);
        serial_transmit("battery power: ");
        serial_transmit(&buf);
        serial_transmit(" W");
        serial_transmit("\r\n");
        data_rx_uart[0] = ' ';
    }
else if (data_rx_uart[4] == 'b' && data_rx_uart[5] == 'o'
&& data_rx_uart[6] == 'o' && data_rx_uart[7] == 's'
&& data_rx_uart[8] == 't') {
    serial_transmit(data_rx_uart);
    serial_transmit("\r\n");
    snprintf(buf, sizeof(buf), "%f", boost_voltage);
    serial_transmit("boost voltage: ");
    serial_transmit(&buf);
    serial_transmit(" V");
    serial_transmit("\r\n");
    snprintf(buf, sizeof(buf), "%f", boost_current);
    serial_transmit("boost current: ");
    serial_transmit(&buf);
    serial_transmit(" A");
    serial_transmit("\r\n");
    snprintf(buf, sizeof(buf), "%f"
,boost_voltage*boost_current);
    serial_transmit("boost power: ");
    serial_transmit(&buf);
    serial_transmit(" W");
    serial_transmit("\r\n");
    data_rx_uart[0] = ' ';
}
}

break;
case 'p':
if (data_rx_uart[1] == 'w' && data_rx_uart[2] == 'm' && data_rx_uart[3] == ' ') {
    serial_transmit(data_rx_uart);
    serial_transmit("\r\n");
freq_pwm = strtoul((char*) &data_rx_uart[4], &pEnd, 10);
if (freq_pwm == 0){
    PORTGbits.RG9 = 0;
    PORTGbits.RG8 = 0;
    PORTDbits.RD1 = 0;
    PORTDbits.RD2 = 0;
    OC1CON = 0;
    OC9CON = 0;
    OC3CON = 0;

```

```

        OC6CON = 0;

T3CONbits.TON = 0;
data_rx_uart[0] = ' ';
}else if ((freq_pwm < 300000)&&(freq_pwm > 100000)){
duty_cycle = strtoul(pEnd, &pEnd, 10);
        if ((duty_cycle <= 499) && (duty_cycle >= 100)){
            set_duty_cycle = (float)duty_cycle/1000;
            dead_time = 0.5 - set_duty_cycle;
            init_pwm(freq_pwm, set_duty_cycle, dead_time);
        }
        else
            serial_transmit("incorrect pwm \r\n");
        data_rx_uart[0] = ' ';
} else
        break;

case 'h' :
if( data_rx_uart[1]=='e' && data_rx_uart[2]=='l' && data_rx_uart[3]=='p'){
        serial_transmit(data_rx_uart);
        serial_transmit("\r\n");

serial_transmit("-----\r\n");
serial_transmit("arrow left and right don't work, backspace works\r\n");
serial_transmit("last command - arrow UP\r\n");
serial_transmit("ADC measurements      - adc measurement \r\n");
serial_transmit("                ex: adc boost ;OR; adc bat \r\n");
        serial_transmit("change pwm      -
pwm freq_in_Hz(100000-500000) duty_cycle_in_1000(1-499) \r\n");
        serial_transmit("                ex: pwm 180000 400 \r\n");
        serial_transmit("emergency stop - pwm 0 \r\n");
serial_transmit("-----\r\n");
        data_rx_uart[0] = ' ';
}else
        break;

    }
    n_rx_uart = 0;
}
}
}

```

B.2 MATLAB Codes

B.2.1 Wireless Power Transfer System Theory

With the following MATLAB code it is possible to obtain the results presented in chapter 3.

```

set(0,'defaulttextInterpreter','latex');
set(groot, 'defaultAxesTickLabelInterpreter','latex');
set(groot, 'defaultLegendInterpreter','latex');

%% Fixed C and RL
clear all; close all; clc;

dataCsv =
load('OCEAN_freq_kreal_kimag2_ltx_rtx_lrx_rrx_l11_r11_l22_r22_lm_rm_salt.csv');

C_1=68e-9;

```

```

C_2=68e-9;
R_L=1:0.1:25;

