

Imperial College London  
Department of Electrical and Electronic Engineering

# **Design of High Frequency Inductive Power Transfer Systems for Integration into Applications**

Juan Manuel Arteaga Sáenz

September 2019

A thesis submitted for the degree of Doctor of Philosophy of  
Imperial College London and the Diploma of Imperial College.



## Abstract

High frequency inductive power transfer (HF-IPT) is a technology with the potential of enabling wireless power to existing and emerging battery powered devices across multiple domains, such as consumer electronics, health-care, electric transport, security and defence. For over a decade, research groups around the world have managed to demonstrate HF-IPT as a proof of concept; however, the lack of presence of the technology in medium or high power commercial products highlights the difficulty of integrating it into real world scenarios.

This thesis is concerned, primarily, with the interaction and trade-offs between the power conversion stages of HF-IPT systems, ranging from the mains to the battery. The system model is structured so that crucial features for commercial exploitation, such as tolerance to misalignment and tolerance to load variation, are considered at each individual stage. Firstly, existing and novel candidate high-frequency power converters are proposed and prototyped to deal with challenges that emerge depending on the application at hand. With these prototype converter blocks, experimental work at the system level includes achieving 88 % DC-DC efficiency at 6.78 MHz and an air-gap of one coil radius. This, to the author's knowledge is the highest end-to-end efficiency achieved in HF-IPT under similar circumstances. Experiments also demonstrate a system architecture, introduced herein, where Class E and Class EF inverters are fed directly from the mains to eliminate one power conversion stage at the transmit side when the mains interface is taken into account.

A technique for HF-IPT design is then introduced with the formulation of a probability-based model which include motion and spatial freedom as design variables, allowing not only for the system to be designed and assessed for variable coupling, but also to consider the motion profile of the device being charged in the system optimisation.

Two HF-IPT systems powering an application are then demonstrated. The first consists of wireless power to a drone without a battery while it hovers over a charging-pad, and the second, wireless charging of an electric scooter. These demonstrations were made possible by implementing the main contributions of this work.





## Declarations

I herewith certify that all material in this thesis which is not my own work has been properly acknowledged.

The copyright of this thesis rests with the author. Unless otherwise indicated, its contents are licensed under a Creative Commons Attribution-Non Commercial 4.0 International Licence (CC BY-NC). Under this licence, you may copy and redistribute the material in any medium or format. You may also create and distribute modified versions of the work. This is on the condition that: you credit the author and do not use it, or any derivative works, for a commercial purpose. When reusing or sharing this work, ensure you make the licence terms clear to others by naming the licence and linking to the licence text. Where a work has been adapted, you should indicate that the work has been changed and describe those changes.

Please seek permission from the copyright holder for uses of this work that are not included in this licence or permitted under UK Copyright Law.



## Acknowledgements

I have had the privilege of spending four years of my life working on expanding the boundaries of knowledge in science and engineering. Achieving this has only been possible with the support of my family, friends and colleagues.

First, I would like to express my gratitude to my supervisor, Prof. Paul D. Mitcheson, for his exceptional technical and analytical support throughout the project. Paul, besides a supervisor, has also been a Tier Point adviser, Tignanello sponsor and a karting rival who finally beat me on the track in July 2019. Our karting rivalry is now 19-1. Having Paul as my PhD supervisor “has really made a difference”. I would also like to express my gratitude to Dr. David C. Yates, who supervises our technical work, and is leading the group in spinning out our technology from the university.

It has been a privilege to work with very talented researchers and postgraduate students at the Wireless Power Lab from Imperial College London in the last four years. Dr. Samer Aldhaher and Dr. George Kkelis were very supportive in the lab in the early stages of the programme. We won awards together and made the most of many conferences, especially WPTC 2016 in Aveiro. Dr. Christopher H. Kwan and Mr. Lingxin Lan have been excellent collaborators in research projects (and co-organisers of WPW 2019) in the last couple of years. I look forward to continue working with them and the newer members of the lab (also WPW 2019 organisers) Mr. Ioannis Nikifrodos and Mr. Nunzio Pucci, in years to come.

I would like to thank my friends from other departments from Imperial College London, especially Minos Skountzos, Giulia Miracca and Dafni Daskalaki-Mountanou. They have been great friends, travel buddies, flatmates, co-authors and they even starred in one of my short films.

I would like to thank my parents, Aurora Sáenz and Juan Eduardo Arteaga, my sister Irene Arteaga, my special lady friend Heidi Lees, my grandparents and the rest of my family for their unconditional support. I would like give a special recognition to Elke Ruge, and Amanda, Elena and Faris Hawila for making me part of their family in the UK and hosting me on multiple occasions during the last four years.

The work presented in this thesis has been financially sponsored by a scholarship from the PINN Programme by the Ministry of Science, Technology and ICT of Costa Rica (MICITT), a bursary from the Department of Electrical and Electronic Engineering of Imperial College London, and a scholarship from the University of Costa Rica (UCR). The research has been sponsored by the Engineering and Physical Sciences Research Council (EPSRC), the Centre for Power Electronics UK, and the UK Government.

Juan M. Arteaga

## Dedication

To Daisy Shelby Prado.



# Contents

|                                                                                                  |           |
|--------------------------------------------------------------------------------------------------|-----------|
| <b>Abstract</b>                                                                                  | <b>3</b>  |
| <b>Acknowledgements</b>                                                                          | <b>7</b>  |
| <b>List of Abbreviations</b>                                                                     | <b>31</b> |
| <b>Preface</b>                                                                                   | <b>35</b> |
| <b>1 Introduction</b>                                                                            | <b>37</b> |
| 1.1 Fundamental Principles that Enable Wireless Power Transfer . . . . .                         | 37        |
| 1.2 Reproduction of One of Faraday’s Experiments . . . . .                                       | 38        |
| 1.3 Techniques for Wireless Power Transfer . . . . .                                             | 39        |
| 1.3.1 Radiative Wireless Power Transfer . . . . .                                                | 40        |
| 1.3.2 Wireless Power Through Capacitive Coupling . . . . .                                       | 41        |
| 1.3.3 Wireless Power Through Magnetic Coupling . . . . .                                         | 42        |
| 1.4 Electromagnetic Exposure Guidelines . . . . .                                                | 42        |
| 1.5 A Brief Review of Commercial Products With Near-Field Wireless Charging Capability . . . . . | 43        |

|          |                                                                        |           |
|----------|------------------------------------------------------------------------|-----------|
| 1.6      | Recent Developments in High Frequency Inductive Power Transfer Systems | 45        |
| 1.7      | Challenges for HF-IPT Systems to be Integrated in Commercial Products  | 46        |
| <b>2</b> | <b>Theoretical Basis of High Frequency Inductive Power Transfer</b>    | <b>48</b> |
| 2.1      | Introduction . . . . .                                                 | 48        |
| 2.2      | Characterisation of the IPT Link . . . . .                             | 49        |
| 2.3      | Optimising the Link for Maximum Efficiency . . . . .                   | 52        |
| 2.3.1    | Resonating the Receiver With an External Capacitor . . . . .           | 53        |
| 2.3.2    | Resonating the Receiver With Self-Resonant Coils . . . . .             | 54        |
| 2.3.3    | The Optimal Load . . . . .                                             | 55        |
| 2.3.4    | Maximum Link Efficiency . . . . .                                      | 55        |
| 2.3.5    | Q-factor of the Coils and the Frequency of Operation . . . . .         | 56        |
| 2.4      | Power Conversion Stages Before and After an IPT Link . . . . .         | 57        |
| 2.4.1    | The Transmit Coil Driver . . . . .                                     | 57        |
| 2.4.2    | The Receiver Induced Electromotive Force Rectifier . . . . .           | 58        |
| 2.4.3    | Auxiliary Power Conversion Stages . . . . .                            | 58        |
| 2.5      | Summary . . . . .                                                      | 59        |
| <b>3</b> | <b>Power Electronic Circuits at Both Ends of a HF-IPT System</b>       | <b>61</b> |
| 3.1      | Introduction . . . . .                                                 | 61        |
| 3.2      | Resonant Inverters for HF-IPT . . . . .                                | 62        |
| 3.2.1    | Class E Inverters . . . . .                                            | 63        |



|       |                                                                                                |    |
|-------|------------------------------------------------------------------------------------------------|----|
| 3.2.2 | Class EF Inverters . . . . .                                                                   | 68 |
| 3.2.3 | Push-Pull Class E and Class EF Inverters . . . . .                                             | 70 |
| 3.2.4 | Power Throughput Control of Class E and Class EF Inverters . .                                 | 71 |
| 3.3   | Rectifiers for HF-IPT . . . . .                                                                | 72 |
| 3.3.1 | Current-Driven Class D Rectifiers for HF-IPT . . . . .                                         | 73 |
| 3.3.2 | Class E Rectifiers for HF-IPT . . . . .                                                        | 75 |
| 3.3.3 | Push-Pull Class E Rectifiers for HF-IPT . . . . .                                              | 76 |
| 3.3.4 | Resistance Compression Networks on HF-IPT Rectifiers . . . . .                                 | 77 |
| 3.3.5 | Current-Driven Class D Rectifiers with Resistance Compression<br>Networks for HF-IPT . . . . . | 78 |
| 3.3.6 | Multiplier Class D Configurations for HF-IPT . . . . .                                         | 80 |
| 3.4   | Inherent Regulation in Resonant Converters for Variable Loads . . . . .                        | 81 |
| 3.5   | Design of the Circuits for Experimental Verification . . . . .                                 | 82 |
| 3.5.1 | Inverter I: Load Tolerant Class E at 13.56 MHz for 150 W . . . . .                             | 82 |
| 3.5.2 | Inverter II: Load Independent Class EF at 13.56 MHz for 150 W .                                | 83 |
| 3.5.3 | Inverter III: Push-Pull Load Independent Class EF at 13.56 MHz<br>for 350 W . . . . .          | 84 |
| 3.5.4 | Rectifier I: RCN Class D at 13.56 MHz for 150 W . . . . .                                      | 84 |
| 3.5.5 | Rectifier II: 3x Voltage-Multiplier Full-Wave Class D at 13.56 MHz<br>for 350 W . . . . .      | 85 |
| 3.6   | Construction of Circuit Prototypes for Experimental Verification . . . . .                     | 86 |
| 3.6.1 | General Considerations in the Components Selection . . . . .                                   | 86 |

|          |                                                                                                                                   |           |
|----------|-----------------------------------------------------------------------------------------------------------------------------------|-----------|
| 3.6.2    | General Considerations in the Design of the Printed Circuit Boards                                                                | 87        |
| 3.7      | Construction of a Setup for Experimental Verification of HF-IPT . . . . .                                                         | 88        |
| 3.7.1    | Laboratory Equipment . . . . .                                                                                                    | 88        |
| 3.7.2    | Laboratory Safety . . . . .                                                                                                       | 89        |
| 3.7.3    | Auxiliary Printed Circuit Boards . . . . .                                                                                        | 90        |
| 3.7.4    | Measurements of the Proposed Prototype Inverters Without a Coupled Receiver . . . . .                                             | 91        |
| 3.8      | Summary . . . . .                                                                                                                 | 92        |
| <b>4</b> | <b>Practical HF-IPT Link Operation and Characterisation</b>                                                                       | <b>93</b> |
| 4.1      | Introduction . . . . .                                                                                                            | 94        |
| 4.2      | Design of a HF-IPT Link with Air-Core Coils . . . . .                                                                             | 95        |
| 4.2.1    | Coil Design Considerations . . . . .                                                                                              | 95        |
| 4.2.2    | Description of the Designed Coils . . . . .                                                                                       | 95        |
| 4.2.3    | Coil Self-Inductance and Equivalent Series Resistance . . . . .                                                                   | 97        |
| 4.2.4    | Coupling Distribution in Space . . . . .                                                                                          | 97        |
| 4.3      | Design and Construction of an Inductively Coupled Test Load . . . . .                                                             | 98        |
| 4.3.1    | Tuning the HF-IPT Test Load at 13.56 MHz . . . . .                                                                                | 100       |
| 4.3.2    | Controlled Detuning of the HF-IPT Test Load . . . . .                                                                             | 102       |
| 4.4      | Experiment I: Initial Investigations into the Effects Produced by Reflected Reactances in Candidate 13.56 MHz Inverters . . . . . | 103       |
| 4.4.1    | Introduction . . . . .                                                                                                            | 103       |

|       |                                                                                             |     |
|-------|---------------------------------------------------------------------------------------------|-----|
| 4.4.2 | Experimental Verification . . . . .                                                         | 104 |
| 4.4.3 | Discussion . . . . .                                                                        | 106 |
| 4.5   | Experiment II: HF-IPT with Alternative Media Wireless Gaps . . . . .                        | 107 |
| 4.5.1 | Introduction . . . . .                                                                      | 108 |
| 4.5.2 | Inductance and Resistance Measurements . . . . .                                            | 108 |
| 4.5.3 | HF-IPT End-to-End Experiments at 13.56 MHz . . . . .                                        | 109 |
| 4.5.4 | Discussion . . . . .                                                                        | 111 |
| 4.6   | Experiment III: Reflected Impedance Estimation Using Waveforms . . . . .                    | 113 |
| 4.6.1 | Introduction . . . . .                                                                      | 113 |
| 4.6.2 | The Reflected Impedance Estimation Principle . . . . .                                      | 113 |
| 4.6.3 | The Reflected Impedance Estimation Technique Using a Class EF<br>Inverter . . . . .         | 114 |
| 4.6.4 | Verification of the Reflected Impedance Estimation Model . . . . .                          | 115 |
| 4.6.5 | Discussion . . . . .                                                                        | 117 |
| 4.7   | Experiment IV: Characterisation of Corroded HF-IPT Coils . . . . .                          | 118 |
| 4.7.1 | Introduction . . . . .                                                                      | 118 |
| 4.7.2 | Association of Power Dissipation on the Coils and Corrosion . . . . .                       | 119 |
| 4.7.3 | Corrosion of the Immersion Silver Finish of the Coils Under Test . . . . .                  | 119 |
| 4.7.4 | Experimental Work Part I: Corroding the Coils Under Test . . . . .                          | 120 |
| 4.7.5 | Experimental Work Part II: DC-DC HF-IPT Testing . . . . .                                   | 121 |
| 4.7.6 | Experimental Work Part III: Materials Characterisation of the Coils<br>Under Test . . . . . | 122 |

|          |                                                                                                                                      |            |
|----------|--------------------------------------------------------------------------------------------------------------------------------------|------------|
| 4.7.7    | Experimental Results of a HF-IPT system Operating with a Corroded Coil . . . . .                                                     | 123        |
| 4.7.8    | Investigation of the Corrosion Mechanism of the Coils Under Test                                                                     | 125        |
| 4.7.9    | Conclusion . . . . .                                                                                                                 | 129        |
| 4.8      | Summary . . . . .                                                                                                                    | 130        |
| <b>5</b> | <b>Simplified HF-IPT Systems AC-Mains Interface</b>                                                                                  | <b>131</b> |
| 5.1      | Introduction . . . . .                                                                                                               | 131        |
| 5.2      | Power Factor Correction and Power Throughput Control at the Receiver                                                                 | 133        |
| 5.3      | Description of the Proposed System Architecture . . . . .                                                                            | 136        |
| 5.4      | Experimental Verification . . . . .                                                                                                  | 139        |
| 5.4.1    | Case Study I: Driving Class E and Class EF Inverters from a 60 Hz Sinusoidal Input Voltage . . . . .                                 | 140        |
| 5.4.2    | Case Study II: Changing the Reflected Load to a Push-Pull Class EF Inverter When Powered From a 60 Hz Sinusoidal Input Voltage .     | 143        |
| 5.4.3    | Case Study III: Operating the Proposed Architecture Using a Large DC-Link Capacitor After the IPT-Rectifier Instead of the PFC Stage | 145        |
| 5.4.4    | Case Study IV: Driving the Inverters from a 60 Hz Input Voltage and Feeding the Load with a Regulated DC Output Voltage . . .        | 147        |
| 5.5      | Summary . . . . .                                                                                                                    | 148        |
| <b>6</b> | <b>Analysis and Design for Dynamic HF-IPT Systems</b>                                                                                | <b>149</b> |
| 6.1      | Introduction . . . . .                                                                                                               | 149        |
| 6.2      | HF-IPT Systems with Variable Coupling . . . . .                                                                                      | 151        |

|       |                                                                                                                                         |     |
|-------|-----------------------------------------------------------------------------------------------------------------------------------------|-----|
| 6.2.1 | Formulating the HF-IPT System End-to-End Efficiency, Input Power and Output Power . . . . .                                             | 152 |
| 6.2.2 | Considering Time in the Formulation of the System Efficiency . .                                                                        | 154 |
| 6.2.3 | Characterising the Coupling Distribution in Space . . . . .                                                                             | 155 |
| 6.2.4 | Characterising the Coupling Distribution in Time . . . . .                                                                              | 156 |
| 6.2.5 | Optimising the Link for Variable Coupling Environments . . . . .                                                                        | 157 |
| 6.3   | Case Study I: Selecting the Alignment Range of Charging for a System With a Known Motion Profile and Without Power Throughput Control . | 158 |
| 6.3.1 | Characterising Performance with Respect to Coupling . . . . .                                                                           | 160 |
| 6.3.2 | Obtaining the Probability Density Function Based on a Known Motion Profile . . . . .                                                    | 161 |
| 6.3.3 | Energy-Efficiency and Output Energy with Respect to Misalignment Range . . . . .                                                        | 164 |
| 6.4   | Case Study II: Optimising a HF-IPT System with Power Throughput Control and Unpredictable Motion . . . . .                              | 166 |
| 6.4.1 | An Overview of the Wirelessly Powered Drone as an Experimental Platform . . . . .                                                       | 166 |
| 6.4.2 | Characterising the Range of Motion of the Wirelessly Powered Drone                                                                      | 168 |
| 6.4.3 | A First Iteration Design of the Rectifier for the Wirelessly Powered Drone . . . . .                                                    | 169 |
| 6.4.4 | Obtaining the Probability Distribution of a Wirelessly Powered Drone from Observation . . . . .                                         | 171 |
| 6.4.5 | Optimised Design of the Rectifier for the Wirelessly Powered Drone                                                                      | 172 |

|          |                                                                                                 |            |
|----------|-------------------------------------------------------------------------------------------------|------------|
| 6.5      | Summary . . . . .                                                                               | 173        |
| <b>7</b> | <b>End-to-End Application Demonstrations of HF-IPT Systems</b>                                  | <b>175</b> |
| 7.1      | Introduction . . . . .                                                                          | 175        |
| 7.2      | The Wirelessly Powered Drone . . . . .                                                          | 176        |
| 7.2.1    | Introduction to Wireless Power for UAVs . . . . .                                               | 177        |
| 7.2.2    | Designed System Overview . . . . .                                                              | 177        |
| 7.2.3    | The HF-IPT System Architecture and Power Throughput Control<br>for the UAV . . . . .            | 178        |
| 7.2.4    | Design of an Inductive Link for Wireless Charging of UAVs . . . .                               | 180        |
| 7.2.5    | Range of Motion of the Wirelessly Powered Drone . . . . .                                       | 182        |
| 7.2.6    | Simulations of Electromagnetic Exposure . . . . .                                               | 182        |
| 7.2.7    | The Inverter Designed to Wirelessly Power a Drone . . . . .                                     | 183        |
| 7.2.8    | The Rectifier Designed to Wirelessly Power a Drone . . . . .                                    | 184        |
| 7.2.9    | A SPICE Simulation on Changes in Load and Coupling . . . . .                                    | 188        |
| 7.2.10   | Measurements on the Wirelessly Powered Drone Demonstrator . .                                   | 190        |
| 7.2.11   | Additional Experiments With Wirelessly Powered Drones . . . . .                                 | 191        |
| 7.2.12   | An Alternative Rectifier for the Wirelessly Powered Drone: A Syn-<br>chronous Class E . . . . . | 192        |
| 7.2.13   | Experiments with a HF-IPT Transmitter on a Drone . . . . .                                      | 193        |
| 7.2.14   | Scaled-up Version of the Wirelessly Powered Drone . . . . .                                     | 194        |
| 7.3      | The Wirelessly Charged E-Scooter . . . . .                                                      | 194        |

|          |                                                                                                  |            |
|----------|--------------------------------------------------------------------------------------------------|------------|
| 7.3.1    | Introduction to Wireless Power to an E-Scooter . . . . .                                         | 195        |
| 7.3.2    | Designed System Overview . . . . .                                                               | 196        |
| 7.3.3    | HF-IPT System Architecture and Power Throughput Control for<br>the E-Scooter . . . . .           | 196        |
| 7.3.4    | Design of an Inductive Link for Wireless Charging of an E-Scooter                                | 197        |
| 7.3.5    | Simulations of Electromagnetic Exposure . . . . .                                                | 198        |
| 7.3.6    | HF-IPT Inverter and Rectifier for the E-Scooter Wireless Charger                                 | 200        |
| 7.3.7    | Measurements on the E-Scooter Demonstrator . . . . .                                             | 200        |
| 7.3.8    | IPT System in Scooter with Battery Charger and Li-ion Battery<br>Powered from Mains AC . . . . . | 204        |
| 7.3.9    | Scaled-up Version of the Wirelessly Charged E-scooter . . . . .                                  | 204        |
| 7.4      | Summary . . . . .                                                                                | 205        |
| <b>8</b> | <b>Conclusion</b>                                                                                | <b>207</b> |
| 8.1      | Overview and Main Findings . . . . .                                                             | 207        |
| 8.2      | Author's Contribution . . . . .                                                                  | 209        |
| 8.3      | Patents, Publications, Awards and Outreach . . . . .                                             | 211        |
| 8.3.1    | Patent Applications . . . . .                                                                    | 212        |
| 8.3.2    | Journal Publications as Lead Author . . . . .                                                    | 212        |
| 8.3.3    | Conference Presentations and Publications as Lead Author . . . .                                 | 212        |
| 8.3.4    | Conference Publications as Co-Author . . . . .                                                   | 214        |
| 8.3.5    | Awards . . . . .                                                                                 | 215        |

|                                           |            |
|-------------------------------------------|------------|
| 8.3.6 Outreach . . . . .                  | 216        |
| 8.4 Suggestions for Future Work . . . . . | 217        |
| <b>References</b>                         | <b>221</b> |