L_1=dataCsv(:,4);
L_2=dataCsv(:,6);
R_1=dataCsv(:,5);
R_2=dataCsv(:,7);
M=dataCsv(:,12);
R_m=dataCsv(:,13);
k_real=dataCsv(:,2);
k_imag=dataCsv(:,3);
L_11=L_1-M;
L_22=L_2-M;
freq = dataCsv(:,1);
w=2*pi*freq;

figure(1)
plot(freq/1000,k_real)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Coupling Coefficient'})

figure(2)
plot(freq/1000,R_m*1000)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Mutual Resistance ( $m\Omega$ )'})

figure(3)
plot(freq/1000,M*1e6)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Mutual Inductance ( $\mu H$ )'})

figure(20)
plot(freq/1000,R_1*1000)
hold on
plot(freq/1000,R_2*1000)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Coils Resistance ( $m\Omega$ )'})
legend({'$R_1$', '$R_2$', 'Location', 'southeast'})
legend('boxoff')

figure(21)
plot(freq/1000,L_1*1e6)
hold on
plot(freq/1000,L_2*1e6)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Coils Inductance ( $\mu H$ )'})
legend({'$L_1$', '$L_2$', 'Location', 'northeast'})
legend('boxoff')

figure(22)
plot(freq/1000,L_11*1e6)

```



```

hold on
plot(freq/1000,L_22*1e6)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Leakage Inductance ( $\mu$  H)})')
legend({'$L_{11}$','$L_{22}$','Location','northeast'})
legend('boxoff')

for n = 1:numel(w)
    Q_tx(n) = w(n)*L_1(n)/R_1(n);
    Q_rx(n) = w(n)*L_2(n)/R_2(n);
    Eff_opt(n) = abs(k_real(n)+j*k_imag(n))^2 * Q_tx(n) * Q_rx(n)
    / ( 1 + sqrt(1+k_real(n)^2*Q_tx(n)*Q_rx(n)) )^2;
    for m=1:numel(R_L)
        Z_L(n,m) = j*w(n)*L_22(n)+1/(j*w(n)*C_2)+R_2(n)+R_L(m);
        Z_2(n,m) = 1/(1/Z_L(n,m)+1/(j*w(n)*M(n)+R_m(n)));
        Z_eq(n,m) = 1/(j*w(n)*C_1)+j*w(n)*L_11(n)+R_1(n)+Z_2(n,m);
        V_i = 10;
        P_i(n,m) = (0.5)*real(V_i^2/Z_eq(n,m));
        P_o(n,m) = real(P_i(n,m)) * real(Z_2(n,m))/(R_1(n)+real(Z_2(n,m)))
        * R_L(m)/(R_2(n)+R_L(m));
        Eff(n,m) = P_o(n,m)/P_i(n,m);
    end
end

figure(4)
plot(freq/1000,Q_tx)
hold on
plot(freq/1000,Q_rx)
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Coils Quality Factor'})
legend({'Primary Coil','Secondary Coil','Location','southeast'})
legend('boxoff')

figure(5)
plot(freq/1000,100*Eff_opt)
grid on
grid minor
xlabel({'Frequency (kHz)'})
ylabel({'Optimum Efficiency (%)'})

figure(6)
plot(freq/1000,P_o(:,41)/max(P_o(:,41)))
hold on
plot(freq/1000,P_o(:,61)/max(P_o(:,41)))
hold on
plot(freq/1000,P_o(:,81)/max(P_o(:,41))) %9ohm
hold on
plot(freq/1000,P_o(:,101)/max(P_o(:,41)))
hold on
plot(freq/1000,P_o(:,121)/max(P_o(:,41)))
grid on
xlabel({'Frequency (kHz)'})
ylabel({'Normalised Output Power'})
legend({'Output Power with $R_L=5 \backslash \Omega$','$Output Power with $R_L=7 \backslash \Omega$'
,'Output Power with $R_L=9 \backslash \Omega$','$Output Power with $R_L=11 \backslash \Omega$'

```

```
, 'Output Power with $R_L=13 \Omega$' }
, 'Location', 'northeast')
legend('boxoff')

figure(7)
plot(freq/1000, 100*Eff_opt)
hold on
plot(freq/1000, 100*Eff(:, 41))
hold on
plot(freq/1000, 100*Eff(:, 61))
hold on
plot(freq/1000, 100*Eff(:, 81)) %9ohm
hold on
plot(freq/1000, 100*Eff(:, 101))
hold on
plot(freq/1000, 100*Eff(:, 121))
grid on
grid minor
xlabel({'Frequency (kHz)'})
ylabel({'Efficiency (%)'})
legend({'Output Power with $R_L=5 \Omega$', 'Output Power with $R_L=7 \Omega$',
'Output Power with $R_L=9 \Omega$', 'Output Power with $R_L=11 \Omega$',
'Output Power with $R_L=13 \Omega$' }
, 'Location', 'northeast')
legend('boxoff')

figure(8)
plot(R_L, 100*Eff(81, :))
hold on
plot(R_L, 100*Eff(101, :))
hold on
plot(R_L, 100*Eff(121, :)) %140khz
hold on
plot(R_L, 100*Eff(141, :))
hold on
plot(R_L, 100*Eff(161, :))
grid on
grid minor
xlabel({'Load ($\Omega$)'})
ylabel({'Efficiency (%)'})
legend({'Output Power with $R_L=5 \Omega$', 'Output Power with $R_L=7 \Omega$',
'Output Power with $R_L=9 \Omega$', 'Output Power with $R_L=11 \Omega$',
'Output Power with $R_L=13 \Omega$' }
, 'Location', 'northeast')
legend('boxoff')

figure(9)
plot(R_L, P_o(81, :)/max(P_o(121, :)))
hold on
plot(R_L, P_o(101, :)/max(P_o(121, :)))
hold on
plot(R_L, P_o(121, :)/max(P_o(121, :))) %140khz
hold on
plot(R_L, P_o(141, :)/max(P_o(121, :)))
hold on
plot(R_L, P_o(161, :)/max(P_o(121, :)))
grid on
```

```

xlabel({'Load ($\Omega$)'})
ylabel({'Normalised Output Power'})
legend({'Output Power with $R_L=5 \ \Omega$', 'Output Power with $R_L=7 \ \Omega$',
'Output Power with $R_L=9 \ \Omega$', 'Output Power with $R_L=11 \ \Omega$',
'Output Power with $R_L=13 \ \Omega$'}
,'Location','northeast')
legend('boxoff')

%% Optimum C and RL
clear all; clc;

dataCsv =
load('OCEAN_freq_kreal_kimag2_ltx_rtx_lrx_rrx_l11_r11_l22_r22_lm_rm_salt.csv');

C_opt=1e-9:1e-9:900e-8;
V_i=10;
C_1=22e-9;
C_2=22e-9;
R_L=1:0.1:25;
L_1=dataCsv(:,4);
L_2=dataCsv(:,6);
R_1=dataCsv(:,5);
R_2=dataCsv(:,7);
M=dataCsv(:,12);
R_m=dataCsv(:,13);
L_11=L_1-M;
L_22=L_2-M;
k_real=dataCsv(:,2);
k_imag=dataCsv(:,3);
freq = dataCsv(:,1);
w=2*pi*freq;

for f = 1:numel(w)
    Q_tx(f) = w(f)*L_1(f)/R_1(f);
    Q_rx(f) = w(f)*L_2(f)/R_2(f);
    Eff_opt(f) = abs(k_real(f)+j*k_imag(f))^2 * Q_tx(f) * Q_rx(f)
    / ( 1 + sqrt(1+k_real(f)^2*Q_tx(f)*Q_rx(f)) )^2;

    for n = 1:numel(R_L) % optimal load with fixed capacitance
        Z_L(f,n) = j*w(f)*L_22(f)+1/(j*w(f)*C_2)+R_2(f)+R_L(n);
        Z_2(f,n) = 1/(1/Z_L(f,n)+1/(j*w(f)*M(f)+R_m(f)));
        Z_eq(f,n) = 1/(j*w(f)*C_1+j*w(f)*L_11(f)+R_1(f)+Z_2(f,n));
        P_i(f,n) = (0.5)*real(V_i^2/Z_eq(f,n));
        P_o(f,n) = real(P_i(f,n)) * real(Z_2(f,n))/(R_1(f)+real(Z_2(f,n)))
        * R_L(n)/(R_2(f)+R_L(n));
        Eff(f,n) = P_o(f,n)/P_i(f,n);
    end

    [R_L_pi_max(f), R_L_max_index(f)] = max(P_i(f,:));
    R_L_max_pi(f) = R_L(R_L_max_index(f));
    [R_L_po_max(f), R_L_max_index(f)] = max(P_o(f,:));
    R_L_max_po(f) = R_L(R_L_max_index(f));
    [R_L_eff_max(f), R_L_max_index(f)] = max(Eff(f,:));
    R_L_max_eff(f) = R_L(R_L_max_index(f));

```