# List of Tables

|     |                                                                                               |     |
|-----|-----------------------------------------------------------------------------------------------|-----|
| 1.1 | SAR limits for general public and occupational exposure [ $\text{W kg}^{-1}$ ] . . . . .      | 43  |
| 3.1 | Candidate inverters and rectifiers for experimental verification . . . . .                    | 82  |
| 3.2 | Component values of Inverter I: Load-tolerant Class E . . . . .                               | 83  |
| 3.3 | Component values of Inverter II: Load-independent Class EF . . . . .                          | 84  |
| 3.4 | Component values of Inverter III: Push-pull load-independent Class EF .                       | 84  |
| 3.5 | Component values of Rectifier I: RCN Class D . . . . .                                        | 85  |
| 3.6 | Component values of Rectifier II: 3x voltage-multiplier full-wave Class D                     | 86  |
| 4.1 | Simulations and measurements of the coils at 13.56 MHz . . . . .                              | 97  |
| 4.2 | Test load tuning . . . . .                                                                    | 102 |
| 4.3 | Inductively coupled load tuning and measured reflected impedance . . . .                      | 103 |
| 4.4 | Input power at the transmit side with different reflected impedances [W]                      | 104 |
| 4.5 | Experimental results of DC to DC HF-IPT with alternative media wireless<br>gaps . . . . .     | 111 |
| 4.6 | Experimental results of DC to DC HF-IPT with a corroded and an uncor-<br>roded coil . . . . . | 124 |

|     |                                                                                                  |     |
|-----|--------------------------------------------------------------------------------------------------|-----|
| 4.7 | Average chemical composition (at.%) of the treated and untreated conductive layer . . . . .      | 128 |
| 6.1 | List of components of the circuits used in the Case Study I . . . . .                            | 159 |
| 6.2 | Constants values for the probability density function . . . . .                                  | 163 |
| 6.3 | Design parameters and components values of the rectifiers . . . . .                              | 173 |
| 6.4 | Average power consumption at the source with the different rectifier designs                     | 173 |
| 7.1 | Specifications of the coils for the wirelessly powered drone demonstrator at 13.56 MHz . . . . . | 182 |
| 7.2 | Design variables of the hybrid Class E Rectifier on the wirelessly powered drone . . . . .       | 188 |
| 7.3 | Components of the hybrid Class E rectifier for the wirelessly powered drone                      | 188 |
| 7.4 | Measurements of coils designed for the wireless charger of an E-scooter .                        | 198 |
| 7.5 | Components of the inverter and rectifier for the wireless charger for the E-scooter . . . . .    | 201 |
| 7.6 | Experimental results of the E-scooter wireless charging system . . . . .                         | 202 |

# List of Figures

|     |                                                                                                                             |    |
|-----|-----------------------------------------------------------------------------------------------------------------------------|----|
| 1.1 | Experimental setup to verify magnetic induction using a galvanometer. . .                                                   | 39 |
| 2.1 | General equivalent circuit of an IPT link. . . . .                                                                          | 49 |
| 2.2 | General equivalent circuits of an IPT link with induced electromotive forces<br>representing the magnetic coupling. . . . . | 50 |
| 2.3 | General equivalent circuits of an IPT link with a reflected impedance. . .                                                  | 51 |
| 2.4 | Equivalent circuits of an IPT system with a resonant receiver. . . . .                                                      | 54 |
| 2.5 | Maximum link efficiency as a function of $k^2 Q_p Q_s$ . . . . .                                                            | 57 |
| 2.6 | Two-port representation of the inverter. . . . .                                                                            | 58 |
| 2.7 | Two-port representation of the rectifier. . . . .                                                                           | 58 |
| 2.8 | Typical IPT system configuration. . . . .                                                                                   | 59 |
| 3.1 | IPT system DC-DC block diagram. . . . .                                                                                     | 62 |
| 3.2 | Output impedance of an inverter for HF-IPT, including the reflected impedance<br>from the receiver. . . . .                 | 62 |
| 3.3 | Class E inverter circuit topology. . . . .                                                                                  | 63 |
| 3.4 | Theoretical waveforms of a ZVS Class E inverter designed for ZDS and<br>with a 50 % duty cycle. . . . .                     | 65 |

|      |                                                                                                                                                                                                           |    |
|------|-----------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|----|
| 3.5  | Theoretical waveforms of a Class E inverter under suboptimal operation.<br>The overloaded condition plotted is when $R_{eq}$ is exceeded at 150 %. Optimal operation is plotted as a dashed line. . . . . | 66 |
| 3.6  | Theoretical waveforms of the Load-tolerant ZVS Class E inverter. The waveforms of the traditional Class E inverter are plotted with a dashed line. . . . .                                                | 67 |
| 3.7  | Class EF inverter circuit topology. . . . .                                                                                                                                                               | 68 |
| 3.8  | Theoretical waveforms of a load independent Class EF inverter at full load. The theoretical waveforms of the inverter when unloaded are plotted with a dashed line. . . . .                               | 69 |
| 3.9  | Push-Pull Class EF inverter circuit topology. . . . .                                                                                                                                                     | 70 |
| 3.10 | Equivalent circuit of the receive side of a HF-IPT system with $\mathcal{E}_s$ as the source of power and $\mathbf{Z}_{sec}$ as the load. . . . .                                                         | 72 |
| 3.11 | Current-driven Class D rectifier circuit topologies. . . . .                                                                                                                                              | 73 |
| 3.12 | Current-driven Class D rectifier circuit topologies in an IPT receiver. . . . .                                                                                                                           | 74 |
| 3.13 | Selected Class E rectifier circuit topologies. . . . .                                                                                                                                                    | 75 |
| 3.14 | Circuit topology of a push-pull hybrid Class E rectifier. . . . .                                                                                                                                         | 76 |
| 3.15 | Resistance compression network conceptual circuit. . . . .                                                                                                                                                | 77 |
| 3.16 | Class D rectifier with a resistance compression network. . . . .                                                                                                                                          | 78 |
| 3.17 | Input resistance of the RCN-Class D rectifier. . . . .                                                                                                                                                    | 79 |
| 3.18 | A 3x voltage-multiplier full-wave Class D rectifier for high output voltage HF-IPT. . . . .                                                                                                               | 80 |
| 3.19 | Photographs of the inverter prototypes. . . . .                                                                                                                                                           | 86 |
| 3.20 | Photographs of the rectifier prototypes. . . . .                                                                                                                                                          | 86 |

|      |                                                                                                                     |     |
|------|---------------------------------------------------------------------------------------------------------------------|-----|
| 3.21 | Photographs of the experimental setup for HF-IPT. . . . .                                                           | 88  |
| 3.22 | Photograph of the input power measuring board. . . . .                                                              | 90  |
| 3.23 | Photographs of additional boards used in the experiments. . . . .                                                   | 91  |
| 3.24 | Drain voltage of candidate inverters. . . . .                                                                       | 91  |
| 4.1  | PCB coils modelled using CST Microwave Studio. . . . .                                                              | 96  |
| 4.2  | Built prototype of the PCB coils. . . . .                                                                           | 96  |
| 4.3  | General equivalent circuit of an IPT link. . . . .                                                                  | 97  |
| 4.4  | Photograph of the coils and setup to measure coupling. . . . .                                                      | 98  |
| 4.5  | Coupling measurements. . . . .                                                                                      | 99  |
| 4.6  | HF-IPT test load circuit diagram. . . . .                                                                           | 99  |
| 4.7  | Photographs of the HF-IPT test load. . . . .                                                                        | 100 |
| 4.8  | Experimental tuning of the test load. . . . .                                                                       | 101 |
| 4.9  | Measured waveforms of Inverter I: Class E driving different loads. . . . .                                          | 105 |
| 4.10 | Measured waveforms of Inverter II: Class EF driving different loads. . . . .                                        | 106 |
| 4.11 | Photograph of IPT experimental setup. . . . .                                                                       | 107 |
| 4.12 | Photographs of the test rig developed for the characterisation of an inductive link with alternative media. . . . . | 109 |
| 4.13 | Measured resistance and reactance of the transmit coil coupled to the receivers under test. . . . .                 | 110 |
| 4.14 | DC to DC HF-IPT system performing wireless power transfer through alternative media. . . . .                        | 111 |

|      |                                                                                                        |     |
|------|--------------------------------------------------------------------------------------------------------|-----|
| 4.15 | Inverter drain voltage waveforms performing wireless power transfer through alternative media. . . . . | 112 |
| 4.16 | Theoretical waveforms of a load independent Class EF inverter with de-tuned loads. . . . .             | 114 |
| 4.17 | Results of the impedance prediction from simulation. . . . .                                           | 116 |
| 4.18 | Results of the impedance prediction from experiments. . . . .                                          | 117 |
| 4.19 | Photograph of the gradually corroded coils in the corrosion chamber. . .                               | 121 |
| 4.20 | Photograph of the experimental test rig for inductive power transfer. . .                              | 121 |
| 4.21 | Thermal images of the coils under test driving 4.3 A at 13.56 MHz. . . . .                             | 124 |
| 4.22 | Images obtained with a digital microscope (100x). . . . .                                              | 126 |
| 4.23 | Images of the untreated coil taken with a digital microscope at 500x. . .                              | 126 |
| 4.24 | Images of the treated coil taken with a digital microscope at 500x. . . . .                            | 127 |
| 4.25 | Electron microscopy images. . . . .                                                                    | 127 |
| 4.26 | X-ray diffraction patterns of the untreated and treated coils. . . . .                                 | 129 |
| 5.1  | DC to DC HF-IPT system diagram with constant $k$ . . . . .                                             | 133 |
| 5.2  | Mains to DC simplified typical HF-IPT system diagram. . . . .                                          | 134 |
| 5.3  | Mains to DC simplified HF-IPT system diagram with the PFC circuitry on the receiver. . . . .           | 134 |
| 5.4  | Mains to DC simplified typical HF-IPT system diagram with a regulated output. . . . .                  | 134 |
| 5.5  | Mains to DC typical and proposed HF-IPT system architecture. . . . .                                   | 136 |

|      |                                                                                                                                                     |     |
|------|-----------------------------------------------------------------------------------------------------------------------------------------------------|-----|
| 5.6  | Theoretical waveforms at the transmit side of a HF-IPT system using the proposed system architecture. . . . .                                       | 137 |
| 5.7  | Theoretical waveforms at the receive side of a HF-IPT system using the proposed system architecture. . . . .                                        | 137 |
| 5.8  | Theoretical waveforms of voltages and currents at key stages of the system. The current is plotted as a dotted line. . . . .                        | 138 |
| 5.9  | Theoretical waveforms of voltages and currents at key stages of the system driving a lighter load. The current is plotted as a dotted line. . . . . | 139 |
| 5.10 | Theoretical waveforms of voltages and currents at key stages of the system with lower $k$ . The current is plotted as a dotted line. . . . .        | 139 |
| 5.11 | 60 Hz AC to regulated DC HF-IPT systems experimental setup. . . . .                                                                                 | 140 |
| 5.12 | Experimental verification of the output current amplitude as a function of input voltage of inverters I and II. . . . .                             | 141 |
| 5.13 | Circuit topologies of Inverters I and II from mains AC. . . . .                                                                                     | 142 |
| 5.14 | Oscilloscope Traces of $V_{\text{rect}}$ (Channel 1), $i_{\text{rect}}$ (Channel 2) and $V_{ds}$ (Channel 3). . . . .                               | 143 |
| 5.15 | Circuit topology of Inverter III from mains AC. . . . .                                                                                             | 143 |
| 5.16 | Experimental verification of the output current amplitude as a function of input voltage of Inverter III. . . . .                                   | 144 |
| 5.17 | Experimental setup using Inverter III: Push-Pull Class EF from a low frequency AC voltage source. . . . .                                           | 144 |
| 5.18 | Experimental waveforms of Inverter III powered from a 60 Hz voltage source.                                                                         | 145 |
| 5.19 | Experimental setup to reflect a non-linear load to the transmitter. . . . .                                                                         | 146 |

|      |                                                                                                                                                                                                                 |     |
|------|-----------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|-----|
| 5.20 | Oscilloscope traces of $V_{ds}$ (Channels 1, 2), $i_o$ (Channel 3) and $i_{rect}$ (Channel 4). . . . .                                                                                                          | 147 |
| 5.21 | Experimental waveforms of the input voltage and current of the inverter when fed from an 60 Hz AC source, and loaded with the proposed rectifier followed by a PFC stage at the receiver and a DC load. . . . . | 148 |
| 6.1  | Block diagram of a HF-IPT system with variable coupling. . . . .                                                                                                                                                | 151 |
| 6.2  | General equivalent circuits of an IPT link with a reflected impedance considering changes in $k$ . . . . .                                                                                                      | 153 |
| 6.3  | Experimental setup for variable $k$ . Coils separation at 17.8 cm. . . . .                                                                                                                                      | 158 |
| 6.4  | Photograph of the linear slide used to change the alignment of the coils. .                                                                                                                                     | 159 |
| 6.5  | Power and efficiency measurements with respect to coupling. . . . .                                                                                                                                             | 160 |
| 6.6  | $k$ against linear misalignment (normalised to a coil diameter distance) for coil separations of 11 cm and 17.8 cm. . . . .                                                                                     | 164 |
| 6.7  | Calculation of output-energy and energy-efficiency with respect to range of $k$ . . . . .                                                                                                                       | 165 |
| 6.8  | Photograph of the wirelessly powered drone. . . . .                                                                                                                                                             | 167 |
| 6.9  | Block diagram of a HF-IPT system with variable coupling and an output voltage regulation stage. . . . .                                                                                                         | 167 |
| 6.10 | Cross section of the IPT-link describing the range of motion of the system.                                                                                                                                     | 168 |
| 6.11 | Wirelessly powered drone coupling distribution in space. . . . .                                                                                                                                                | 169 |
| 6.12 | Capture of a key frame for position tracking of the drone. . . . .                                                                                                                                              | 171 |
| 6.13 | Histogram of coupling obtained from 3600 samples in one minute. . . . .                                                                                                                                         | 172 |



|      |                                                                                                                                                                                                                                                                      |     |
|------|----------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|-----|
| 7.1  | Photograph of the wirelessly powered drone hovering over a charging pad.                                                                                                                                                                                             | 178 |
| 7.2  | Block diagram of a HF-IPT system (including reflected impedances) with variable coupling and an output voltage regulation stage. . . . .                                                                                                                             | 179 |
| 7.3  | Photograph of the off-the-shelf drone with its battery (left) next to the modified drone with a plated copper coil, the rectifier board and the DC-DC converter (right). . . . .                                                                                     | 181 |
| 7.4  | Coils diagrams for the wirelessly powered drone demonstrator. . . . .                                                                                                                                                                                                | 181 |
| 7.5  | SAR simulations with an $i_p$ of 5.5 A. . . . .                                                                                                                                                                                                                      | 183 |
| 7.6  | Experimental waveforms of the drain voltage of Inverter II: Class EF while wirelessly powering the drone at three throttle values. . . . .                                                                                                                           | 184 |
| 7.7  | Circuit topology of the implemented hybrid Class E rectifier. . . . .                                                                                                                                                                                                | 184 |
| 7.8  | Equivalent impedances of the receiver of an IPT system with the proposed impedance model of the hybrid Class E rectifier. . . . .                                                                                                                                    | 185 |
| 7.9  | SPICE simulation results performing step-changes between critical conditions in $k$ and load resistance. . . . .                                                                                                                                                     | 189 |
| 7.10 | Distribution of losses in the wirelessly powered drone demonstrator from simulation. . . . .                                                                                                                                                                         | 190 |
| 7.11 | Photograph of two wirelessly powered drones fed from a single transmitter.                                                                                                                                                                                           | 192 |
| 7.12 | Photograph of a DJI Matrice 100 drone with an on-board HF-IPT transmitter. . . . .                                                                                                                                                                                   | 193 |
| 7.13 | Photograph of the wirelessly charged drone taking off from the developed transmitter platform. The photograph was taken in the Maritime Autonomous Systems Trials and Evaluation Centre located at Turnchapel Wharf, Plymouth, UK, on the 11th of June 2019. . . . . | 194 |

|                                                                                                                                 |     |
|---------------------------------------------------------------------------------------------------------------------------------|-----|
| 7.14 E-Rider Model 18 Electric Scooter. . . . .                                                                                 | 196 |
| 7.15 E-Rider HF-IPT system block diagram. . . . .                                                                               | 197 |
| 7.16 Coils diagrams for the wireless charging E-scooter. . . . .                                                                | 198 |
| 7.17 SAR simulations results for the wirelessly charged E-scooter in critical cases.                                            | 199 |
| 7.18 Circuit topology of the inverter and rectifier implemented for the wirelessly<br>charged E-scooter. . . . .                | 201 |
| 7.19 Photographs of the E-scooter wireless charger prototype. . . . .                                                           | 202 |
| 7.20 Distribution of losses in the wireless charged E-scooter from simulation. .                                                | 203 |
| 7.21 Photograph of the 600 W wireless charging experiment of the on-board<br>battery of a Govecs S2.4 Electric Scooter. . . . . | 205 |
| 8.1 Photographs of the awards presentations. . . . .                                                                            | 216 |

# List of Abbreviations

|           |                                                               |
|-----------|---------------------------------------------------------------|
| AC        | alternating current                                           |
| CPT       | capacitive power transfer                                     |
| $C_{oss}$ | output capacitance of a transistor                            |
| DC        | direct current                                                |
| ESR       | equivalent series resistance                                  |
| FET       | field-effect transistor                                       |
| GaN       | gallium nitride                                               |
| HF-IPT    | high frequency inductive power transfer                       |
| ICNIRP    | International Commission on Non-Ionizing Radiation Protection |
| IEEE      | Institute of Electrical and Electronics Engineers             |
| IPT       | inductive power transfer                                      |
| ISM       | industrial, scientific and medical                            |
| LED       | light-emitting diode                                          |
| PFC       | power factor correction                                       |
| PCB       | printed circuit board                                         |
| PWM       | pulse-width modulation                                        |

|     |                                |
|-----|--------------------------------|
| RCN | resistance compression network |
| RF  | radio frequency                |
| RMS | root mean square               |
| SAR | specific absorption rate       |
| SiC | silicon carbide                |
| THD | total harmonic distortion      |
| UAV | unmanned aerial vehicle        |
| WPT | wireless power transfer        |
| ZVS | zero-voltage switching         |
| ZCS | zero-current switching         |
| ZDS | zero-derivative switching      |

# List of Symbols

Phasor quantities are bold

|                                |                                                  |
|--------------------------------|--------------------------------------------------|
| $j$                            | imaginary unit                                   |
| $\eta_{\text{link}}$           | IPT link efficiency                              |
| $\eta_{\text{rect}}$           | IPT rectifier efficiency                         |
| $\eta_{\text{dc-o}}$           | IPT system DC to DC efficiency                   |
| $\eta_{\mathcal{E}-\text{o}}$  | IPT receiver AC to DC efficiency                 |
| $\eta_{\text{dc}-\mathcal{E}}$ | IPT transmitter DC to AC efficiency              |
| $\eta_{\text{mains-reg}}$      | IPT system mains AC to regulated DC efficiency   |
| $\Phi_{\text{B}}$              | magnetic flux                                    |
| $C_{\text{s}}$                 | resonant capacitance at the receiver             |
| $\mathcal{E}$                  | induced electromotive force                      |
| $\mathcal{E}_{\text{s}}$       | induced electromotive force on the receive coil  |
| $\mathcal{E}_{\text{p}}$       | induced electromotive force on the transmit coil |
| $E_{\text{o}}$                 | IPT system output energy                         |
| $E_{\text{dc}}$                | IPT system input energy                          |

|                    |                                                     |
|--------------------|-----------------------------------------------------|
| $f_{\text{ip}}$    | transmit-coil current frequency                     |
| $i_{\text{p}}$     | transmit-coil current                               |
| $i_{\text{s}}$     | receive-coil current                                |
| $k$                | coupling factor                                     |
| $M_{\text{ps}}$    | mutual inductance of the transmit and receive coils |
| $L_{\text{p}}$     | self-inductance of the transmit coil                |
| $L_{\text{s}}$     | self-inductance of the receive coil                 |
| $R_{\text{p}}$     | equivalent-series-resistance of the transmit coil   |
| $R_{\text{s}}$     | equivalent-series-resistance of the receive coil    |
| $V_{\text{mains}}$ | mains voltage                                       |
| $V_{\text{rect}}$  | rectified-mains voltage                             |
| $V_{\text{dc}}$    | DC input voltage                                    |
| $V_{\text{o}}$     | IPT-rectifier output voltage                        |
| $V_{\text{reg}}$   | IPT system regulated output voltage                 |
| $P_{\text{o}}$     | IPT system output power                             |
| $P_{\text{dc}}$    | IPT system input power                              |
| $R_{\text{ac}}$    | IPT system AC load                                  |
| $R_{\text{o}}$     | IPT system DC load                                  |
| $R_{\text{eq}}$    | reflected resistance on the transmit coil           |
| $X_{\text{eq}}$    | reflected reactance on the transmit coil            |
| $Z_{\text{eq}}$    | reflected impedance on the transmit side            |
| $Z_{\text{sec}}$   | input impedance of the IPT-rectifier                |

# Preface

This thesis is organised as follows:

- Chapter 1 introduces fundamental concepts of wireless power transfer and the system integration challenges of the technology.
- Chapter 2 formulates the known theoretical principles of inductive power transfer.
- Chapter 3 introduces candidate power converter topologies and demonstrates the design and construction of the circuits.
- Chapter 4 introduces techniques to characterise and operate a high frequency inductive power transfer link with air-core coils in ideal and non-ideal conditions.
- Chapter 5 proposes an alternative system-level design where the inverter is fed directly from a low frequency alternating current source, representing the mains.
- Chapter 6 proposes a motion characterisation model using probability to design dynamic high frequency inductive power transfer systems for the highest energy-efficiency.
- Chapter 7 showcases the design and construction of two end-to-end high frequency inductive power transfer systems capable of wirelessly powering two applications.
- Chapter 8 concludes the thesis.





# Chapter 1

## Introduction

The content in this chapter comprises an introduction to high frequency inductive power transfer (HF-IPT) primarily focusing on the potential for commercial exploitation of the technology.

### 1.1 Fundamental Principles that Enable Wireless Power Transfer

Michael Faraday first demonstrated that an electric current could be generated from a magnetic source in 1831 [1]. By this time, André-Marie Ampère had already proven that magnetic force is generated from electric current [2], but at that point no reciprocal effects were yet directly documented. Faraday's hypothesis was then that since a magnetic force was generated from electric current, electric current should be somehow generated from magnetic action.