```

for c=1:numel(C_opt) % optimal capacitance with optimal load
    Z_L_C_opt(f,c) = j*w(f)*L_22(f)+
    1/(j*w(f)*C_opt(c))+R_2(f)+R_L_max_eff(f);
    Z_2_C_opt(f,c) = 1/(1/Z_L_C_opt(f,c)+1/(j*w(f)*M(f)+R_m(f)));
    Z_eq_C_opt(f,c) = 1/(j*w(f)*C_opt(c)+
    j*w(f)*L_11(f)+R_1(f)+Z_2_C_opt(f,c);
    P_i_C_opt(f,c) = (0.5)*real(V_i^2/Z_eq_C_opt(f,c));
    P_o_C_opt(f,c) = real(P_i_C_opt(f,c))
    * real(Z_2_C_opt(f,c))/(R_1(f)+real(Z_2_C_opt(f,c)))
    * R_L_max_eff(f)/(R_2(f)+R_L_max_eff(f));
    Eff_C_opt(f,c) = P_o_C_opt(f,c)/P_i_C_opt(f,c);
end

[C_opt_pi_max(f), C_opt_max_index(f)] = max(P_i_C_opt(f,:));
C_opt_max_pi(f) = C_opt(C_opt_max_index(f));
[C_opt_po_max(f), C_opt_max_index(f)] = max(P_o_C_opt(f,:));
C_opt_max_po(f) = C_opt(C_opt_max_index(f));
[C_opt_eff_max(f), C_opt_max_index(f)] = max(Eff_C_opt(f,:));
C_opt_max_eff(f) = C_opt(C_opt_max_index(f));

for n = 1:numel(R_L) % optimal load with optimal capacitance
    Z_Lb(f,n) = j*w(f)*L_22(f)+1/(j*w(f)*C_opt_max_eff(f))+R_2(f)+R_L(n);
    Z_2b(f,n) = 1/(1/Z_Lb(f,n)+1/(j*w(f)*M(f)+R_m(f)));
    Z_eqb(f,n) = 1/(j*w(f)*C_opt_max_eff(f))+j*w(f)*L_11(f)+R_1(f)+Z_2b(f,n);
    P_ib(f,n) = (0.5)*real(V_i^2/Z_eqb(f,n));
    P_ob(f,n) = real(P_ib(f,n)) * real(Z_2b(f,n))/(R_1(f)+real(Z_2b(f,n)))
    * R_L(n)/(R_2(f)+R_L(n));
    Effb(f,n) = P_ob(f,n)/P_ib(f,n);
end

[R_L_pi_maxb(f), R_L_max_indexb(f)] = max(P_ib(f,:));
R_L_max_pib(f) = R_L(R_L_max_indexb(f));
[R_L_po_maxb(f), R_L_max_indexb(f)] = max(P_ob(f,:));
R_L_max_pob(f) = R_L(R_L_max_indexb(f));
[R_L_eff_maxb(f), R_L_max_indexb(f)] = max(Effb(f,:));
P_out_max_eff(f) = P_ob(f,R_L_max_indexb(f));
R_L_max_effb(f) = R_L(R_L_max_indexb(f));

end

figure(14)
subplot 211
plot(freq(66:381)/1000,R_L_max_effb(66:381))
grid on
grid minor
xlabel({'Frequency (kHz)'})
ylabel({'Optimal $R_L$ ($\Omega$)'})
subplot 212
plot(freq(66:381)/1000,C_opt_max_eff(66:381)*1e9)
grid on
grid minor
xlabel({'Frequency (kHz)'})
ylabel({'Optimal $C_{\text{resonance}}$ (nF)'})

```

```

figure(15)
plot(freq/1000, Eff_opt*100)
hold on
plot(freq/1000, R_L_eff_maxb*100)
grid on
grid minor
xlabel({'Frequency (kHz)'})
ylabel({'Efficiency (%)'})
legend({'Optimal Efficiency', 'Efficiency'}, 'Location', 'southeast')
legend('boxoff')

figure(16)
plot(freq/1000, P_out_max_eff/max(P_out_max_eff))
grid on
grid minor
xlabel({'Frequency (kHz)'})
ylabel({'Normalised Output Power (W)'})

```

B.2.2 Class-DE Theory

With the following MATLAB code it is possible to obtain the results relative to the previously presented class-DE theoretical formulations. from chapter 4.