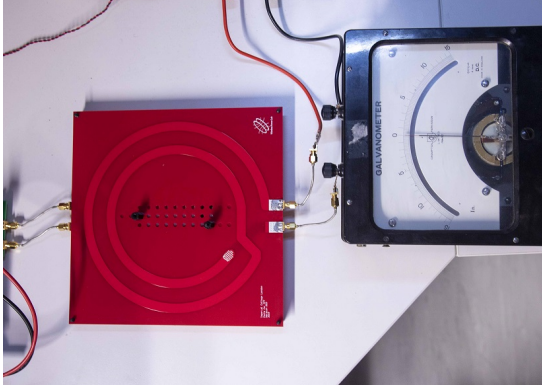
In his experiments, Faraday was able to detect electric current in an electric circuit powered by a variable magnetic source. The circuit under test consisted of solely passive components, namely a copper-wire coil connected to a current sensing device known as a galvanometer. The latter, first developed by Johann Schweigger in 1820 [3], was

the crucial instrument that allowed Faraday to confirm his hypothesis. Two magnetic power sources were utilised in the experiments: the first was a primary copper-wire coil connected to a battery, placed closely to the coil of the circuit under test. No effect was seen initially, but peaks of current were detected by the galvanometer when the battery in the primary circuit was either switched on or switched off. Small currents were also detected using a second power source comprised of a magnet in motion in proximity to the coil of the circuit under test. These experiments, proving magnetic induction, became the first demonstrations of one of the most important concepts of electrical engineering. Faraday's proven concept was soon after used for practical purposes, such as the invention of the dynamo in 1832 and the transformer in 1836.

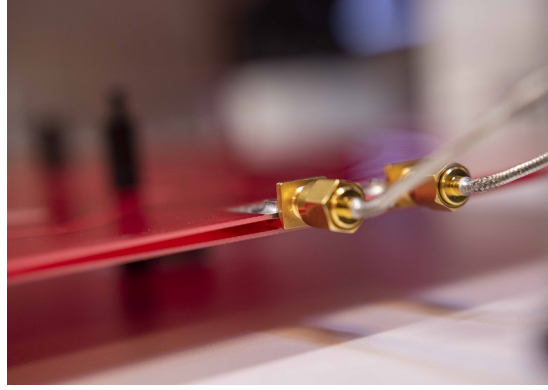
The mathematical representation of this phenomenon was not introduced by Faraday in his published work. This formulation, still used today, is credited to James Clerk Maxwell. Maxwell's published work in electromagnetism includes the Faraday-Maxwell equation, which describes the electromotive force generated in a conductor when a variable magnetic field is applied to it. He used Faraday's concept of lines of magnetic force [4], published in 1852 and proposed a mathematical model [5] in 1855 to describe it. Maxwell expanded his theory on electromagnetism in [6], introducing the concept of displacement current and later introduced electromagnetic waves in [7]. This mathematical model, published in 1865, became the theoretical basis of the study of electricity, magnetism and optics [3]; and that includes wireless power transfer (WPT).

## **1.2 Reproduction of One of Faraday's Experiments**

With the purpose of revisiting one of the experiments with which Faraday first observed electric current being generated from magnetic action, an experimental rig using the coils constructed for the following chapters of this thesis was built (Fig. 1.1). The equipment used in these experiments was a direct current (DC) power supply at the transmit side, two two-turn printed circuit board (PCB) coils (later discussed in Chapter 4), aligned



(a) Experimental setup using a galvanometer



(b) Detail of the coils placement

Figure 1.1: Experimental setup to verify magnetic induction using a galvanometer.

one on top of the other (Fig. 1.1), and at the receive side a galvanometer and a resistor connected in series with the receive coil. As in Faraday's experiment, it was expected that peaks of current at the receive coil would be measured when the power supply was switched on or off at the transmit side.

As expected, the experiments results show that when the power supply was switched on from 0 A to a maximum current of 11.55 A, a subtle movement in the galvanometer's needle, peaking at approximately  $50\text{ }\mu\text{A}$ , was observed, and conversely, at approximately  $-50\text{ }\mu\text{A}$  when the power supply was switched off.

### 1.3 Techniques for Wireless Power Transfer

Many power transfer mechanisms fit the broad definition of WPT. However, most often, the techniques which are associated with the term, are divided into two categories: far-field and near-field, or radiative and non-radiative wireless power transfer. These techniques directly employ the scientific breakthroughs of Ampère and Faraday, and the mathematical models of Maxwell to achieve WPT via magnetic induction (near-field), and alternatively by electromagnetic radiation (far-field).

There are alternative methods to inductive or radiative WPT, for example capacitive power transfer (CPT), which is discussed in Section 1.3.2, wireless power through acoustic

waves [8, 9], and systems which use lasers, or light to excite a photovoltaic receptor [10].

### 1.3.1 Radiative Wireless Power Transfer

Far-field WPT or radiative WPT consists of transferring power wirelessly using electromagnetic radiation. Transfer of power is done by driving a transmitting antenna with high frequency alternating current to create an electromagnetic beam from which the receiver can draw power. The receiver in such systems is comprised of a receiving antenna and a rectifier (the two together are referred to as a rectenna), and the load, which in itself may also include other power conversion stages for output power regulation.

In omnidirectional propagation WPT systems the end-to-end efficiency is a serious problem [11], due to the losses in the medium, product of the isotropic direction of propagation of electromagnetic waves. Antenna arrays and electromagnetic reflectors and lenses can be utilised to guide the electromagnetic waves from the transmitting point to the receiver, and draw more power from the same source, and thus achieve a higher efficiency. However, delimiting the propagation of energy to a narrow line of sight thereby increases the energy density of the electromagnetic waves, which depending on the environment and application, could be limited by regulations. In addition, an object's interruption of the line of sight would interrupt the transfer of power. Interesting solutions have been proposed with smart antennas, which can significantly improve the overall performance of radiative WPT systems for low power applications [12], as the possibility of creating a dynamic electromagnetic beam which can be optimised for different scenarios.

Even though radiative WPT systems are mostly proposed for low power applications, there are exceptions. In [13], for example, a far-field WPT system was employed to wirelessly power a 270 W unmanned aerial vehicle (UAV) while it hovered 15 m away from the transmitting antenna without any other source of power. This was an impressive achievement, especially considering the results were published in 1969; however, the low efficiency inherent of this WPT technique required a power transmission source

(2.45 GHz magnetron) of 5 kW. Also, a complex control system was required to direct the electromagnetic beam towards the UAV rectenna.

### 1.3.2 Wireless Power Through Capacitive Coupling

Capacitive Power Transfer (CPT) is one of the two technologies that are often referred to as near-field WPT or non-radiative WPT. In principle, CPT takes advantage of the capacitance between two electrodes in air to achieve WPT. These systems require two capacitances to create a circuit that separate the transmit and receive sides. The operation is similar to that of inductive power transfer (IPT), and similar topologies can be used in the power conversion circuits. The main difference with IPT is that instead of coils, capacitive systems require conductive plates to form the capacitance which links the transmit and receive sides.

Achievements of the technology were reviewed in [14] and positive results were reported for air-gaps smaller than 1 mm. Also, efficient power transfer at larger air-gaps can be achieved with additional compensating networks comprised of additional capacitances and inductances. For example, in [15], the additional network allowed 100 W of power to be transferred through an air-gap of 180 mm with a DC-DC efficiency of 66 %, where most of the losses are stated to be in the compensating networks required by the capacitive link. This technology has been proposed in high power applications, however the high intensity of the electric field is a challenge that still needs solving for this technology to be implemented in applications such as electric vehicles [16]. Achievements in terms of efficiency, power throughput and power density using CPT are still inferior to those found in the literature for IPT; but nonetheless, CPT is an interesting alternative to IPT where significant improvements have been shown recently [17, 18], and in some applications could be more convenient to integrate.

### 1.3.3 Wireless Power Through Magnetic Coupling

The most widely used near-field WPT technology is IPT. IPT can be utilised to transfer power through air-gaps of tens of centimetres or even a few meters [19]. The range of power transfer goes from a few milliwatts to several kilowatts, and end-to-end efficiencies higher than 90 % can be achieved [20].

The work presented in this thesis is concerned with HF-IPT systems, which operate at the first two industrial, scientific and medical (ISM) bands, 6.78 MHz and 13.56 MHz. Therefore, a theoretical formulation (Chapter 2), the practical considerations and the experimental verification of this technique are further developed in the following chapters of this thesis.

## 1.4 Electromagnetic Exposure Guidelines

Wireless power solutions such as the one in [13], where a UAV was powered through a radio frequency (RF) beam, are limited to very specific applications where firstly high efficiency is not a priority, and secondly, no humans are meant to be close to the electromagnetic beam. This is due to possible harmful effects from exposure to high intensity and high frequency electric and magnetic fields.

In order to determine a margin at which the magnetic and electric field strengths are considered safe to humans, the International Commission on Non-Ionizing Radiation Protection (ICNIRP) has published a widely accepted set of guidelines [21, 22]. At frequencies higher than 10 MHz, only thermal effects are considered to be of risk, and therefore the restrictions are only based on the specific absorption rate (SAR). This is the case for most of the experimental work presented in this thesis, which is at 13.56 MHz. Table 1.1 shows the SAR limits specified in [21], which also coincide with the regulations in the Institute of Electrical and Electronics Engineers (IEEE) C39.1-2005 standard [23]. For frequencies lower than 10 MHz, the electric field exposure caused by the internal

induced electromotive force also has to be assessed because of the potential for harmful effects on the nervous system.

Table 1.1: SAR limits for general public and occupational exposure [ $\text{W kg}^{-1}$ ]

|                                | General public | Occupational |
|--------------------------------|----------------|--------------|
| Whole body average SAR         | 0.08           | 0.4          |
| Localised SAR (head and trunk) | 2              | 10           |
| Localised SAR (limbs)          | 4              | 20           |

The ICNIRP and IEEE regulations for electric and magnetic field exposure, [21] and [23], respectively, give different incident reference field levels that assure compliance with the SAR limits specified in Table 1.1 and the limits of electric fields produced by the internal induced electromotive force. These reference field levels have been proposed to guarantee compliance in the worst cases. Nonetheless, since in near-field wireless power technologies the field intensity distribution is not uniform, it is more precise to evaluate exposure limits using simulations and specifically evaluate thermal (SAR) and non-thermal (internal induced electric fields) effects [24, 25, 26]. Examples of exposure simulations for HF-IPT systems integrated on an application are presented in Chapter 7.

## 1.5 A Brief Review of Commercial Products With Near-Field Wireless Charging Capabilitiy

Wireless power transfer is a desirable feature in most electronic devices in the market. It significantly simplifies the battery charging process to the consumer, and eliminates plugin power connectors which can be expensive and reliability-concerning, especially in applications such as portable electronic devices and electric vehicles.

IPT has had a successful commercial integration in low power applications, such as passive RFID [27] and the electric toothbrush [28]. Higher power systems, however, pose a more difficult technical challenge. Notwithstanding that some experiments proposing large

amounts of power to be transferred wirelessly date back to the decade of the 1970s, it was in the late 1990s that a research group from the University of Auckland, including John Boys, Andrew Green, Fred Nassenstein and later Grant Covic [29], successfully integrated a relatively high power IPT system to wirelessly power industrial machines [30].

In the latter half of the decade of the 2000s, a company called Witricity was spun out from the Massachusetts Institute of Technology (MIT), looking to commercialise IPT at high frequencies with self-resonant coils, after publishing promising results in *Nature* [31], a prestigious British multidisciplinary scientific journal. However, the self-resonant technology proposed in [31] was not released commercially, and now the company changed its focus to commercialising IPT for electric vehicles [32], a technology which is more mature and has been successfully demonstrated in real environments [33, 34], by building on the principles of operation of the first commercialised solutions by the group from the University of Auckland. Witricity, after acquiring Qualcomm Halo (also including wireless power transfer intellectual property from the University of Auckland) now holds more than 900 patents (and more than 550 pending) and is a leading entity in the development of wireless charging to electric vehicles standards [35].

In the portable electronic devices market, the Qi standard, which offers wireless power solutions for low power devices ( $<5\text{ W}$ ) using IPT at kilohertz frequencies, has had successful commercial exploitation, as there are more than 3500 products [36] with a Qi compatible wireless charging receiver. At these power levels, an alternative to Qi is the AirFuel Alliance [37] which proposes RF IPT or HF-IPT as the core technology. This standard, which already has a products lined up [38], may become more relevant in years to come. The advantages of operating at megahertz frequencies and recent improvements in high frequency power converters could trigger this technology to become a strong alternative to Qi in the low and mid-power wireless charging applications market.



## 1.6 Recent Developments in High Frequency Inductive Power Transfer Systems

Notwithstanding that the physical principles that govern IPT are the same at kilohertz and megahertz frequencies, HF-IPT is not achieved by solely increasing the frequency of a system designed to operate in the kilohertz range. The power conversion stages and the coils employed for HF-IPT can be fundamentally different to those designed for kilohertz IPT. The main reason to increase the frequency of operation of an IPT system to the first ISM bands (6.78 MHz and 13.56 MHz) is allowing for the link to be constructed with low inductance but high Q coils, and thus eliminate or minimise the use of magnetically permeable materials. HF-IPT coils can be constructed simply by shaping a wire conductor or PCB trace to the required form using one or a few turns. Because of the difference in the coils and the circuits employed at both ends of the system (Chapter 3), improvements in the state of the art of kilohertz IPT do not necessarily result in significant improvements in HF-IPT technology and conversely.

The most challenging aspect of HF-IPT is driving the transmit coil and rectifying the high frequency induced voltage on the receiver without significantly increasing the switching losses of the power converters. This requires opting for alternative power converter topologies. This topic is expanded in Chapter 3.

Most of the work recently published in the field of HF-IPT tackles challenges in one of the power conversion stages. The technology has significantly improved since the introduction of wide bandgap devices, namely silicon carbide (SiC) and gallium nitride (GaN) [39], into the market. Achieving high efficiency is possible with silicon mosfets, e.g. [40]; however, the high voltage and current ratings, the low conduction losses and the low gate and output capacitances of SiC and GaN devices are beneficial for the development of HF-IPT. GaN field-effect transistors (FETs), specifically, are very attractive for power converters that require switching at the first ISM bands since these can be switched with low gate voltage and charge.

The literature that covers demonstrations of existing end-to-end HF-IPT systems is not extensive. However, there are examples in which high end-to-end efficiencies are achieved, most notably and recently in [41, 42, 43]. The work in [43] showcases HF-IPT using single-turn air-core coils operating in the first three ISM bands and demonstrates an end-to-end efficiency higher than 85 % with an output power of 500 W at 6.78 MHz, using a Class D inverter. In addition to good efficiencies with simple air-core coils, notable achievements in control of HF-IPT systems can be found in the literature. For example, the low power (>5 W) battery to battery power transfer demonstrated in [44] was made possible by using a CMOS monolithic reconfigurable bidirectional transceiver which exchanges its configuration from a differential Class-D inverter when transmitting power, to a full-wave rectifier when it's receiving.

The maturity of HF-IPT systems has now improved to the point where it could soon start being integrated in commercial products.

## **1.7 Challenges for HF-IPT Systems to be Integrated in Commercial Products**

The process of integrating a new technology into a commercial application can be complicated and time consuming. Electronic commercial products require a number of certifications to ensure safety and to guarantee that the product does not interfere with other systems; for example, in the case of HF-IPT there should be no interference with RF communications. There are various ways of taking a product into the market: a company can simply deploy a product, if it has the means to do so, and grab the market by reaching it first; or alternatively, the technology can undertake a standardisation process to guarantee inseparability.

The work presented in this thesis aims to demonstrate the capabilities of the technology and showcase how it can operate whilst integrated into an application. In order to do

this, several techniques are investigated which have then fed into the proof-of-concept application examples.

## Chapter 2

# Theoretical Basis of High Frequency Inductive Power Transfer

The content in this chapter comprises a review of known theory, summarised and expressed to support the author's contributions in the following chapters.

### 2.1 Introduction

The basic principles of inductive power transfer (IPT) can be explained and formulated using linear circuit models which are practical for an application-driven design. A two-coil IPT system is used to derive the equations; however, the analysis can be extrapolated if there was a mutual inductance between more than two IPT coils. The analysis is presented considering a fixed frequency, and thus expressed in phasor notation.

The model presented does not account for non-linearities in the components, it treats the self inductance of the coils as constant (since the frequency is fixed and the coils are proposed to be air-cored) and does not consider parasitic effects such as capacitive coupling between the coils.

The phasor quantities are marked in bold.

## 2.2 Characterisation of the IPT Link

A two-coil IPT link can be modelled using the circuit diagram in Fig. 2.1. Here, a primary circuit, comprised of a current source  $i_p$  and a transmit coil (represented as a self-inductance  $L_p$  and an equivalent series resistance  $R_p$ ) feeds, by the means of magnetic coupling, a secondary circuit comprised of a receive coil ( $L_s$  and  $R_s$ ) and an impedance  $Z_{sec}$ . The exchange of power between the coils can be quantified using the Maxwell-

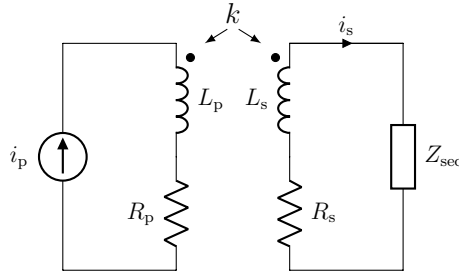


Figure 2.1: General equivalent circuit of an IPT link.

Faraday equation:

$$\mathcal{E} = -\frac{\partial \Phi_B}{\partial t}; \quad (2.1)$$

where an induced electromotive force ( $\mathcal{E}$ ) is generated from a variable magnetic flux ( $\Phi_B$ ), and Ampère's circuital law,

$$\oint_C \mathbf{B} \cdot d\boldsymbol{\ell} = \mu_0 i_{enc}, \quad (2.2)$$

where a circulating current generates a magnetic field, and henceforth a magnetic flux:

$$\Phi_B = \iint_s \mathbf{B} \cdot d\mathbf{s}. \quad (2.3)$$

The  $\mathcal{E}$  generated in a conductor is the superposition of the  $\mathcal{E}$  generated by its own circulating current (which defines self-inductance) and the  $\mathcal{E}$  generated by external sources

(which defines mutual inductance):

$$\mathcal{E} = \mathcal{E}_L + \mathcal{E}_{M_{L:1}} + \cdots + \mathcal{E}_{M_{L:n}} = -L \frac{di_L}{dt} + M_{L:1} \frac{di_1}{dt} + \cdots + M_{L:n} \frac{di_n}{dt}. \quad (2.4)$$

The  $\mathcal{E}$  in a conductor can therefore be modelled as a set of dependent voltage sources in series representing the sources of power external to the circuit and a self-inductance of the conductor. Thus, the inductively coupled circuits in Fig. 2.1 can be represented as in Fig. 2.2.

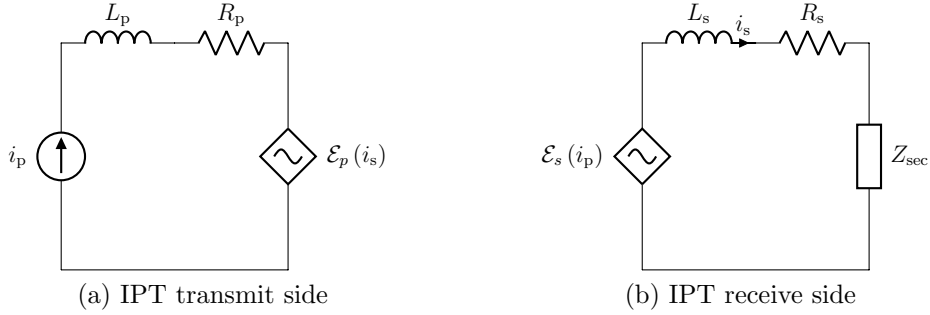


Figure 2.2: General equivalent circuits of an IPT link with induced electromotive forces representing the magnetic coupling.

If the frequency of operation is constant, an expression can be derived for the induced electromotive force on the receive coil ( $\mathcal{E}_s$ ), produced by the transmit-coil current ( $i_p$ ) in phasor notation, in terms of: the imaginary unit ( $j$ ), the transmit-coil current frequency ( $f_{i_p}$ ), and the mutual inductance of the transmit and receive coils ( $M_{ps}$ ):

$$\mathcal{E}_s = j2\pi f_{i_p} M_{ps} i_p. \quad (2.5)$$

If coupling factor ( $k$ ) is defined in terms of  $M_{ps}$ , the self-inductance of the transmit coil ( $L_p$ ) and the self-inductance of the receive coil ( $L_s$ ) as:

$$k = \frac{M_{ps}}{\sqrt{L_p L_s}}, \quad (2.6)$$

then (2.5) can be expressed as:

$$\mathcal{E}_s = j2\pi f_{i_p} k \sqrt{L_p L_s} i_p. \quad (2.7)$$

The induced electromotive force on the transmit coil ( $\mathcal{E}_p$ ) produced by the receive-coil current ( $i_s$ ):

$$i_s = \frac{\mathcal{E}_s}{Z_{\text{sec}} + R_s + j2\pi f_{i_p} L_s}, \quad (2.8)$$

can be calculated, considering that the frequency of  $i_s$  is equal to  $f_{i_p}$ , as:

$$\mathcal{E}_p = \frac{j2\pi f_{i_p} k \sqrt{L_p L_s} \mathcal{E}_s}{Z_{\text{sec}} + R_s + j2\pi f_{i_p} L_s}. \quad (2.9)$$

This allows  $\mathcal{E}_p$  to be calculated in terms of  $i_p$ :

$$\mathcal{E}_p = -\frac{4\pi^2 f_{i_p}^2 k^2 L_p L_s i_p}{Z_{\text{sec}} + R_s + j2\pi f_{i_p} L_s}, \quad (2.10)$$

and hence, as a reflected impedance on the transmit side ( $Z_{\text{eq}}$ ):

$$Z_{\text{eq}} = \frac{\mathcal{E}_p}{i_p}. \quad (2.11)$$

The equivalent circuit in Fig. 2.1 can therefore be restructured as shown in Fig. 2.3, where (2.7) defines the source of power on the receive side, and the reflected impedance

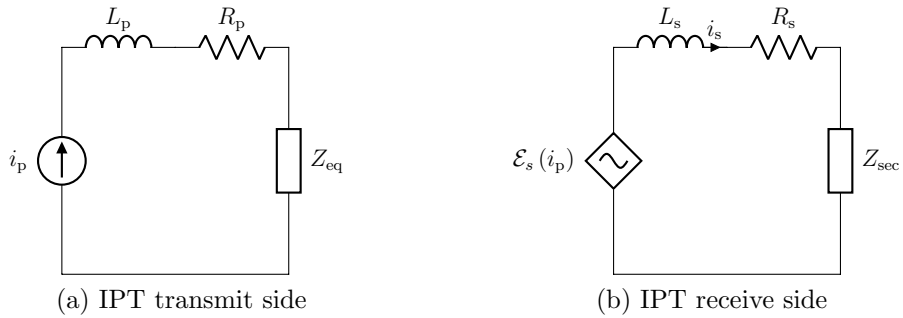


Figure 2.3: General equivalent circuits of an IPT link with a reflected impedance.

on the transmit side can be calculated by separating the real and imaginary elements of the impedances ( $\mathbf{Z}_{\text{sec}} = R_{\text{sec}} + \mathbf{j}X_{\text{sec}}$ , and  $\mathbf{Z}_{\text{eq}} = R_{\text{eq}} + \mathbf{j}X_{\text{eq}}$ ):

$$R_{\text{eq}} = \frac{4\pi^2 f_{\text{ip}}^2 k^2 L_{\text{p}} L_{\text{s}} (R_{\text{s}} + R_{\text{sec}})}{(R_{\text{s}} + R_{\text{sec}})^2 + (2\pi f_{\text{ip}} L_{\text{s}} + X_{\text{sec}})^2}, \quad (2.12)$$

and,

$$X_{\text{eq}} = -\frac{4\pi^2 f_{\text{ip}}^2 k^2 L_{\text{p}} L_{\text{s}} (2\pi f_{\text{ip}} L_{\text{s}} + X_{\text{sec}})}{(R_{\text{s}} + R_{\text{sec}})^2 + (2\pi f_{\text{ip}} L_{\text{s}} + X_{\text{sec}})^2}. \quad (2.13)$$

The IPT circuit model in Fig. 2.3 formulated by the equations (2.7), (2.12) and (2.13) can describe an IPT system where variables that are relatively easy to measure and optimise for are expressed for the circuits at both ends of the system.

## 2.3 Optimising the Link for Maximum Efficiency

The IPT link efficiency ( $\eta_{\text{link}}$ ), as modelled in Fig. 2.3, can be calculated as:

$$\eta_{\text{link}} = \frac{i_{\text{s}}^2 R_{\text{sec}}}{i_{\text{p}}^2 (R_{\text{p}} + R_{\text{eq}})}, \quad (2.14)$$

from which the expression:

$$\eta_{\text{link}} = \frac{P_{\text{o}}}{P_{\text{i}}} = \frac{4\pi^2 f_{\text{ip}}^2 k^2 L_{\text{p}} L_{\text{s}} R_{\text{sec}}}{4\pi^2 f_{\text{ip}}^2 k^2 L_{\text{p}} L_{\text{s}} (R_{\text{sec}} + R_{\text{s}}) + R_{\text{p}} \left( (R_{\text{sec}} + R_{\text{s}})^2 + (X_{\text{sec}} + 2\pi f_{\text{ip}} L_{\text{s}})^2 \right)}, \quad (2.15)$$

is derived.

Once the construction of the coils, the geometry of the link and the frequency of operation are determined, the impedance  $\mathbf{Z}_{\text{sec}}$  allows the link to be operated with different objectives. For example, some low-power systems take priority in maximising the output power [19, 45]; others, in applications such as biomedical implants, are required to be designed for minimum heat dissipation on the receive side [26, 46]; and most commonly,



$\mathbf{Z}_{\text{sec}}$  is designed for maximum link efficiency [40, 47, 48, 49].

There are two conditions which need to be satisfied in the design of  $\mathbf{Z}_{\text{sec}}$  to achieve the maximum link efficiency: resonating the receive coil, and achieving the optimal load.

### 2.3.1 Resonating the Receiver With an External Capacitor

From (2.15) the maximum link efficiency with respect to  $X_{\text{sec}}$  is achieved when:

$$\frac{\partial \eta_{\text{link}}(X_{\text{sec}_{\text{opt}}})}{\partial X_{\text{sec}}} = 0,$$

which results in:

$$X_{\text{sec}_{\text{opt}}} = -2\pi f_{\text{ip}} L_{\text{s}}. \quad (2.16)$$

This corresponds to resonating the receive coil with a negative reactance. When the receive-side is at resonance,  $\mathcal{E}_{\text{s}}$  and  $\mathbf{i}_{\text{s}}$  are in phase, and the reflected reactance on the transmit coil ( $X_{\text{eq}}$ ) is zero. Hence, for a fixed  $R_{\text{sec}}$  the magnitude of  $\mathbf{i}_{\text{s}}$  and consequently the output power, are maximised for a given magnitude of  $\mathbf{i}_{\text{p}}$ , when  $L_{\text{s}}$  is resonated out.

This condition is commonly achieved by introducing a capacitance across (or in series with) the receive coil. The selection between these two options to resonate  $L_{\text{s}}$  is dictated by the type of rectifier, which is expanded in Chapter 3. The circuit diagrams of an IPT system with a resonant receiver are shown in Fig. 2.4, with  $\mathbf{Z}_{\text{sec}}$  expressed as a resonant capacitance at the receiver ( $C_{\text{s}}$ ) and an IPT system AC load ( $R_{\text{ac}}$ ). It should be noted that in parallel-tuned IPT systems, achieving  $X_{\text{sec}_{\text{opt}}}$  depends not only on  $C_{\text{s}}$ , but also on  $R_{\text{ac}}$ .

The equivalent series resistance (ESR) losses of  $C_{\text{s}}$  are not taken into account in this model, since these tend to be very low in comparison to the losses in the coil, especially for high frequency inductive power transfer (HF-IPT) systems, where the capacitors used are most commonly based on the C0G (NP0) dielectric, and these have an “ultra-low”

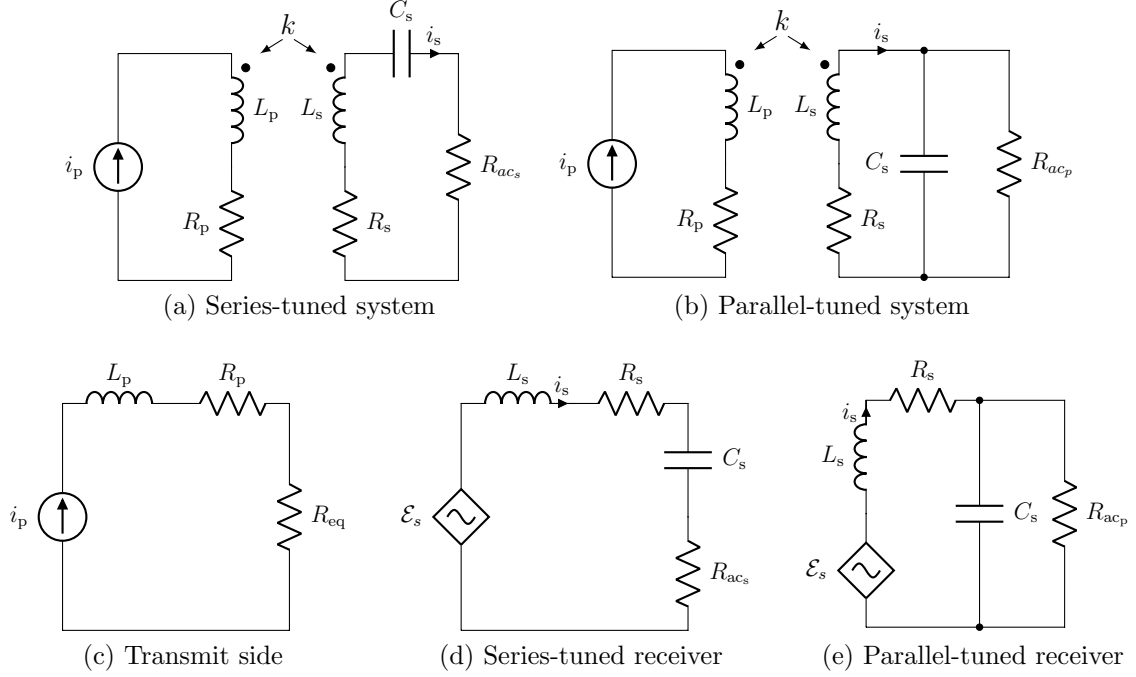


Figure 2.4: Equivalent circuits of an IPT system with a resonant receiver.

ESR [50].

### 2.3.2 Resonating the Receiver With Self-Resonant Coils

Notwithstanding that the principle of resonating the receive coil with a capacitance dates back to Tesla's experiments in the 19th century [51], and that the principle was demonstrated in industrial applications in the early 90s [29, 30]; researchers from the Massachusetts Institute of Technology, in [31], introduced self-resonant coils. These intrinsically possess a distributed capacitance in their construction (the distributed capacitance of an inductor can be calculated as in [52]), which resonates the self-inductance of the coil at a particular frequency. In their experiments, a link efficiency of 40 % was achieved over an air-gap of four coil diameters (i.e., a coupling factor lower than 2 %). That system was built using self-resonant coils with a Q factor of 950 at their resonating frequency, which was 9.9 MHz.

Self-resonant coils appear to be difficult to implement in commercial applications since they have a natural resonant frequency which depends on their construction; hence, they

do not allow for external tuning to change the resonant frequency. Also, practically, it is convenient to allow for changes in  $C_s$  in order to tune out the parasitic residual reactances of the power converters that drive the coils and possible changes in the self and mutual inductance caused once the system is integrated into an application.

### 2.3.3 The Optimal Load

The second criteria for the design of  $\mathbf{Z}_{\text{sec}}$  that maximises the link efficiency is the optimal value for  $R_{\text{sec}}$ , which can be found by solving:

$$\frac{\partial \eta_{\text{link}}(R_{\text{secopt}})}{\partial R_{\text{sec}}} = 0,$$

for  $R_{\text{sec}}$ . The optimal load can be formulated, generally, as:

$$R_{\text{secopt}} = \sqrt{\frac{4\pi^2 f_{\text{ip}}^2 k^2 L_p L_s R_s}{R_p} + (X_{\text{sec}} + 2\pi f_{\text{ip}} L_s)^2 + R_s^2}, \quad (2.17)$$

and as:

$$R_{\text{secopt}} \Big|_{X_{\text{secopt}}} = \sqrt{\frac{4\pi^2 f_{\text{ip}}^2 k^2 L_p L_s R_s}{R_p} + R_s^2}, \quad (2.18)$$

when the self-inductance of the coil is at resonance with  $X_{\text{sec}}$ .

### 2.3.4 Maximum Link Efficiency

The maximum link efficiency, achieved when  $\mathbf{Z}_{\text{sec}}$  is designed as specified in (2.16) and (2.17), is:

$$\eta_{\text{linkmax}} = \frac{4\pi^2 f_{\text{ip}}^2 k^2 L_p L_s}{\left( \sqrt{R_p R_s} + \sqrt{R_p R_s + 4\pi^2 f_{\text{ip}}^2 k^2 L_p L_s} \right)^2}. \quad (2.19)$$

From this expression, it can be noted that the parameters that determine the maximum achievable  $\eta_{\text{link}}$  are  $k$ ,  $f_{\text{ip}}$ , and the self-inductance and the ESR of the coils.

### 2.3.5 Q-factor of the Coils and the Frequency of Operation

In [47], the often quoted expression for maximum  $\eta_{\text{link}}$ , which can be derived from (2.19), is expressed as a function of  $k$  and the Q-factor of the coils:

$$\eta_{\text{link}_{\text{max}}} = \frac{k^2 Q_p Q_s}{(1 + \sqrt{1 + k^2 Q_p Q_s})^2}, \quad (2.20)$$

where the Q factor of the coils are:

$$Q_p = \frac{2\pi f_{\text{ip}} L_p}{R_p}, \quad (2.21)$$

and,

$$Q_s = \frac{2\pi f_{\text{ip}} L_s}{R_s}, \quad (2.22)$$

respectively. The Q-factor of the coils depends of the self-inductance and ESR (determined by shape, construction and materials), and the frequency of operation. The dimensionless  $k^2 Q_p Q_s$ , indicates the Q factor required to achieve good efficiencies for a particular value of coupling. Maximum  $\eta_{\text{link}}$  is shown as a function of  $k^2 Q_p Q_s$  in Fig. 2.5.

High efficiency at low  $k$  is a typically desirable feature in IPT system design since, fundamentally, lower values of  $k$  imply a greater separation between the coils, and thus higher spatial freedom.

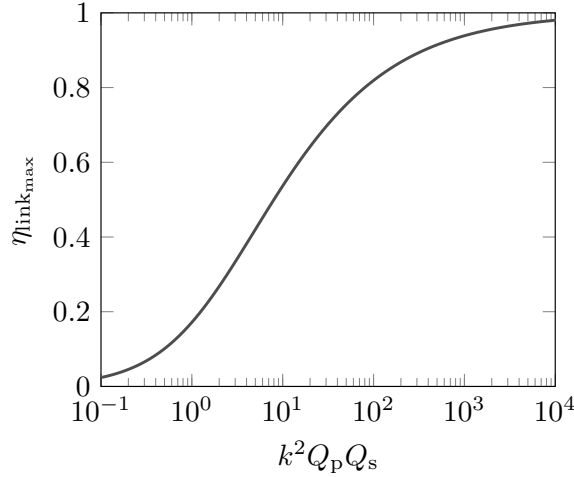


Figure 2.5: Maximum link efficiency as a function of  $k^2 Q_p Q_s$ .

## 2.4 Power Conversion Stages Before and After an IPT Link

The design of an IPT system is a bigger task than just the optimisation of the link. The power electronic circuits required to generate  $\mathbf{i}_p$  at the transmit side and emulate  $\mathbf{Z}_{sec}$  at the receive side (when a rectifier is employed) while delivering the required amount of power by the load from the power source available are also fundamental challenges in the design of the system.

### 2.4.1 The Transmit Coil Driver

The transmit coil driver of an IPT system, represented as a two port circuit in Fig. 2.6, is a power inverter fed from an external DC input voltage ( $V_{dc}$ ) which outputs a high frequency sinusoidal current  $\mathbf{i}_p$  at the frequency selected to drive the transmit coil  $f_{ip}$ .

There are a number of challenges in the design of an inverter for IPT, as for example, driving not only a resistive load, but the reactance of  $L_p$ , which in most cases needs to be tuned out (at least in part) with a capacitive or capacitive-inductive network to reduce the VA rating on the active components. An inverter for IPT is also expected to achieve

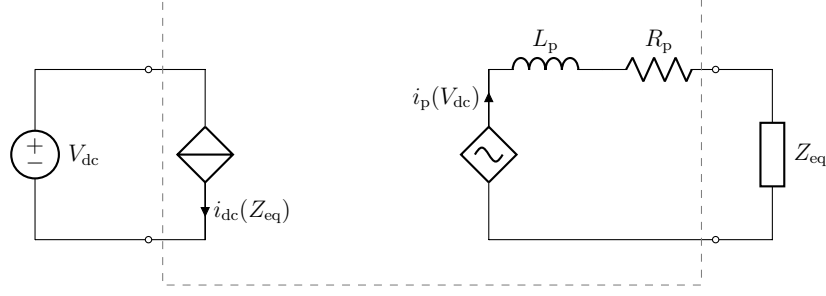


Figure 2.6: Two-port representation of the inverter.

high efficiency, low harmonic distortion, and tolerance to variations in the load.

### 2.4.2 The Receiver Induced Electromotive Force Rectifier

The first order linear approximation of the circuit that is required to transform the  $\mathcal{E}_s$  into a useful DC voltage, is shown in Fig. 2.7.

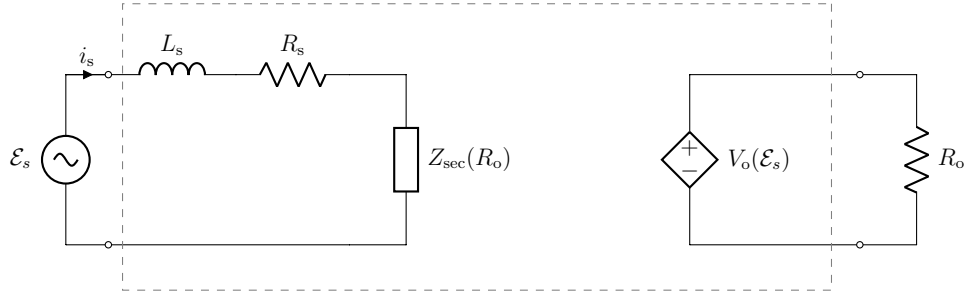


Figure 2.7: Two-port representation of the rectifier.

The design of a rectifier for IPT poses challenges which are somewhat similar to the ones of the inverter; for example, achieving good efficiency, providing a DC output voltage with low variations, and dealing with changes in both  $\mathcal{E}_s$  and the IPT system DC load ( $R_o$ ).

### 2.4.3 Auxiliary Power Conversion Stages

From Figs. 2.6 and 2.7, it can be observed that there is a relation between  $V_{dc}$  at the transmit side and  $R_o$  at the receive side. Once the IPT link is designed, the power delivered to the load can be regulated by varying the amplitude of  $i_p$ , or the resistance

of  $R_o$ . Thus, power throughput control can be performed at either side of the system, or as a combined effort.

The diagram in Fig. 2.8 exemplifies an IPT system architecture where power throughput control can be performed using additional power conversion stages at both ends of the system. Power throughput control can be done directly on the inverter or the rectifier with variations on the switching signals; however, in HF-IPT it is often preferred to operate the inverter and the rectifier in open-loop and include the additional power conversion blocks shown in Fig. 2.8. The reasoning of this preference is further explained in Section 3.2.4.

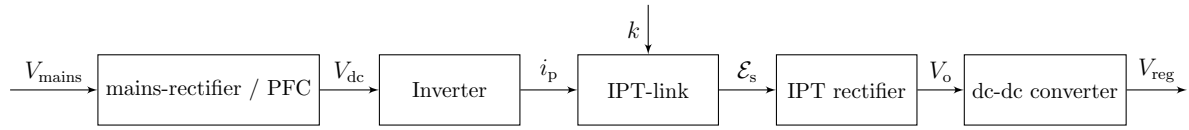


Figure 2.8: Typical IPT system configuration.

An end-to-end IPT system comprises multiple power conversion stages at both ends of the system. This not only results in many design variables to allow for optimisation in a particular application, but also raises challenges especially in achieving high efficiencies, considering that each individual power conversion stage incorporates losses to the system.

## 2.5 Summary

This chapter outlines the theoretical principles of IPT and formulates the interaction between the transmit and receive side circuits. The fundamental theory of IPT, summarised in this chapter, is further developed specifically for HF-IPT and tested experimentally in the following chapters of the thesis. Chapter 3 focuses on the design of the power conversion blocks before and after the HF-IPT link and Chapter 4 focuses on the characterisation of the interaction between the coils for different cases.





# Chapter 3

## Power Electronic Circuits at Both Ends of a HF-IPT System

The content in this chapter comprises a review of known and original power electronic circuit topologies and design techniques, an overview of the design parameters of some of the prototype circuits building blocks used later in the thesis, and original work involving resistance compression network (RCN) Class D rectifiers for inductive power transfer (IPT). This has been published in IEEE Transactions on Power Electronics: [53] and in two IEEE conference proceedings: [54, 55].

### 3.1 Introduction

This chapter is concerned with the electronic circuit topologies typically employed to transform to and from megahertz frequencies in high frequency inductive power transfer (HF-IPT) systems from DC to DC. Fig. 3.1 shows the fundamental power conversion stages of a HF-IPT system. The diagram includes three power conversion blocks: an inverter, which generates a high frequency transmit-coil current ( $i_p$ ), an IPT-link, which describes the properties of the coils, and an IPT rectifier which is in charge of transforming the induced electromotive force on the receive coil ( $\mathcal{E}_s$ ) into a DC voltage. Here, the

inputs of the system are the DC input voltage ( $V_{dc}$ ) and coupling factor ( $k$ ). The latter is formulated as an input in order to allow for variable coupling in the design of the system.

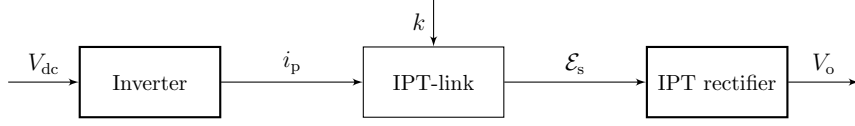


Figure 3.1: IPT system DC-DC block diagram.

This chapter first describes some of the main topologies for megahertz switching inverters and rectifiers, and then proposes a set of candidate topologies with different features, allowing choice in the inverter and the rectifier with the required features for a specific application. The candidate circuit designs proposed in this chapter are utilised in the following chapters of the thesis.

## 3.2 Resonant Inverters for HF-IPT

HF-IPT inverters are required to feed loosely coupled inductive links with high frequency alternating current, where the impedance to be driven is the sum of the impedance of the coil and the reflected impedance by the receive side, as shown in Fig. 3.2. Therefore, HF-IPT inverters require a tuning network to at least in part, resonate the transmit-coil inductance and drive the coil more efficiently (with a power factor closer to one) to reduce the VA rating of the active components.

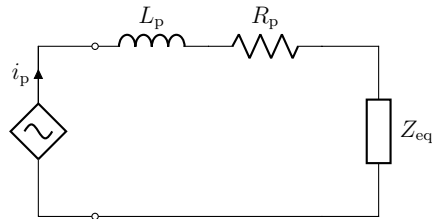


Figure 3.2: Output impedance of an inverter for HF-IPT, including the reflected impedance from the receiver.

There are two main families of inverters which are utilised in HF-IPT systems: Class D,

which are comprised of an input half or full-bridge switching network and a resonant circuit at the output; and alternatively, Class E, which have a single low-side switch (push-pull variations have two low-side switches) with an inductance at the input, from the source to the switch (similar to that of a boost converter) and an output resonant circuit.

Class D inverters are very popular at low frequency ( $<300$  kHz) IPT [56, 57, 34]. The output resonant tank can be designed in different configurations, achieving different responses to changes in coupling or load, and the input bridge-type switching network allows for elaborate control techniques with a broad range of operational benefits [58, 59]. At high frequencies, notwithstanding that newly developed wide-bandgap devices may allow for solutions where Class D inverters match or overcome the efficiencies achieved with Class E, e.g. [43], Class D inverters are less popular due to the difficulty of driving the high-side transistor efficiently [60, 61].

The three inverter designs proposed in this chapter, and utilised in the other chapters of the thesis, belong to the Class E family.