```

%% Theoretical
clear all; clc;
close all;

RL = 1:0.001:25;
D = 0.34:0.01:0.48;

alpha = (2*pi*(1-2.*D) + sin(4*pi.*D)) ./ ((sin(2*pi.*D)).^2);
beta = (sin(2*pi.*D)).^2;

for i=1:length(RL)
    wCs(i,:) = beta./(RL(i)*pi);
end

for i=1:length(alpha)
    wLx(:,i) = beta(i).*alpha(i)./(2*pi*wCs(:,i));
end

lim=1;
RL_target = 1:1:7;
for i=1:length(RL_target)
    wCs_target(i,:) = beta./(RL_target(i)*pi);
    for n=1:length(D)
        if min(abs(wCs(i,n)-wCs_target(i,n)))<lim
            [wCs_target_value(i,n), wCs_target_index(i,n)]
            = min(abs(wCs(:,n)-wCs_target(i,n)));
            wCs_target_value(i,n) = wCs(wCs_target_index(i,n),n);
            wLx_target_value(i,n) = beta(n)*alpha(n)/(2*pi*wCs_target_value(i,n)); % 2*
        end
    end
end
end

```

```

figure(20)
hold on
plot(wCs, wLx)
for i=1:length(wLx(1,:))
    str=sprintf('D = %0.2f', D(i));
    text(wCs(1,i),wLx(1,i),str)
end
hold on
plot(wCs_target_value',wLx_target_value','--o')
for i=1:length(RL_target)
    str=sprintf('Req = %0.1f', RL_target(i));
    text(wCs_target_value(i,1) , 3+wLx_target_value(i,1),str)
end
ylabel({'wLx'})
xlabel({'wCs'})

%% Matching
Cs = [10/2 22/2 47/2 62/2]*1e-9;
for m=1:length(Cs)

    clearvars -except Cs m; clc;
    %close all;

    D = 0.25:0.00001:0.49;
    RL = 1:0.1:25;

    dataCsv = load(
        'OCEAN2freq_kreal_kimag2_ltx_rtx_lrx_rrx_l11_r11_l22_r22_lm_rm_salt.csv');
    freq = dataCsv(:,1); w0=2*pi*freq;
    L_1=dataCsv(:,4); L_2=dataCsv(:,4);
    R_1=dataCsv(:,9); R_2=dataCsv(:,9);
    k_real=dataCsv(:,2); k_imag=dataCsv(:,3);
    M=k_real.*sqrt(L_1.*L_2);
    L_11=L_1-M; L_22=L_2-M;
    R_m = dataCsv(:,13);
    C = 68*1e-9;
    fmin = 143e3; fmax = 900e3;
    [fmin, fminIndex] = min(abs(freq-fmin));
    [fmax, fmaxIndex] = min(abs(freq-fmax));
    fInterval = fminIndex:fmaxIndex;
    f = repmat(freq(fInterval),1,length(RL));
    w = repmat(w0(fInterval),1,length(RL));
    L = repmat(L_1(fInterval),1,length(RL));
    R1 = repmat(R_1(fInterval),1,length(RL));
    R2 = repmat(R_1(fInterval),1,length(RL));
    Rm = repmat(R_m(fInterval),1,length(RL));
    k = repmat(k_real(fInterval),1,length(RL));
    RL_ = repmat(RL,size(k,1),1);

    Li = L-1./(w.^2*C);
    div = w.^2.*Li.^2+(R2+RL_+Rm).^2;
    L22 = ( (Rm.*(Li-k.*L)+k.*L.*(R2+RL_)) .* (R2+RL_+Rm)
        - Li.*Rm.*(R2+RL_) + w.^2.*Li.*k.*L.*(Li-k.*L) ) ./ div;
    R22 = ( w.^2.*Li.*( Rm.*(Li-k.*L)+k.*L.*(R2+RL_) )

```