### 3.2.1 Class E Inverters

The Class E inverter (Fig. 3.3) was first introduced in [62]. Its input switching network

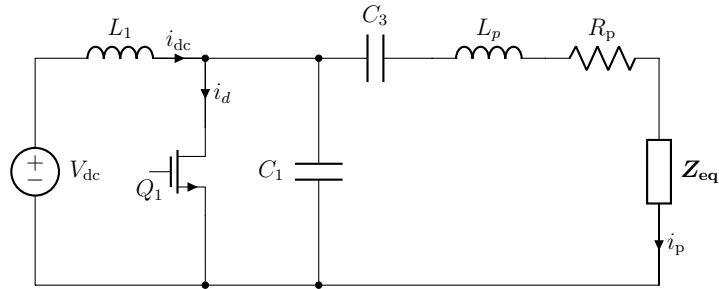


Figure 3.3: Class E inverter circuit topology.

is similar to that of the boost converter where a large inductance conducting DC current ( $i_{dc}$  in  $L_1$  in Fig. 3.3) is switched by a transistor (denoted as the field-effect transistor (FET) transistor  $Q_1$  in Fig. 3.3) to create an alternating current (AC) output. Unlike

a boost converter, where a half-wave rectifier is used to obtain a DC output voltage, a Class E inverter uses a resonant tank to drive a load (reflected impedance on the transmit side ( $\mathbf{Z}_{eq}$ ) in Fig. 3.3) with an AC current  $\mathbf{i}_p$ . In practice, this type of inverter is preferred at high frequencies because the required size of the reactive components ( $L_1$  in particular) decreases with frequency, and most importantly it is at high frequencies where the advantageous switching properties of this topology are exploited. Class E inverters can be designed either for zero-current switching (ZCS) and zero-voltage switching (ZVS) [63]. The ZCS Class E circuit topology diverts slightly from the one in Fig. 3.3 in that  $C_1$  is eliminated; however, especially at high frequencies, the output capacitance of a transistor ( $C_{oss}$ ) has to be accounted for, and since the topology used for ZVS Class E (Fig. 3.3) includes  $C_1$  across the drain to source of the transistor as a design variable, the inverter can be designed so that  $C_{oss}$  is included, or absorbed, in the parameter  $C_1$ . Thus, ZVS Class E inverters are most often preferred over the ZCS alternative at megahertz frequencies.

The circuit components shown in Fig. 3.3, along with the frequency and duty cycle at which  $Q_1$  is driven, can be utilised to design the inverter, considering the requirements and ranges of the load ( $\mathbf{Z}_{eq}$ ) and the restrictions of the parameters mentioned. In addition to ZVS, ZVS Class E inverters are designed to achieve optimum performance (maximum theoretical efficiency) when the zero-derivative switching (ZDS) condition (when the slope of the voltage is zero at turn-on) is also satisfied [63].

The power output capability, a variable that specifies how much power can be extracted with respect to the voltage and current ratings of a device, is a key parameter to consider since it restrains to an extent the range of the transistor's duty cycle: In ZVS Class E inverters the power output capability is highest at a 50 % duty cycle [63]; and thus, this parameter is often fixed at that value. In case the duty cycle is increased or decreased from 50 % to suit the requirements of a design, the power output capability decreases, therefore requiring a higher rated device for a given power demand.

In this topology, the power output capability is more crucial than in Class D topologies since as shown in the theoretical voltage and current waveforms of the transistor (Fig. 3.4),

the voltage across the switching device is much higher ( $\approx 3.5$  times) than the inverter's

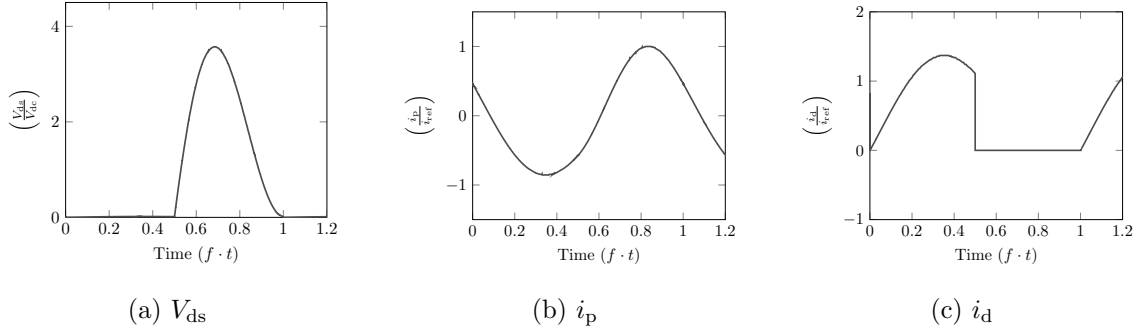


Figure 3.4: Theoretical waveforms of a ZVS Class E inverter designed for ZDS and with a 50 % duty cycle.

input voltage, therefore requiring, in many cases, high-voltage rated devices to achieve a relatively low output power. This is an important restriction in the device selection part of designing Class E inverters.

In summary, the optimum operation of a ZVS Class E inverter (waveforms in Fig. 3.4) is achieved by designing the components of the topology (Fig.3.3), the duty cycle of the switching device and frequency of operation, to meet the ZVS and ZDS conditions. These variables, however, are often restricted when the inverter is designed to drive a HF-IPT system. For example, the frequency of operation, in most designs, is fixed at an industrial, scientific and medical (ISM) band, the inductance of  $L_p$  depends of the construction of the coil, and hence the application's spatial restrictions, and the load ( $Z_{eq}$ ) depends on the link properties and design.

### Resistive Load Changes on Class E Inverters

The effects of changing the resistive element of the load have been extensively addressed in the literature: first, if the load resistance is decreased, the drain voltage increases drastically (Fig. 3.5 (a)), the output current increases (Fig. 3.5 (b)), and the switch conducts negative current (Fig. 3.5 (c)). Conversely, when the load resistance increases, ZVS is lost (Fig. 3.5 (d)), the output current decreases (Fig. 3.5 (e)), and the switch current decreases (Fig. 3.5 (f)).

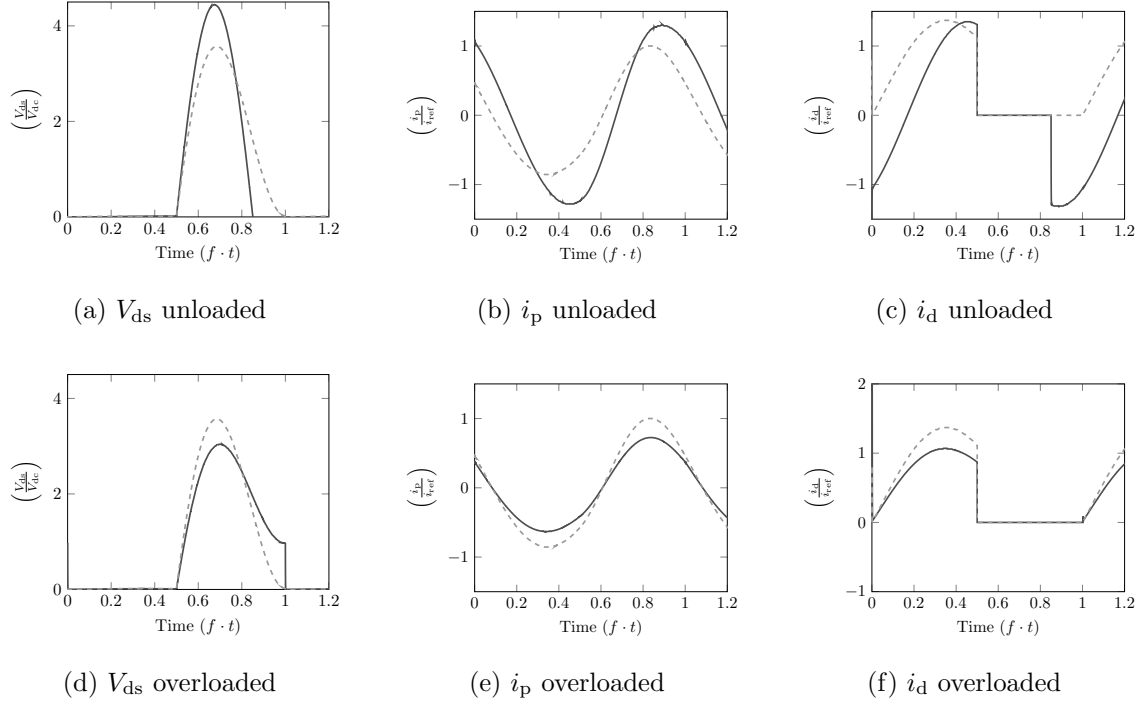


Figure 3.5: Theoretical waveforms of a Class E inverter under suboptimal operation. The overloaded condition plotted is when  $R_{eq}$  is exceeded at 150%. Optimal operation is plotted as a dashed line.

Load-independent solutions have been formulated for the Class E inverter [64], but nonetheless, these solutions give constant output voltage amplitude and not, as is required in HF-IPT, a constant output current amplitude, to achieve constant magnetic field amplitude. Moreover, in [65], authors conclude that the mathematical solutions for constant output current amplitude load independent Class E are unpractical to implement.

### A Load-Tolerant Design for HF-IPT

From the suboptimal operation modes of the ZVS Class E inverter (Fig. 3.5), the case in which ZVS is lost (when the load resistance is higher than its optimum one) involves a drastic increase in switching losses, and therefore should be avoided. Conversely, the case in which the load resistance is lower than the optimal one can be in cases tolerated, suggesting that alternative designs for this inverter without the ZDS condition could be possible.

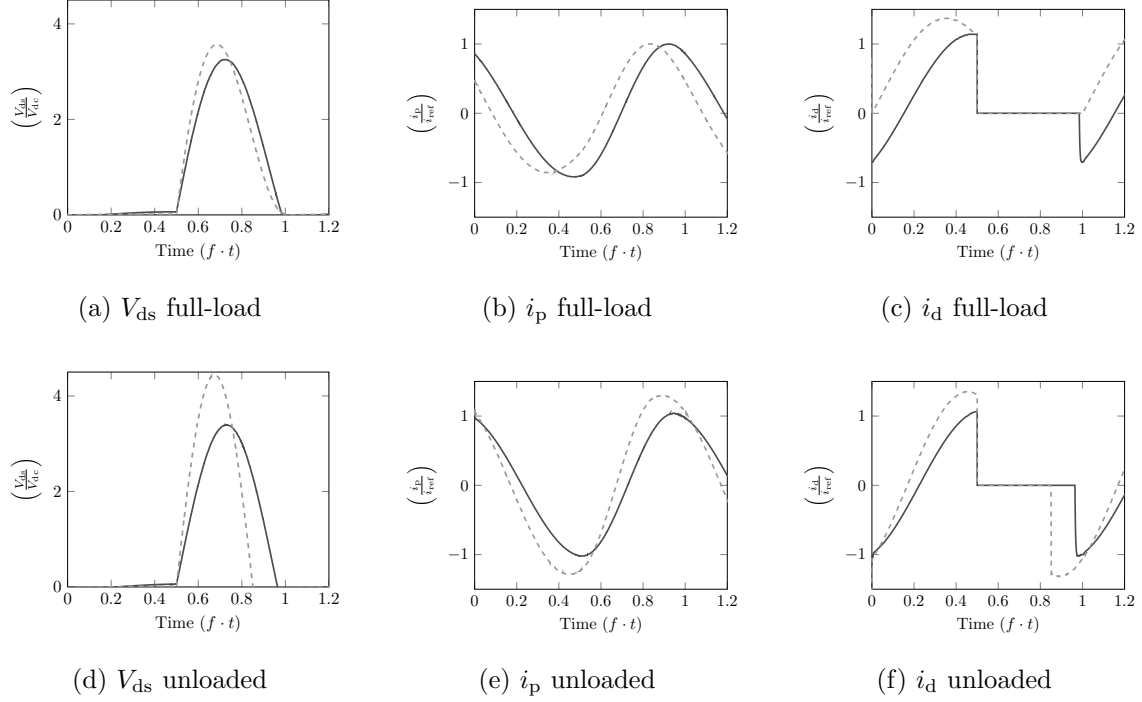


Figure 3.6: Theoretical waveforms of the Load-tolerant ZVS Class E inverter. The waveforms of the traditional Class E inverter are plotted with a dashed line.

Interestingly, if the ZVS Class E inverter is designed without the ZDS restriction, a solution with load tolerant properties can be found as follows. The tuning is done so that the anti-parallel diode of the transistor never conducts at full load, and that it conducts for a very short amount of time when unloaded. The waveforms in Fig. 3.6, and specifically the drain voltage at  $f \cdot t = 1$  at full load and unloaded, demonstrates how the performance of this proposed design, despite not being ideal in that the transistor conducts negative current at any load, is much less affected by changes in the resistive load, and hence showcases a more robust solution for HF-IPT.

### Alternative Designs for Class E Inverters

The aim of achieving optimum operation with the Class E inverter has led to design modifications to the resonant nature of the topology. An interesting approach, with respect to the ones previously mentioned, is accounting for  $L_1$  to be a part of the resonant circuit, as opposed to an infinite choke. This was first proposed in [66]. A finite choke Class E grants another variable to design with; however, this requires additional filtering

at the inverter's input since  $L_1$  then conducts substantial AC currents.

Another alternative design can be proposed by integrating additional reactive components to the resonant tank with the purpose of making impedance transformations [40, 48, 49, 67, 68]; however, additional components generate additional losses, and complexity, making the system more difficult to tune, and likely less efficient.

### 3.2.2 Class EF Inverters

The Class EF (Fig. 3.7) is a resonant inverter topology from the Class E family. It includes

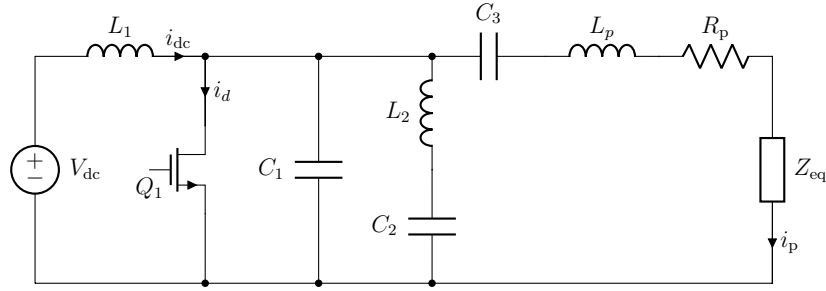


Figure 3.7: Class EF inverter circuit topology.

an LC network comprised of  $C_2$  and  $L_2$  which helps shape the voltage and current through the transistor. This LC network grants an additional two degrees of freedom in the circuit design.

This topology has been extensively analysed in literature [69, 70, 71] and design equations for different tuning methods have been proposed. The main purpose of integrating the additional LC branch (when compared to a Class E inverter) is to reduce the peak voltage from drain to source in the transistor. This is achieved by tuning  $C_2$  and  $L_2$  at the second or third harmonic of the switching frequency, allowing shaping the voltage across the transistor, and thus reducing the peak voltage and granting higher power output capability.

This topology is also known as  $\Phi_2$  when the tuning also includes a finite  $L_1$  to grant a higher degree of freedom in the design [72].



## Load-Independent Class EF

By formulating the state-space equations of the Class EF inverter, solutions can be found so that the drain voltage converges to zero at turn-on for a broad range of load resistances. Interestingly, the solution for load-independent operation in the Class EF inverter [73], which is similar to that found for a Class E inverter in [64] achieves the load-independent criteria but in addition produces a constant amplitude output current as opposed to the constant amplitude output voltage achieved with the Class E. This characteristic makes the load-independent Class EF inverter more suitable to drive an IPT-coil, without requiring an impedance matching network.

Load independent operation is achieved, unlike in the tuning variations where the LC branch is tuned at the second or third harmonic ( $EF_2$ ,  $EF_3$ ), by tuning the  $L_2 C_2$  branch not at a particular harmonic of the switching frequency, but between the first and second harmonic. Infinite mathematical solutions that comply with this condition can be found by solving the state space equations of this topology; however, not all of these solutions are practical. Practical solutions are usually found by optimising the duty cycle for maximum power output capability of the transistor [73, 69, 70], which unlike the ZVS Class E, is at 30 %. Fig. 3.8 shows the theoretical waveforms of a particular solution for load-

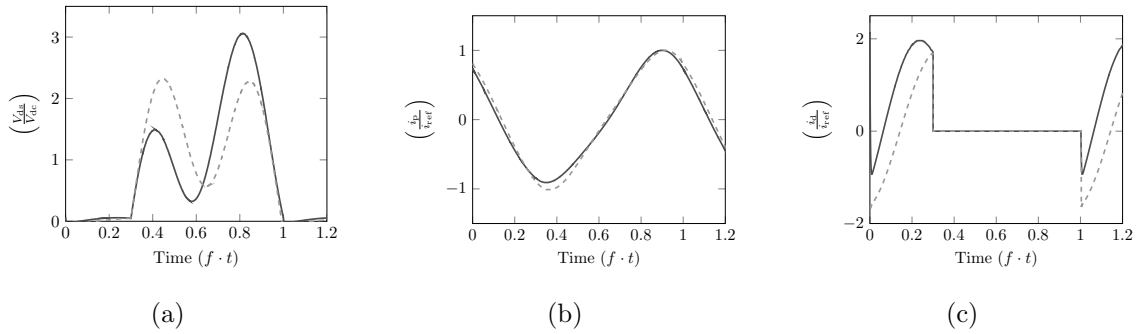


Figure 3.8: Theoretical waveforms of a load independent Class EF inverter at full load. The theoretical waveforms of the inverter when unloaded are plotted with a dashed line.

independent operation. The load independent feature can be identified by observing that the drain voltage waveform converges to zero at  $f \cdot t = 1$  independent of the load, demonstrating that indeed ZVS is achieved for the entire load range, and that the anti-

parallel diode of the transistor never conducts. Additionally, the amplitude and phase of  $i_p$  (Fig. 3.8) do not change, significantly, with load resistance.

The amplitude of  $i_p$  is proportional to  $V_{dc}$  and inversely proportional to the reactance of  $C_1$  ( $X_{C_1}$ ). This feature allows performing high level control on a HF-IPT system that uses this inverter by changing the input voltage. Nonetheless, since the average drain voltage of the transistor is equal to  $V_{dc}$ , by changing its value, the effective capacitance  $C_1$  changes due to the voltage-dependence of the non-linear  $C_{oss}$ . Thus, large variations in  $V_{dc}$  can cause the inverter to become detuned in some designs.

### 3.2.3 Push-Pull Class E and Class EF Inverters

The Class E and Class EF topologies can be constructed using a push-pull variation (Fig. 3.9) [74, 75]. The principle consists of operating the two branches of the inverter with a phase-shift of  $180^\circ$ . This allows for a higher output power, and lower harmonic distortion.

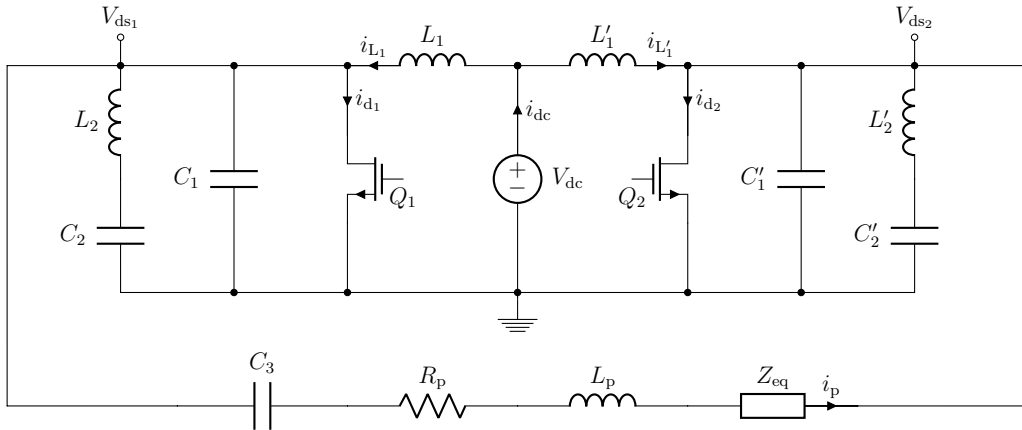


Figure 3.9: Push-Pull Class EF inverter circuit topology.

The main challenges of implementing this type of system is maintaining the symmetry between the two branches once the system starts to operate. Changes in the component values, due to temperature, for example, or discrepancies between the switching signals (the phase shift not being exactly  $180^\circ$ , or differences in duty cycle) can overload one of

the branches, and severely affect the overall performance of the power converter. The design equations of single-ended topologies can be utilised considering that the transmit coil and  $\mathbf{Z}_{\text{eq}}$  are fed between the two branches of the inverter.

### 3.2.4 Power Throughput Control of Class E and Class EF Inverters

Traditional pulse-width modulation (PWM) power converters use the duty cycle as the control variable from which the output voltage, and input and output current can be shaped for specific purposes: output voltage regulation to feed electronic equipment, output current regulation to feed current driven loads such as light emitting diodes, and input current shaping in applications requiring active power factor correction (PFC) circuits.

Controlling such variables in Class E or Class EF inverters is more challenging since the advantageous switching characteristics of this family of inverters depend on frequency and duty cycle [76], and thus, switching frequency and duty cycle are not preferred candidate variables to perform power throughput control.

Most systems that use this type of inverter therefore operate the switch in open loop and control the power throughput with an additional power converter at the input, which varies the voltage fed to the inverter and consequently  $i_p$ . Alternatives such as on-off control on the switching device can also be integrated but with a limited control capability [77]. Other alternative solutions are possible. For example, in [68, 78], a saturable reactor was included in the resonant output network to achieve variable inductance using an external source of DC current. Such solutions, despite being interesting, are not practical due to the excessive losses generated and the limitations on the range of control.

With respect to the dynamic response of the inverters, from the reactive components, only the input filter, which includes  $L_1$ , has to be considered [79], since it is the only reactive component of the inverter that stores a sufficient amount of energy not to

be discarded in such analysis. It should be noted, however, that the representation of the reflected impedance on the transmit side may need to be modelled in such way that it represents the components that store energy at the receive side depending on their values.

The inverters proposed in this thesis operate with constant frequency and duty cycle and changes in the output power from the transmit side are performed by changing the input voltage of the inverter, since changes to  $\mathbf{Z}_{\text{eq}}$ , which is the other variable that determines output power, are produced by control on the receive side.

### 3.3 Rectifiers for HF-IPT

At the receive-side of a HF-IPT system (Fig. 3.1), the circuit in charge of transforming the  $\mathcal{E}_s$  into a DC voltage is the IPT rectifier. Fig. 3.10 shows an equivalent circuit where the  $\mathcal{E}_s$  is the source of power and the input impedance of the IPT-rectifier ( $\mathbf{Z}_{\text{sec}}$ ) is the load.

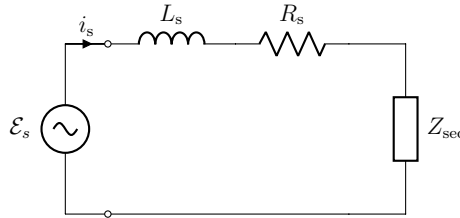


Figure 3.10: Equivalent circuit of the receive side of a HF-IPT system with  $\mathcal{E}_s$  as the source of power and  $\mathbf{Z}_{\text{sec}}$  as the load.

IPT rectifiers can be synchronous (uses controlled switches), or non-synchronous (uses diodes). Synchronous rectifiers allow for a higher degree of control and can be more efficient due to the lower conduction losses that can be achieved with transistors as opposed to diodes. Nonetheless, employing them for HF-IPT is not only more complicated due to the additional circuitry, but difficult to implement [80, 81], as triggering the signals that drive the switch or switches correctly and tuning is most challenging, practically at megahertz.

Similar to resonant inverters, rectifiers are divided into two main categories, Class D and Class E, and the circuit topologies are equal to those of the inverters. Their main characteristics are also similar to those of the inverters. A basic comparison of non-synchronous Class D and Class E rectifiers for HF-IPT can be found in [82].

The work presented in this section is concerned exclusively with non-synchronous rectifiers, and the only end-to-end application example with a synchronous rectifier in the thesis can be found in Section 7.2.12, published in [83, 84], where an alternative solution to power up a drone without a battery with a synchronous Class E rectifier is mentioned.

### 3.3.1 Current-Driven Class D Rectifiers for HF-IPT

Current-driven Class D rectifiers (Fig. 3.11) can be constructed using a half or full diode bridge (transistors can be used instead of diodes in synchronous solutions), respectively, for half and full wave rectification. Each solution has different properties and equations [63], and both can be considered for HF-IPT.

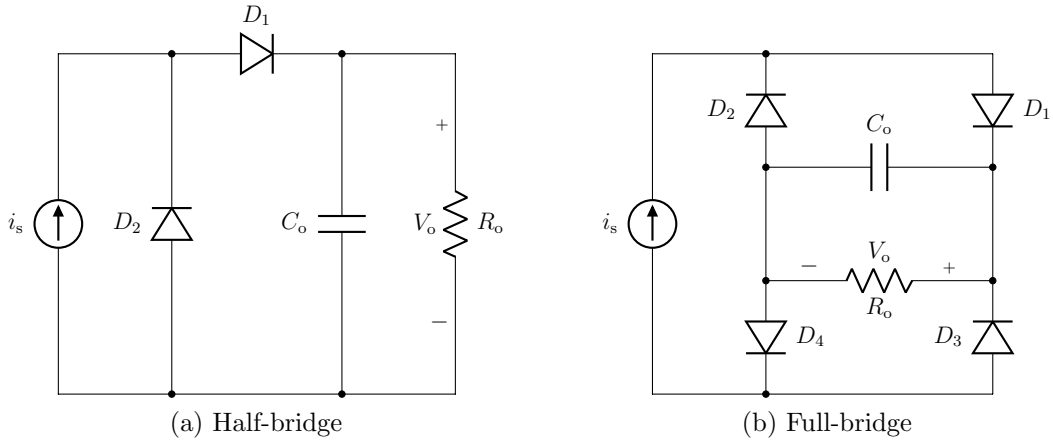


Figure 3.11: Current-driven Class D rectifier circuit topologies.

Since in HF-IPT systems, the value of  $\mathbf{Z}_{\text{sec}}$  is essential to maximise the link efficiency (2.15), the equation that describes the input impedance of the rectifier with respect to the DC load is fundamental for system design. A first order approximation of  $\mathbf{Z}_{\text{sec}}$  is given

by [63] as:

$$\mathbf{Z}_{\text{sec}} \Big|_{\text{HB-Class D}} = \frac{2R_o}{\pi^2 \eta_{\text{rect}}}, \quad (3.1)$$

for the half-bridge, and:

$$\mathbf{Z}_{\text{sec}} \Big|_{\text{FB-Class D}} = \frac{8R_o}{\pi^2 \eta_{\text{rect}}}, \quad (3.2)$$

for the full bridge, expressed in terms of the IPT system DC load ( $R_o$ ) and the IPT rectifier efficiency ( $\eta_{\text{rect}}$ ). According to this model, the input impedance of Class D rectifiers is resistive; however, the parasitic junction capacitance of the diodes can make a Class D rectifier behave like a Class DE rectifier [85], which has an input reactance different than zero.

Class D rectifiers, when integrated in an IPT environment, tend to behave differently than the theoretical model of [63] since the resonant circuit at the receive side, when the load or coupling changes, is no longer well represented as an ideal current source. Therefore, Class D rectifiers for HF-IPT can be more accurately described using the circuits diagrams in Fig. 3.12, where the receive-coil current ( $i_s$ ) depends on  $R_o$ , and consequently  $\mathbf{Z}_{\text{sec}}$ .

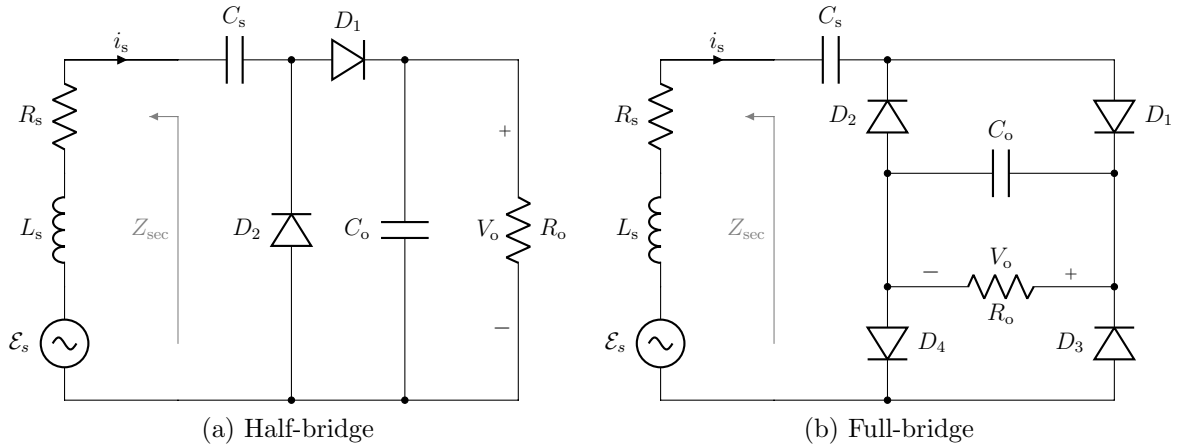


Figure 3.12: Current-driven Class D rectifier circuit topologies in an IPT receiver.

Voltage-driven Class D rectifiers can also be implemented in HF-IPT by parallel-tuning

the receive coil, incorporating an output inductive filter, and using  $D_2$  in Fig. 3.11 (a) as a freewheeling diode to the inductive filter, not a freewheeling diode for the source. Since this topology requires a large inductance in the output LC filter, it was not selected in the work presented in the following chapters of this thesis.

### 3.3.2 Class E Rectifiers for HF-IPT

The Class E rectifier operates with the same principle as the Class E inverter, but with opposite power flow. An AC current or voltage source (there are voltage-driven and current-driven variants) is rectified using a diode (or transistor if synchronous) which with the aid of a resonant tank performs soft-switching. Conveniently, the topologies include a capacitance in parallel with the diode, allowing the design of the passives so that the junction capacitance of the diode is absorbed. Interestingly, however, the junction capacitance losses, which are not often considered, or detailed in the data sheets, can significantly affect the performance of a rectifier [86].

The circuit diagrams for a theoretical current-driven Class E rectifier, and a hybrid Class E rectifier in an IPT receiver [87] are shown in Fig. 3.13.

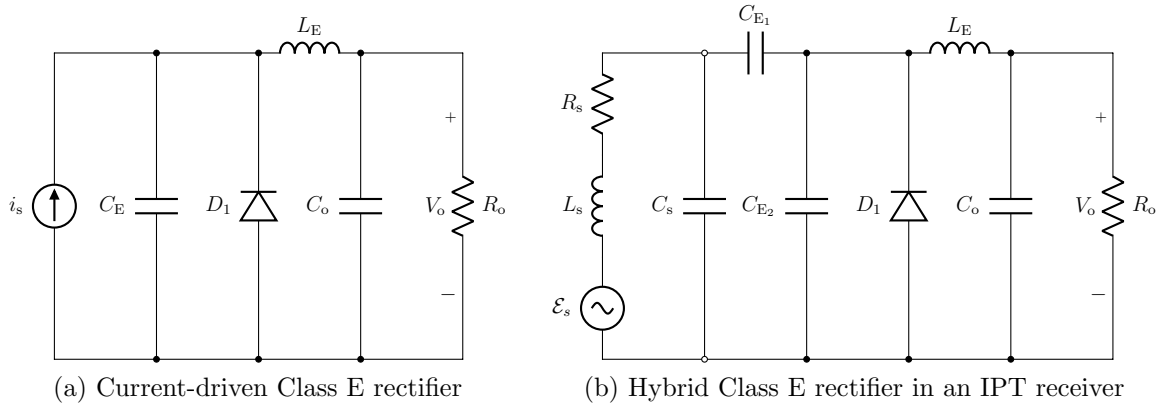


Figure 3.13: Selected Class E rectifier circuit topologies.

The duty cycle of Class E rectifiers changes inherently with load, and negative conduction of the device (see this for Class E inverters in Fig.3.5) does not occur if the diode is ideal. As in Class E inverters, the voltage rating of the device is required to be more than three

times the DC voltage ( $> 3V_o$ ) which narrows to an extent the candidate diode technologies for mid-power applications. Considering the range of power of the aimed applications of this thesis, all the Class E solutions are proposed with silicon carbide (SiC) Schottky diodes.

Very detailed design guidelines for non-synchronous Class E rectifiers for HF-IPT can be found in [87, 88]. In that work, the  $\mathbf{Z}_{\text{sec}}$ , and hence  $\mathbf{Z}_{\text{eq}}$ , (Fig. 2.3), are characterised in terms of  $R_o$ , and design guidelines are given to achieve low reflected reactance on the transmit coil ( $X_{\text{eq}}$ ) for a broad range of  $R_o$ . This feature is essential to avoid detuning (losing resonance at the receive side) a HF-IPT system when required to operate across a large range of loads.

### 3.3.3 Push-Pull Class E Rectifiers for HF-IPT

Similar to Class E or Class EF inverters, Class E rectifiers can be constructed in push-pull or full-bridge configurations. Fig. 3.14 shows the circuit topology of a push-pull hybrid Class E rectifier, which can be designed for low reactance deviation, i.e. low  $X_{\text{eq}}$ .

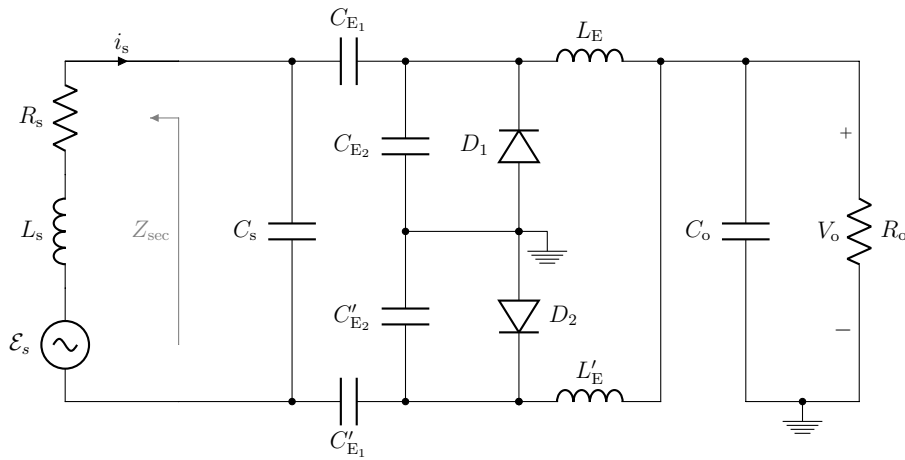


Figure 3.14: Circuit topology of a push-pull hybrid Class E rectifier.



### 3.3.4 Resistance Compression Networks on HF-IPT Rectifiers

Employing a RCN on a rectifier was first done in [89, 90] to minimise changes in the input resistance of a rectifier when  $R_o$  changes. This principle was afterwards implemented in RF transmission lines for radiative wireless power transfer (WPT) systems [91], and was introduced for HF-IPT in the author's published works [54, 53].

In principle, RCNs allows  $R_o$  to be simultaneously fed from two rectifiers, each containing an additional reactive impedance connected either in series or in parallel, depending if the rectifier is current-driven or voltage-driven (Fig. 3.15). Each rectifier is modelled as an

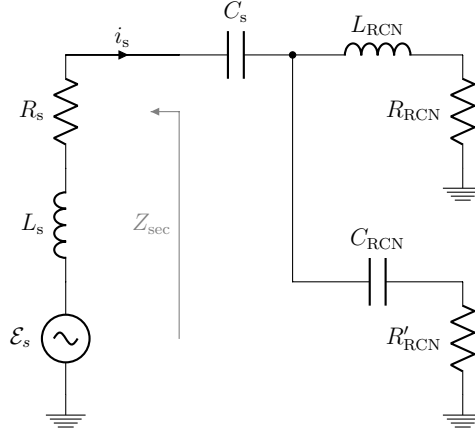


Figure 3.15: Resistance compression network conceptual circuit.

equal resistance ( $R_{RCN} = R'_{RCN}$ ), and the two supplemental reactive impedances ( $L_{RCN}$  and  $C_{RCN}$ ) should be designed as complex-conjugates at the frequency of operation (the transmit-coil current frequency ( $f_{ip}$ )), i.e. with an equal reactance magnitude:

$$X_{RCN} = 2\pi f_{ip} L_{RCN} = \frac{1}{2\pi f_{ip} C_{RCN}}, \quad (3.3)$$

thus that the self-inductance of the receive coil ( $L_s$ ) can be tuned out with the resonant capacitance at the receiver ( $C_s$ ), independent of  $R_o$  at  $f_{ip}$ :

$$Z_{sec} \Big|_{f_{ip}} = \frac{L_{RCN}}{2R_{RCN}C_{RCN}} \left( 1 + \frac{C_{RCN}R_{RCN}^2}{L_{RCN}} \right) - j \left( \frac{1}{2\pi f_{ip} C_s} \right). \quad (3.4)$$

This impedance transformation creates a compressed dependency between the rectifier's

input resistance ( $R_{\text{sec}} = \text{Re}(Z_{\text{sec}})$ ) and  $R_{\text{RCN}}$ , which should be clear from Fig. 3.17, and hence between  $R_{\text{sec}}$  and  $R_o$ .  $R_{\text{sec}}$  also depends of the magnitude of the reactance of the supplemental reactive impedances ( $X_{\text{RCN}}$ ), which allows the magnetic link to be operated with a close to optimal impedance (2.15) across a wide range of  $R_o$ .  $X_{\text{RCN}}$  can be treated as an independent design variable.

The concept of RCNs can be extrapolated to higher orders to achieve better compression, as proposed for transmission-line RCN rectifiers in [91], where four rectifiers were connected to the source via two stages of reactive impedances. This approach, however, is not practical for HF-IPT because of the additional losses and the unpractical challenges of measuring and tuning for such high order networks.

### 3.3.5 Current-Driven Class D Rectifiers with Resistance Compression Networks for HF-IPT

The principle of RCNs can be employed on HF-IPT systems with two half-wave Class D rectifiers, a capacitance ( $C_{\text{RCN}}$ ) and an inductance ( $L_{\text{RCN}}$ ) (Fig. 3.16). This proposed

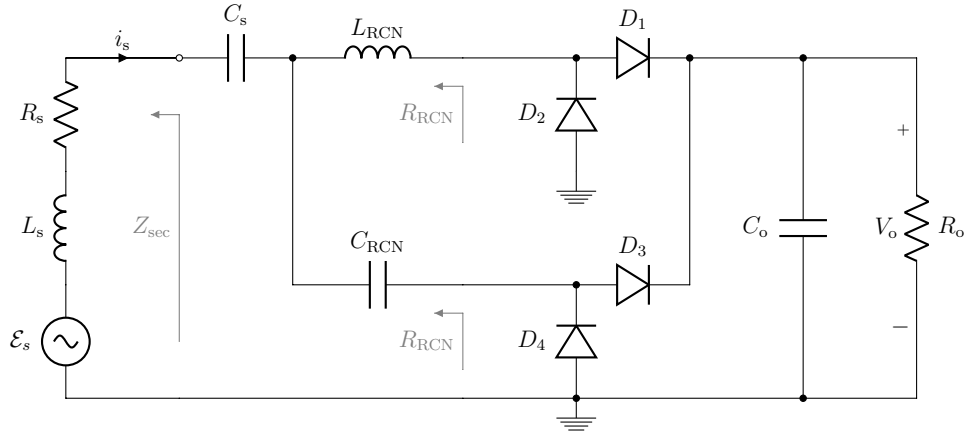


Figure 3.16: Class D rectifier with a resistance compression network.

topology uses a first order RCN and two current-driven Class D rectifiers, which since they both feed the same load  $R_o$  simultaneously, the equivalent load for each rectifier is  $2R_o$ . Therefore, the input impedance of each rectifier in terms of  $R_o$  and in accordance

to (3.1) is:

$$R_{\text{RCN}} = \frac{4R_o}{\pi^2 \eta_{\text{rect}}}. \quad (3.5)$$

Thus,  $R_{\text{sec}}$  can be calculated as proposed in (3.4), where  $R_{\text{RCN}}$  is determined by (3.5). There are, however, limitations on the accuracy of this expressions since when  $R_{\text{RCN}}$  is greater than  $X_{\text{RCN}}$ , the Class D rectifiers enter a discontinuous mode of operation, i.e. the diodes do not always conduct, where the cause of such phenomenon is that at low loaded Q factors, the sinusoidal current waveform at the input of the Class D rectifiers deteriorates, which is something that a first order model does not account for.

Fig. 3.17 shows a plot of (3.4) with an  $R_{\text{RCN}}$  determined by (3.5) against a SPICE simulation, with ideal components, demonstrating how the input resistance of this proposed

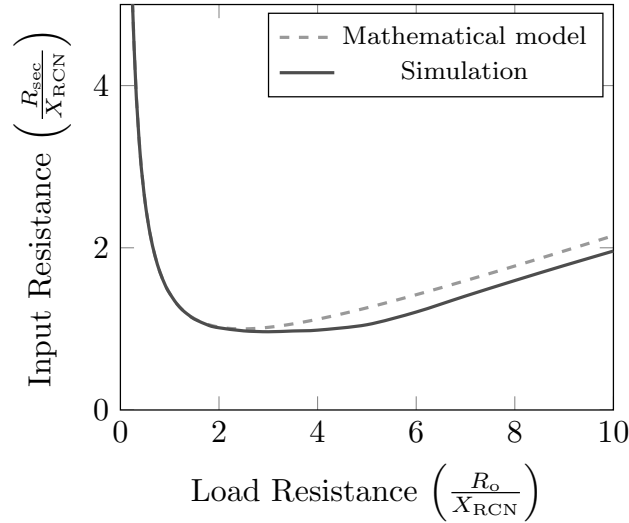


Figure 3.17: Input resistance of the RCN-Class D rectifier.

topology has a compressed dependence on  $R_o$  and how it also depends of the design variable  $X_{\text{RCN}}$ . Therefore, the RCN Class D rectifier can be designed to inherently present the optimal load to the IPT system for a broad range of loads.

### 3.3.6 Multiplier Class D Configurations for HF-IPT

As demonstrated with the RCN Class D rectifiers, more than one Class D rectifier can be used to simultaneously feed a DC load,  $R_o$ . These, while not necessarily needed to feed the load in parallel, can also be stacked in series (Fig. 3.18). This concept, introduced as

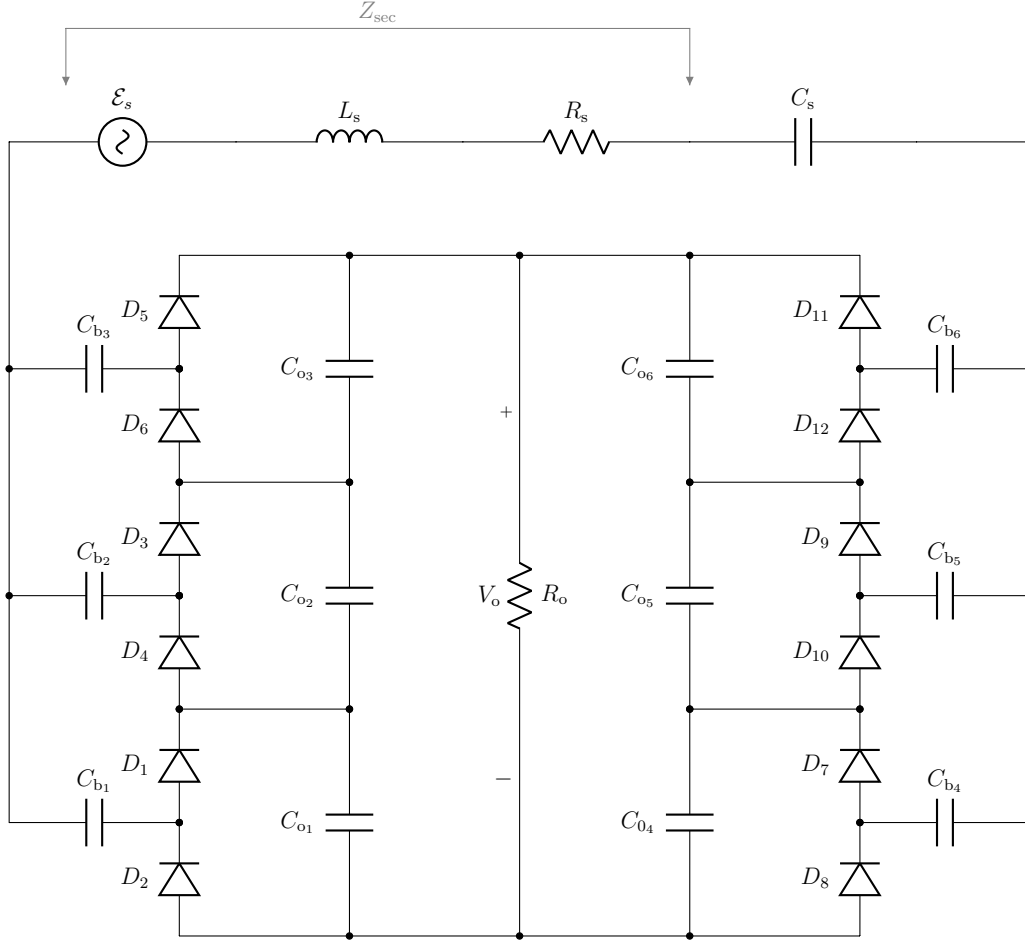


Figure 3.18: A 3x voltage-multiplier full-wave Class D rectifier for high output voltage HF-IPT.

the Cockcroft-Walton Voltage Multiplier [92] can comprise an arbitrary number of stages in order to multiply the voltage (the IPT-rectifier output voltage ( $V_o$ )) by a factor equal to the number of stages.