```

+ Rm.*(R2+RL_).*(R2+RL_+Rm) - w.^2.*k.*L.*(Li-k.*L).*(R2+RL_+Rm) ) ./ div;
Leq = L22+Li-k.*L;
Req = R22+R1;

beta = (sin(2*pi.*D)).^2;
for i=1:length(D)
    R_needed(:,i) = beta(i)./(w(:,1).*pi*Cs(m));
end

alpha = (2*pi*(1-2.*D) + sin(4*pi.*D)) ./ ((sin(2*pi.*D)).^2);
for i=1:length(D)
    Lx(:,i) = beta(i)*alpha(i)./(2*pi.*(w(:,1)).^2.*Cs(m));
end

for r=1:length(RL)
    for i=1:length(D)
        lim = 1e-6;
        if min(abs(Lx(:,i)-Leq(:,r)))<lim
            [w_target_value_L(i), w_target_index_L(i)]
            = min(abs(Lx(:,i)-Leq(:,r)));
            w_target_value_L(i) = w(w_target_index_L(i),1);
            f_target_value_L(i) = w_target_value_L(i)/(2*pi);
        end
        lim = 10;
        if min(abs(R_needed(:,i)-Req(:,r)))<lim
            [w_target_value_R(i), w_target_index_R(i)]
            = min(abs(R_needed(:,i)-Req(:,r)));
            w_target_value_R(i) = w(w_target_index_R(i),1);
            f_target_value_R(i) = w_target_value_R(i)/(2*pi);
        end
    end
end

lim = 3000;
if min(abs(f_target_value_L - f_target_value_R))<lim
    [f_final(r), f_final_index]
    = min(abs(f_target_value_L - f_target_value_R));
    f_final(r) = f_target_value_L(f_final_index);
    Req_aux(r) = Req(w_target_index_L(f_final_index),r);
    Leq_aux(r) = Leq(w_target_index_L(f_final_index),r);
    D_final(r) = D(f_final_index);
    if (f_final(r)==f(length(f)))
        f_final(r) = NaN;
        D_final(r) = NaN;
    end
    f_final(r);
    D_final(r);
    wCs(r) = 2*pi*f_final(r)*Cs(m);
    wLx(r) = 2*pi*f_final(r)*Leq_aux(r);
else
    f_final(r) = NaN;
    D_final(r) = NaN;
end
end
end

```

```

for i=1:length(wCs)
    if (wCs(i) == 0)
        wCs(i) = NaN;
    end
end
for i=1:length(wLx)
    if (wLx(i) == 0)
        wLx(i) = NaN;
    end
end

CL0 = [0 0 0]/256;
CL1 = [0 113 188]/256;
CL2 = [119 172 48]/256;
CL3 = [217 83 25]/256;
CL4 = [0.5 0.5 0.5];
CL5 = [0 0 0]/256;
LineWidth = 2.2;
LineWidthx = 0.5;
AxesLineWidth = 1.4;
MarkerSize = 18;
FontSize = 12;
set(0,'defaulttextInterpreter','latex');
set(groot, 'defaultAxesTickLabelInterpreter','latex');
set(groot, 'defaultLegendInterpreter','latex');
CL0 = [0 0 0]/256;
CL1 = [0 113 188]/256;
CL2 = [119 172 48]/256;
CL3 = [217 83 25]/256;
CL4 = [0.5 0.5 0.5];
CL5 = [0 0 0]/256;
LineWidth = 2.2;
LineWidthx = 0.5;
AxesLineWidth = 1.4;
MarkerSize = 18;
FontSize = 12;
set(0,'defaulttextInterpreter','latex');
set(groot, 'defaultAxesTickLabelInterpreter','latex');
set(groot, 'defaultLegendInterpreter','latex');

figure(1)
subplot 211
hold on; box on; grid on; grid minor;
plot(RL,f_final/1000)
ylabel({'frequency (kHz)'})
xlabel({'$R_L$ ($\Omega$)'})
set(gca,'Linewidth',AxesLineWidth,'FontSize',FontSize,
,'XMinorTick','on','YMinorTick','on');
subplot 212
hold on; box on; grid on; grid minor;
plot(RL,D_final)
ylabel({'$D$'})
xlabel({'$R_L$ ($\Omega$)'})
set(gca,'Linewidth',AxesLineWidth,'FontSize',FontSize,
,'XMinorTick','on','YMinorTick','on');

```



```

figure
hold on
plot(RL,Leq_aux*1e6)
ylabel({'$L_{eq}$ ($\mu H$)'})
xlabel({'$R_L$ ($\Omega$)'})

figure
hold on
plot(RL,Req_aux)
ylabel({'$R_{eq}$ ($\Omega$)'})
xlabel({'$R_L$ ($\Omega$)'})

figure(20)
hold on
plot(wCs,wLx)
str=sprintf('2$C_s$ = %d nF', 2*Cs(m)*1e9);
text(nanmedian(wCs),max(wLx),str)

end

```


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