When a single DC load is fed with multiple rectifiers, the equivalent load fed by each rectifier, is a fraction of  $R_o$ , and also, with respect to the source, all the rectifiers are fed in parallel. Therefore the voltage multiplier principle can be used as an impedance transformation from  $R_{sec}$  to  $R_o$ , which depends of how the individual rectifiers are placed

in the topology.

Fig. 3.18 shows a rectifier design example of this principle where a voltage multiplier rectifier, consisting of six individual Class D rectifiers (or three full bridge rectifiers) is implemented. Here, each full bridge rectifier feeds a load equal to  $R_o/3$ , in parallel, and therefore:

$$\mathbf{Z}_{\text{sec}} \Big|_{f_{\text{ip}}} = \frac{8R_o}{9\pi^2\eta_{\text{rect}}} - j \left( \frac{1}{2\pi f_{\text{ip}} C_s} \right). \quad (3.6)$$

This candidate topology is utilised for high output voltage systems required in the following chapters of this thesis.

### 3.4 Inherent Regulation in Resonant Converters for Variable Loads

The inherent regulation properties of resonant converters can sometimes be used to maintain soft switching, for instance, non-synchronous Class E rectifiers inherently change the duty cycle of the diode when the dc-load changes, and consequently achieve ZVS. Furthermore, in [87] the hybrid Class E rectifier not only achieves ZVS, but also can be designed to inherently regulate the output voltage as the load changes. The inherent regulation properties of resonant converters have been utilised for multiple purposes, for example in [93] the resonant nature of a LCC converter was utilised for power factor correction, and in [94] an inherent over-current protection feature was implemented by placing an auxiliary circuit in parallel with the resonant capacitor of a LLC resonant converter which shorts the capacitor and detunes the resonant tank when a fault shorts out the load.

The mentioned converters exploit some of the inherent regulation properties of resonant converters to allow for changes in load and on the geometry of the link.

## 3.5 Design of the Circuits for Experimental Verification

In this work, a set of candidate circuits were designed to perform experiments concerning system optimisation methods and technology demonstrators that are proposed in the following chapters of the thesis. The candidate circuits were initially tuned for ideal transmit and receive coils, as described and characterised in Chapter 4. The candidate designs are listed in Table 3.1. In addition to these, additional converters were designed

Table 3.1: Candidate inverters and rectifiers for experimental verification

| Name         | Topology and tuning                 | Frequency [MHz] | Power [W] |
|--------------|-------------------------------------|-----------------|-----------|
| Inverter I   | Load-tolerant Class E               | 13.56           | 150       |
| Inverter II  | Load-independent Class EF           | 13.56           | 150       |
| Inverter III | Push-pull load-independent Class EF | 13.56           | 350       |
| Rectifier I  | RCN Class D                         | 13.56           | 150       |
| Rectifier II | Voltage multiplier Class D          | 13.56           | 350       |

and tuned for other coils in Chapter 7, where the shape and construction divert from the ones in Chapter 4 due to the requirements and size restrictions of the application.

### 3.5.1 Inverter I: Load Tolerant Class E at 13.56 MHz for 150 W

A load-tolerant Class E inverter (Section 3.2.1) was designed and tuned at 13.56 MHz with the component values specified in Table 3.2, and a duty cycle of 50 %, for maximum power output capability. The inverter was designed to drive the transmit coil proposed in Chapter 4 with a reflected resistance of  $0\ \Omega$  to  $10\ \Omega$ . The switching device for the proposed inverter is a gallium nitride (GaN) FET with a voltage blocking capability of 650 V and a maximum drain current of 15 A (GS66504B by GaN Systems). The switching signal was provided by an on-board signal generator, and amplified by a high-speed power-FET driver by Intersil (ISL55110).

Table 3.2: Component values of Inverter I: Load-tolerant Class E

| Topology | $L_1$ [ $\mu\text{H}$ ] | $L_2$ [ $\text{nH}$ ] | $C_1$ [ $\text{pF}$ ] | $C_2$ [ $\text{pF}$ ] | $C_3$ [ $\text{pF}$ ] |
|----------|-------------------------|-----------------------|-----------------------|-----------------------|-----------------------|
| Class E  | 88                      | -                     | $81 + C_{\text{oss}}$ | -                     | 163                   |

### 3.5.2 Inverter II: Load Independent Class EF at 13.56 MHz for 150 W

A load-independent Class EF inverter (Section 3.2.2) was designed and tuned at 13.56 MHz with a duty cycle of 30 %, for maximum power output capability. The design equations, obtained from [73], are as follows: the resonant frequency of the  $C_2$ - $L_2$  branch is:

$$\frac{1}{2\pi\sqrt{L_2 C_2}} = 1.67 f_{i_p}, \quad (3.7)$$

where a set of practical load-independent design equations are derived considering that the inverter was designed to drive the transmit coil proposed in Chapter 4. The switching device for the proposed inverter is a GaN FET with a voltage blocking capability of 650 V and a maximum drain current of 15 A (GS66504B by GaN Systems). The switching signal was provided by an on-board signal generator, and amplified by a high-speed power-FET driver by Intersil (ISL55110). The other design equations from [73] are:

$$\begin{aligned} 2\pi f_{i_p} R_{\text{eqmax}} C_1 &= 0.118614, \\ \frac{C_1}{C_2} &= 1.329341, \\ \frac{|i_p| R_{\text{eqmax}}}{V_{\text{dc}}} &= 0.286373, \\ \frac{X_{\text{res}}}{R_{\text{eqmax}}} &= 0.342106, \end{aligned} \quad (3.8)$$

where  $X_{\text{res}}$  is residual reactance:

$$X_{\text{res}} = 2\pi f_{i_p} L_p - \frac{1}{2\pi f_{i_p} C_3}. \quad (3.9)$$

These design equations were found for  $R_{\text{eqmax}} = 6\Omega$ , resulting in the component values listed in Table. 3.3.

Table 3.3: Component values of Inverter II: Load-independent Class EF

| Topology | $L_1$ [ $\mu\text{H}$ ] | $L_2$ [ $\text{nH}$ ] | $C_1$ [ $\text{pF}$ ]  | $C_2$ [ $\text{pF}$ ] | $C_3$ [ $\text{pF}$ ] |
|----------|-------------------------|-----------------------|------------------------|-----------------------|-----------------------|
| Class EF | 88                      | 267                   | $156 + C_{\text{oss}}$ | 176                   | 130                   |

### 3.5.3 Inverter III: Push-Pull Load Independent Class EF at 13.56 MHz for 350 W

A push-pull Class EF inverter (Section 3.2.3) was designed using the design equations proposed for the Inverter II: (3.7), (3.8), but since the self-inductance of the transmit coil ( $L_p$ ) and  $\mathbf{Z}_{\text{eq}}$  are distributed between the two branches of the inverter, the component values differ from the previous design, as shown in Table 3.4. The inverter was designed to drive the transmit coil proposed in Chapter 4. The switching device for the proposed inverter is a GaN FET with a voltage blocking capability of 650 V and a maximum drain current of 30 A (GS66508B by GaN Systems). The switching signal was provided by an on-board signal generator, and amplified by a single channel high-speed, low-side gate driver by Texas Instruments (UCC27512-EP).

Table 3.4: Component values of Inverter III: Push-pull load-independent Class EF

| Topology           | $L_1$ & $L'_1$ [ $\mu\text{H}$ ] | $L_2$ & $L'_2$ [ $\text{nH}$ ] | $C_1$ & $C'_1$ [ $\text{pF}$ ] | $C_2$ & $C'_2$ [ $\text{pF}$ ] | $C_3$ [ $\text{pF}$ ] |
|--------------------|----------------------------------|--------------------------------|--------------------------------|--------------------------------|-----------------------|
| Push-pull Class EF | 88                               | 66                             | $270 + C_{\text{oss}}$         | 470                            | 128                   |

### 3.5.4 Rectifier I: RCN Class D at 13.56 MHz for 150 W

A RCN current-driven Class D rectifier (Section 3.3.5) was designed and tuned at 13.56 MHz, with  $C_s$  designed to tune out the self-inductance of the receive coil, which is specified in Chapter 4. The RCN network was designed by selecting the magnitude of the reactance



of the reactive components ( $C_{\text{RCN}}$ ,  $L_{\text{RCN}}$ ):  $X_{\text{RCN}}$ , to be slightly lower than the optimal load of the link (at a coupling corresponding to a coil separation of 8 cm) in order to achieve the optimal load with an  $R_o$  slightly higher or lower than  $\frac{\pi^2}{4}X_{\text{rcn}}$ , which is when the minimum value of  $R_{\text{sec}}$  (equal to  $X_{\text{rcn}}$ ) is achieved.

It is worth noticing that according to (3.4) and the plots in Fig. 3.17,  $R_{\text{sec}}$  can be equal to the optimal load at two values of  $R_o$ , however the greater value of the two, achieves a higher efficiency given that the current in the diodes is lower at higher values of  $R_o$ .

The component values selected are specified in Table 3.5. The diodes used are 120 V and 10 A, Schottky diodes by ON Semiconductor (FSV10120V).

Table 3.5: Component values of Rectifier I: RCN Class D

| Topology    | $L_{\text{RCN}}$ [nH] | $C_s$ [pF] | $C_{\text{RCN}}$ [nF] | $C_o$ [μF] |
|-------------|-----------------------|------------|-----------------------|------------|
| RCN Class D | 42                    | 112        | 3.11                  | 1          |

### 3.5.5 Rectifier II: 3x Voltage-Multiplier Full-Wave Class D at 13.56 MHz for 350 W

A three-stage voltage-multiplier full-wave Class D rectifier (Section 3.3.6, Fig. 3.18) was designed for high output voltage systems.  $C_s$  was designed to resonate with the receive coil in Chapter 4 and  $C_b$  is a DC-blocking capacitance to allow each current-driven Class D rectifier to operate at different voltage offsets. The output capacitance of the two current-driven Class D rectifiers are connected in series to achieve a higher output voltage than a full-wave current-driven Class D rectifier.

The component values designed are presented in Table. 3.6, and the diodes used for this prototype are 100 V, 8 A Schottky diodes by STMicroelectronics (STPS8H100).

Table 3.6: Component values of Rectifier II: 3x voltage-multiplier full-wave Class D

| Topology                                | $C_s$ [pF] | $C_{b1-6}$ [nF] | $C_{o1-6}$ [ $\mu$ F] |
|-----------------------------------------|------------|-----------------|-----------------------|
| 3x voltage-multiplier full-wave Class D | 112        | 0.1             | 1                     |

## 3.6 Construction of Circuit Prototypes for Experimental Verification

The above proposed inverters and rectifiers were designed and built for the experiments of the following chapters of the thesis. Fig. 3.19 shows photographs of the inverters and

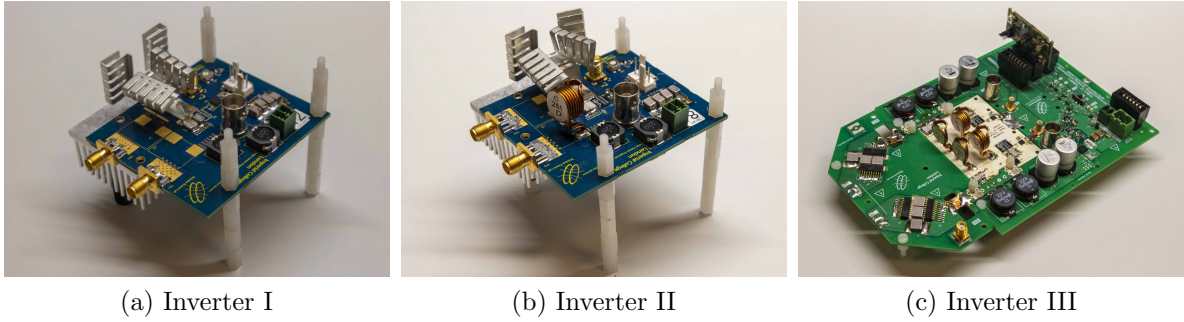


Figure 3.19: Photographs of the inverter prototypes.

Fig. 3.20 shows photographs of the rectifiers.

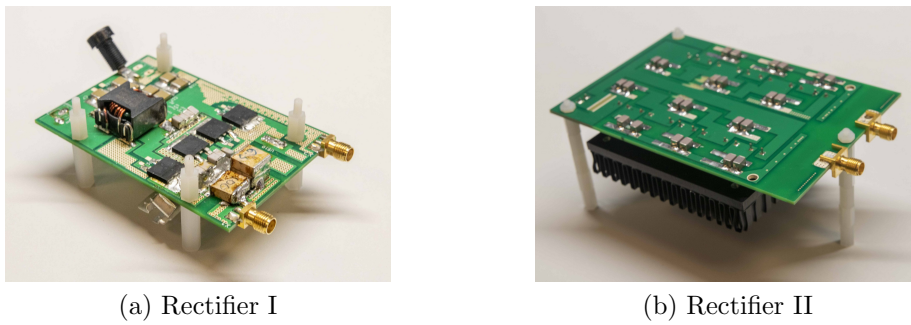


Figure 3.20: Photographs of the rectifier prototypes.

### 3.6.1 General Considerations in the Components Selection

The following considerations were taken in the design and construction of the circuits:

- The coil can be connected to the inverter or the rectifier using two shorted edge-mount SubMiniature version A (SMA) connectors (one at each terminal of the coil), visible at the front of the inverters in Fig. 3.19, and in the rectifiers in Fig. 3.20.
- The capacitors of the resonant tanks use the C0G (NP0) dielectric. Specifically the selected models were QUAD HIFREQ Series from Vishay and the C Series by TDK.
- The cored inductors used are from the WE-PD SMT Power Inductor series by Würth Elektronik. In addition, inductors with in-house winding like the one in Fig. 3.20(b) use Powder Core Solutions by Micrometals.
- The air-core inductors used are from the 2014VS series by Coilcraft.

### 3.6.2 General Considerations in the Design of the Printed Circuit Boards

As discussed in the following chapters, the main tuning of the system is done at the transmit side, and thus an extended set of considerations were taken in the design of these boards:

- The inverters were build with two separate input voltage terminals, one for the inverter's input voltage and one for the gate driver's voltage to allow for changes on the gate voltage.
- The inverter boards have an on-board signal generator at 13.56 MHz and an RF connector to allow driving it with an external signal if required.
- The inverter boards include a voltage test point at the drain of the transistors.

Inverter I and Inverter II have the same board layout Fig. 3.19 (a) and Fig. 3.19 (b), allowing these two configurations to be easily compared. Inverter III uses a modular board

design, including an Insulated Metal Substrate (IMS) board in the switching networks for better thermal conductivity. The inverters printed circuit boards (PCBs) in Fig. 3.19, were designed by Dr. Samer Aldhafer (Inverters I, II and III), and the author designed the rectifier PCBs in Fig 3.20 (Rectifiers I and II).

### 3.7 Construction of a Setup for Experimental Verification of HF-IPT

The experimental work presented in this thesis was performed at the Wireless Power Lab of Imperial College London. Fig. 3.21 shows a typical experimental setup for the WPT experiments presented.

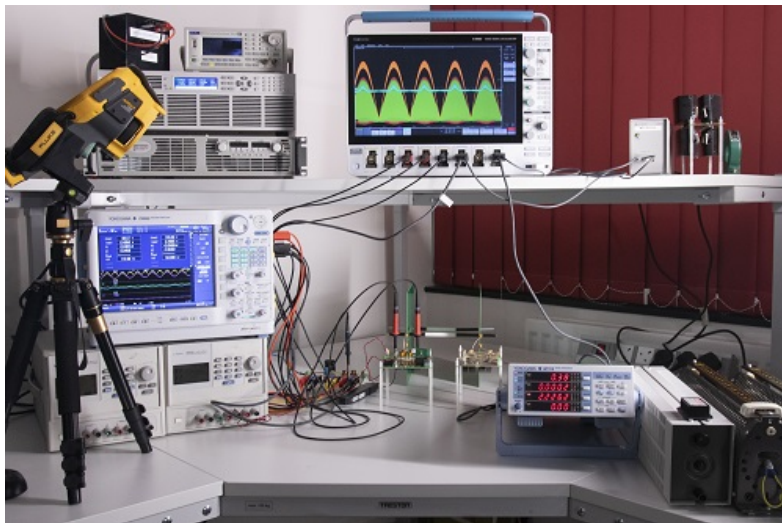


Figure 3.21: Photographs of the experimental setup for HF-IPT.

#### 3.7.1 Laboratory Equipment

The main laboratory equipment used in the experiments presented in this thesis is listed as following:

- Power supplies: Two main power supplies were used, a N8741A System DC Power Supply by Agilent Technologies (0 V to 300 V, 0 A to 11 A), and an AC6802A

AC Power Supply by Keysight (0 V to 270 V, 0 A to 5 A, ratings specified in root-mean-square (RMS), with a maximum frequency of 500 Hz). In addition, the voltage supply used to power the gate drivers of the inverters is an Agilent U8032A.

- Signal generators: In cases where the input signals of the gate drivers in the inverter boards were not generated internally, a 50 MHz Function Generator by Advanced Test and Measurement Instruments and Power Supplies was utilised.
- Power Meters: A WT332E Digital Power Meter by Yokogawa was used to measure system input power and a WT310 Digital Power Meter by Yokogawa was used to measure system output power. In some experiments, a WT210 Digital Power Meter by Yokogawa was also utilised, and in Chapter 5 some experiments used a PX8000 Precision Power Scope by Yokogawa.
- Oscilloscopes and probes: Most of the waveforms presented in this thesis were captured using an 8-Channel Tektronix 5 Series MSO, however a Teledyne LeCroy HDO4104 and a Teledyne LeCroy 4054MS were also utilised. Other than the passive voltage probes that come with the oscilloscopes, high voltage probes (PHV 1000 100:1 by PMK and 1000:1 PPR 6 kV by LeCroy) were used to measure the drain voltage of the inverters. The system input and output current was measured using Keysight N2783B current probes, and the coil current was measured using a 6585 Current Monitor by Pearson.
- Impedance analyser: The impedance measurements presented in this thesis were done using a Keysight E4990A impedance analyser with a 42941A impedance probe.
- Thermal camera: Thermal images presented in this thesis were captured with a Fluke Ti400 Thermal Imager.

### 3.7.2 Laboratory Safety

The experimental work presented in this thesis followed a rigorous safety protocol including the general safety rules from Imperial College London and the specific rules from

the Wireless Power Lab. Risk assessments were generated for every test-rig and a Narda safety system was used to ensure not exceeding relevant electromagnetic fields exposure limits. Any experiment involving voltages over 50 V was conducted within a protective insulating box with minimal openings allowed for probing.

### 3.7.3 Auxiliary Printed Circuit Boards

A set of auxiliary PCBs were fabricated for the development of the experiments in this thesis. A circuit board with a diode bridge (for the AC input voltage experiments in Chapter 5), a common-mode filter and the test-cable connections necessary to measure power at the input of the HF-IPT system was constructed as an interface between the power supplies and the inverters (Fig. 3.22). Other boards (e.g. the rectifier evaluation

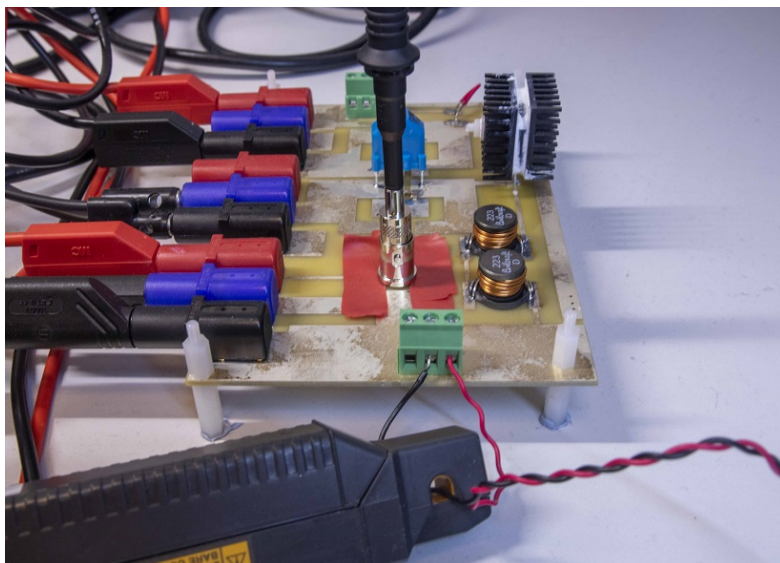
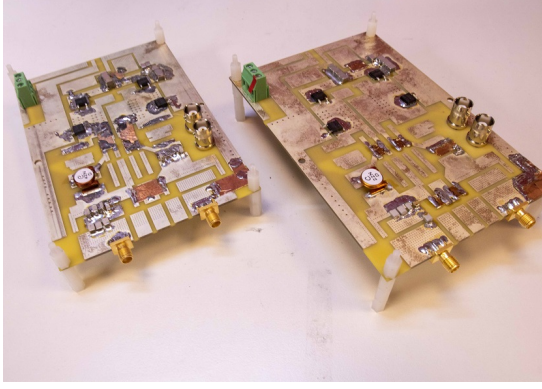
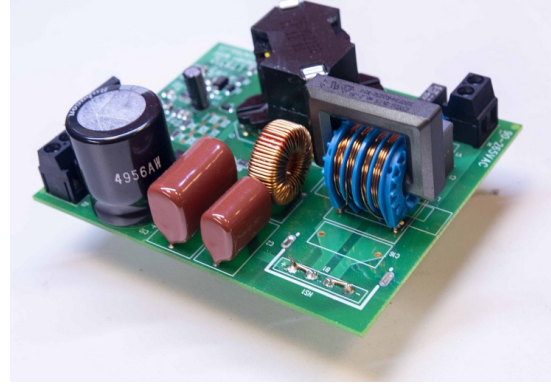


Figure 3.22: Photograph of the input power measuring board.

board in Fig. 3.23 (a)) were constructed for the development of intermediate designs, and a number of evaluations boards (e.g. the PFC stage in Fig. 3.23 (b)) for power factor correction and voltage regulation. Throughout the thesis, when utilised, these are specified in the respective sections.



(a) Rectifiers

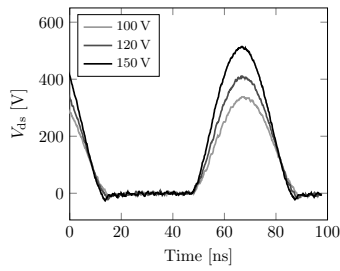


(b) PFC evaluation board by Linear Technology

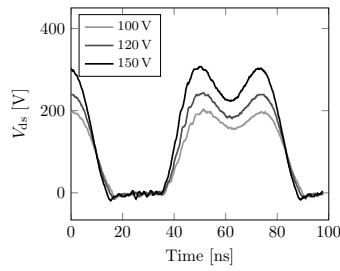
Figure 3.23: Photographs of additional boards used in the experiments.

### 3.7.4 Measurements of the Proposed Prototype Inverters Without a Coupled Receiver

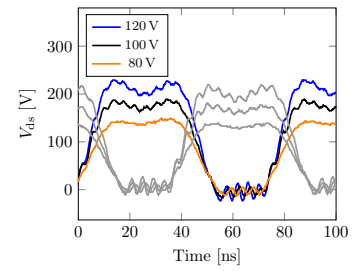
Once built, the prototype inverters were fed from a DC power supply at three input voltages, as an initial test when unloaded: 100, 120 and 150 V for inverters I and II, and at 80, 100 and 120 V for inverter III. The three inverters have a different input-voltage-to-output-current gain, which is later analysed in Chapter 5. Fig. 3.24 shows the drain voltage waveforms of the three candidate inverters. Inverter I and II were measured using



(a) Inverter I



(b) Inverter II



(c) Inverter III

Figure 3.24: Drain voltage of candidate inverters.

the 100:1 voltage probe and the waveforms of Inverter III with two 1000:1 voltage probes.

In these experiments, for inverters I and 2, the power consumption of the gate driver was measured at 86 mW when driving the transistor at 6.78 MHz and 140 mW when driving the transistor at 13.56 MHz. For Inverter III the power consumption at the gate roughly

doubles. The gate and the power consumption of an external fan (used as a precautionary measure), measured at 715 mW, are not included in the efficiency calculations in the following chapters of the thesis.

## **3.8 Summary**

This chapter covers the theoretical and practical background of the power electronic circuits which can be used in HF-IPT systems, and emphasises the topologies that are most relevant to the development of the proposed systems architectures and tuning techniques described in the latter chapters of this thesis. The description of the specific designs and construction of the experimental prototypes, and an experimental setup are also described.



# Chapter 4

## Practical HF-IPT Link Operation and Characterisation

The objective of this chapter is to consider the effects of operating a high frequency inductive power transfer (HF-IPT) system in which the link is constrained by certain practical considerations that would be seen in real deployments outside of lab conditions. This includes operation of the wireless link with alternative dielectric materials between the coils (e.g. soil or water), operation of system with corroded coils (e.g. due to the effects of weathering), and understanding the behaviour of non-optimally loaded links on the inverter behaviour. The author's original work presented in this chapter involves the characterisation of an inductive power transfer (IPT) link with air-core coils, the construction of a test load, the experimental verification of Class E and Class EF inverters dealing with detuned loads, a study of HF-IPT through alternative media wireless gaps, the introduction of a reflected impedance estimation technique using waveforms (in collaboration with Mr. Lingxin Lan from the Wireless Power Lab at Imperial College London), and the investigation of the effects produced by corrosion on a HF-IPT link. The latter is a collaboration with Mr. Epameinondas Skountzos from the Department of Materials and Dr. Kyra L. Sedransk Campbell from the Department of Chemical Engineering, Imperial College London. Some of the results presented in this chapter have

been published in three IEEE conference proceedings: [95, 96, 97]. The work presented in this chapter is also relevant to the IEEE conference proceedings [98], where foreign objects are detected by observing the inverter waveforms.

## 4.1 Introduction

Air-core coils, advantageously, can be constructed simply and cheaply using a copper pipe conductor or a printed circuit board (PCB). These, however, are in most cases required to be driven at higher frequencies since the absence of a magnetically permeable core results in lower mutual and self inductance. Since the induced electromotive force ( $\mathcal{E}$ ) from one coil to another depends of  $\sqrt{L_p L_s}$ , the coupling factor ( $k$ ) and the transmit-coil current frequency ( $f_{ip}$ ), (2.7), an increase in frequency can compensate for low inductances and coupling.

There are a number of factors that confine the design options for air-core coils. For example, the Q-factor does not increase indefinitely with frequency, (2.21) and (2.22), as  $R_p$  and  $R_s$  also increase with frequency due to skin and proximity effects. Also, as the length of the conductor that forms the coil ( $l_{coil}$ ) increases to a significant fraction of the wavelength of electromagnetic waves at that frequency, the coil starts to radiate, and thus becomes a parasitic antenna. Therefore, design options such as the shape and number of turns of the coil are limited by:

$$l_{coil} \ll \frac{c}{f_{ip}}, \quad (4.1)$$

where  $c$  is the speed of light. Practically, the length of the conductor should be kept shorter than a tenth of the wavelength at the frequency of operation. With respect to the integration of a HF-IPT system with air-core coils on an application, there are also challenges that arise from not confining the magnetic flux to a core. Metal objects close to the link reflect an impedance on the transmit and receive coils, which might detune the system. This fundamental aspect of the design of HF-IPT systems is addressed in

the examples presented in Chapter 7.

## 4.2 Design of a HF-IPT Link with Air-Core Coils

This section describes the designed air-core coils to operate at 13.56 MHz which are then used throughout this chapter as a basis for the experimental work.

### 4.2.1 Coil Design Considerations

The self-inductance of planar coils can be estimated as proposed in [99], and the equivalent series resistance (ESR) of the coils can be calculated from simulations, as done in [100], where the distribution of current in the conductor, along with the resistivity of the material determine the ESR as a function of frequency. The shape of the coils should be selected taking into account the mutual inductance of two coils, which can be calculated as proposed in [101].

The experimental framework of this thesis proposes a set of HF-IPT systems that operate at megahertz frequencies. For the design of the coils proposed here, a  $f_{ip} = 13.56$  MHz was considered, and also the condition in (4.1) was accounted for.

### 4.2.2 Description of the Designed Coils

Two equal air-core coils were designed and constructed on a PCB with a circular planar shape. Eleven sets of mounting holes were drilled on the PCB in order to recreate a particular link configuration in further experiments, using nylon spacers. The coils have two turns and an external diameter of 210 mm. The width of the conductor is 10 mm and the separation between the turns is 16 mm. The connection between the coils and the power converters at both ends was achieved with two shorted coaxial cables (the shield and the inner conductor are shorted) and shorted edge-mount SubMiniature version A (SMA)

connectors. This configuration was chosen in order to allow integrating a current probe to measure the transmit-coil current ( $i_p$ ). The designed coils can be seen in Fig. 4.1, and a photograph of the built prototype can be seen in Fig. 4.2. The variations of inductance due to changes in the position of the coaxial cables connecting the coils to the boards is considered (4.1 (a), (b)).

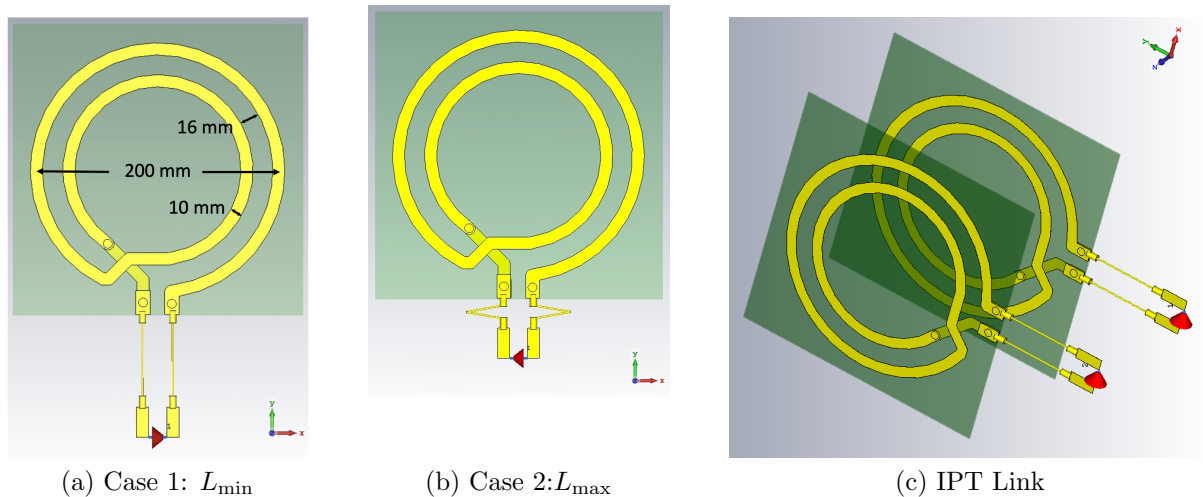


Figure 4.1: PCB coils modelled using CST Microwave Studio.

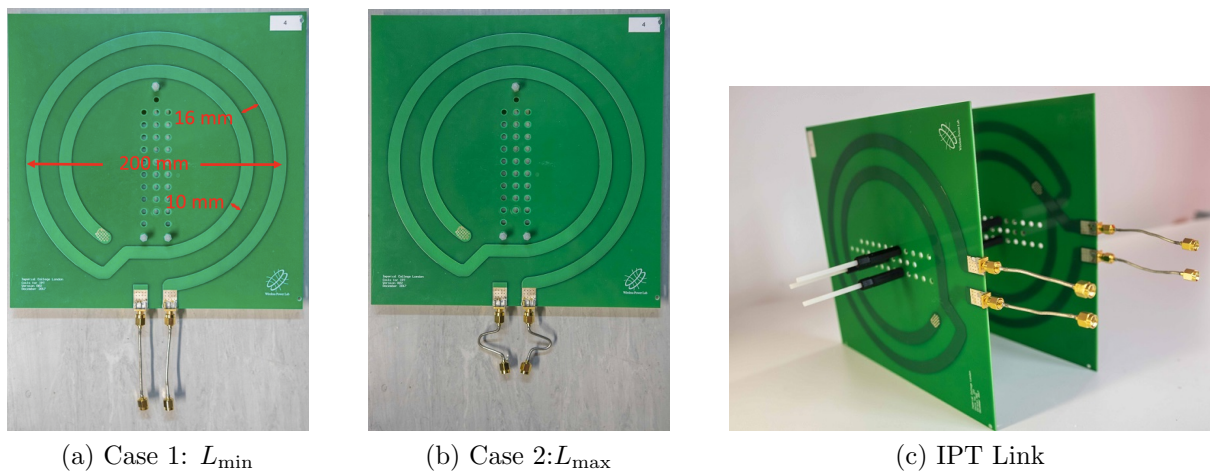


Figure 4.2: Built prototype of the PCB coils.

The design of each coil was done using simulations and once built, the parameters were verified. Then, the coupling factor between the coils was calculated in function of separation and misalignment. The properties of the coils and  $k$  were obtained from electromagnetic simulations using CST Microwave Studio and measurements were performed using

the impedance analyser.

### 4.2.3 Coil Self-Inductance and Equivalent Series Resistance

In the design of the individual power electronic circuits at both ends of the system, it is necessary to account for the parameters of the link. Fig. 4.3, shows the variables to be considered. First, the inductance and the ESR of a single coil were obtained from

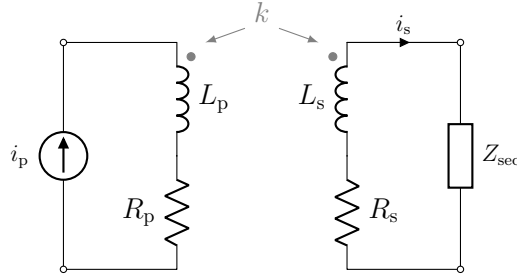


Figure 4.3: General equivalent circuit of an IPT link.

simulations and from measurements using the impedance analyser. The results obtained from simulations are preferred in the design of the link and the circuits since the ESR of the coils is difficult to measure accurately when the Q-factor is high [102, 103]. The results from simulations and measurements are presented in Table 4.1.

Table 4.1: Simulations and measurements of the coils at 13.56 MHz

| Description                       | Inductance (nH) | Resistance (mΩ) | $Q$ factor |
|-----------------------------------|-----------------|-----------------|------------|
| Simulation at case 1: $L_{\min}$  | 1155.7          | 191.2           | 515        |
| Simulation at case 2: $L_{\max}$  | 1200.4          | 189.2           | 541        |
| Measurement at case 1: $L_{\min}$ | 1181.0          | 202.7           | 496        |
| Measurement at case 2: $L_{\max}$ | 1185.2          | 227.2           | 444        |

### 4.2.4 Coupling Distribution in Space

Once the coils have been characterised, individually, the coupling between them can be calculated for different relative positions. Simulations in CST were performed to determine coupling at coil separations of 10 mm to 250 mm, and linear misalignments

of 0 mm to 250 mm.  $k$  can also be verified using an impedance analyser by measuring the equivalent inductance of the primary coil, coupled to the secondary coil when the secondary coil is kept in open circuit and shorted:

$$k = \sqrt{1 - \frac{L_{p,s-\text{short}}}{L_{p,s-\text{open}}}}. \quad (4.2)$$

This technique was used in the proposed IPT link as shown in the photograph in Fig. 4.4, and the results of the measured  $k$  are summarised in the two cases shown in Fig 4.5: the coupling distribution with respect to distance, and the coupling distribution with respect to misalignment.

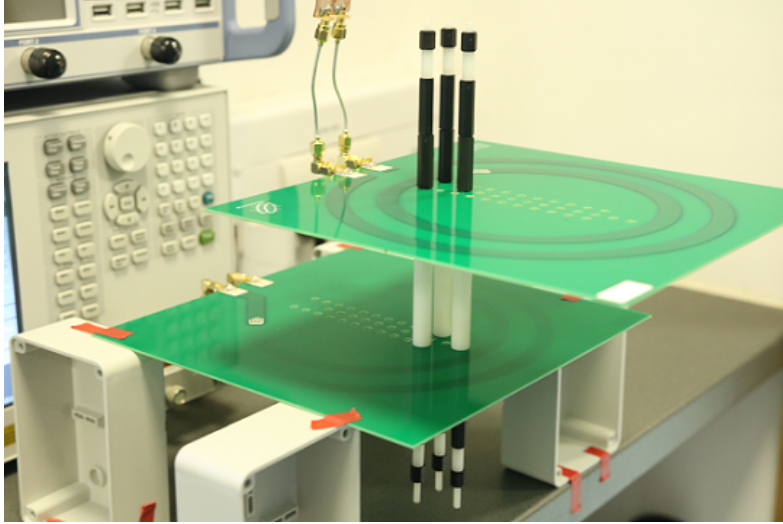
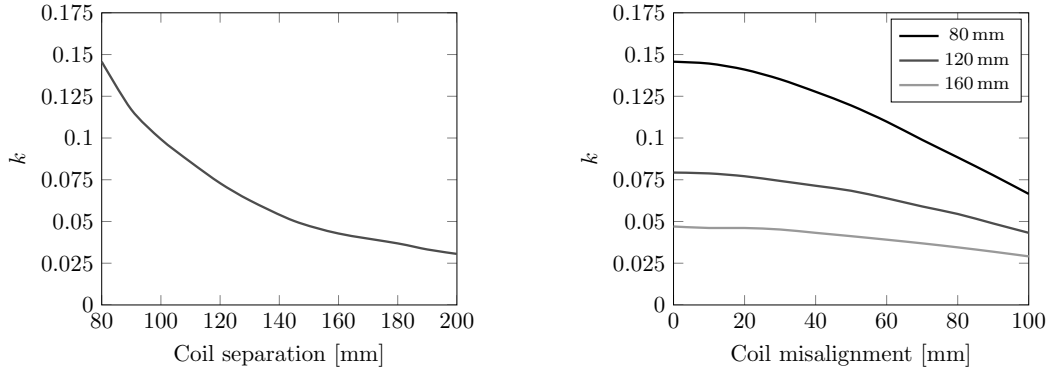


Figure 4.4: Photograph of the coils and setup to measure coupling.

### 4.3 Design and Construction of an Inductively Coupled Test Load

A circuit comprised of a capacitance and a resistance was designed and built to provide a well tuned and measurable load for an IPT system under test. The circuit diagram describing it is presented in Fig. 4.6.  $R_{\text{sec}}$  is comprised of twelve resistors of  $100\ \Omega$  (LTO100F15R00JTE3) in parallel, resulting in a total resistance of  $8.41\ \Omega$ , measured with



(a) Coupling distribution with aligned coils      (b) Coupling distribution at three coil separations

Figure 4.5: Coupling measurements.

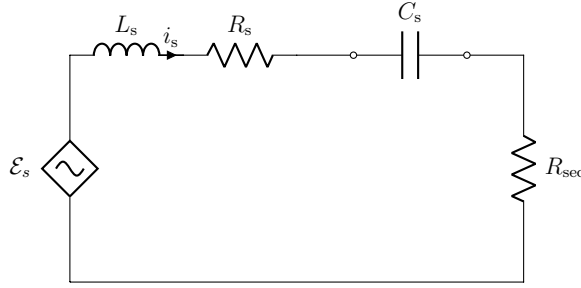


Figure 4.6: HF-IPT test load circuit diagram.

the impedance analyser at 13.56 MHz. Additional measurements were performed at different temperatures (heating up the heatsink via the load resistors to 70°) to verify that as the load temperature increases once the system is running, no significant variations in resistance would be present. A photograph of the test load tuned to resonate with the proposed PCB coils at 13.56 MHz is shown in Fig. 4.7.

The test load was designed to have a loaded-Q:

$$Q_{\text{secloaded}} = \frac{2\pi f_{\text{ip}} L_s}{R_s + R_{\text{sec}}}, \quad (4.3)$$

higher than 10 ( $\approx 11.7$ ) with the proposed coils at resonance at 13.56 MHz, in order to load the transmit side at low  $k$ , and therefore allow for high variations in the reflected reactance on the transmit coil ( $X_{\text{eq}}$ ) with small changes in the resonant capacitance at the receiver ( $C_s$ ).

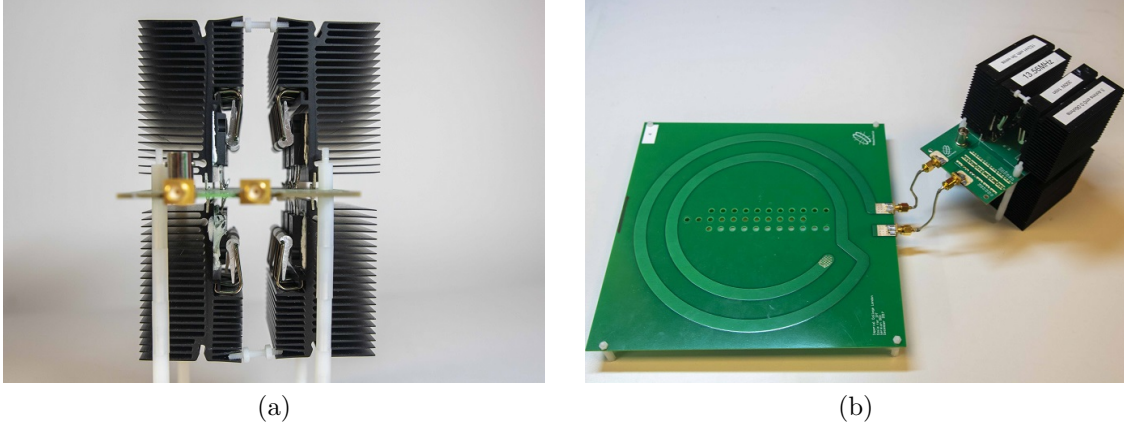


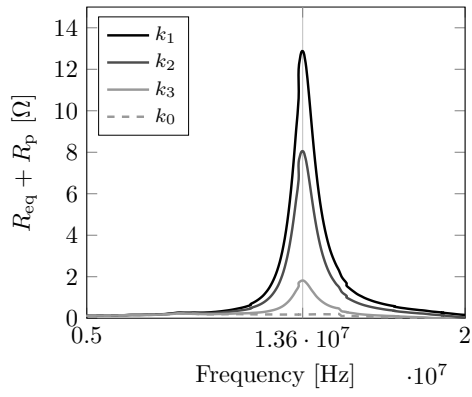
Figure 4.7: Photographs of the HF-IPT test load.

### 4.3.1 Tuning the HF-IPT Test Load at 13.56 MHz

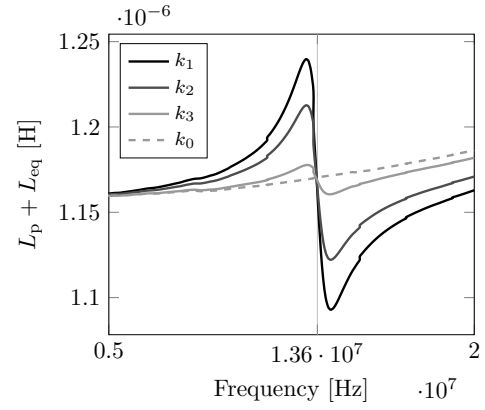
The receive coil inductance, in addition to the parasitic inductance of the board and the inductance of the cables and connectors used to connect the test load PCB to the coil, was tuned out with an external capacitance ( $C_{\text{sec}}$  in Fig. 4.6) using the impedance analyser.

First, the two terminals of the transmit coil which connect to the inverter were used as test points on the impedance analyser to perform inductance and resistance measurements with respect to frequency on the transmit coil. For these measurements, the inverter was disconnected from the coil. Second, the receive coil, connected to the test load, is placed closely to the transmit coil to reflect an impedance on the transmit coil (see equations (2.12) and (2.13)). In the experiments  $k$  was fixed at three arbitrary couplings ( $k_1, k_2, k_3$  in Fig. 4.8). In accordance with (2.12) and (2.13), the maximum reflected resistance is achieved when the receive coil is at resonance, and this also corresponds to when the reflected reactance is zero. Thus, the resonant frequency of the receiver (the receive coil connected to the test load) is identifiable at the point of the maximum equivalent resistance and when the measured inductance is equal to the one measured on the transmit coil when uncoupled from the receive coil ( $k_0$ ). Finally,  $C_s$  is adjusted until the resonant frequency of the receiver matches the frequency at which the load should be tuned. Fig. 4.8 shows how the maximum resistance is measured at exactly 13.56 MHz,

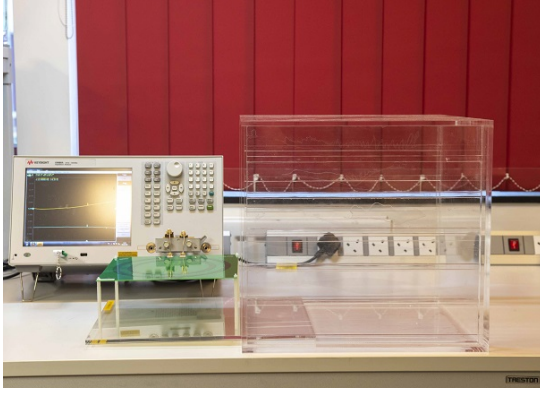




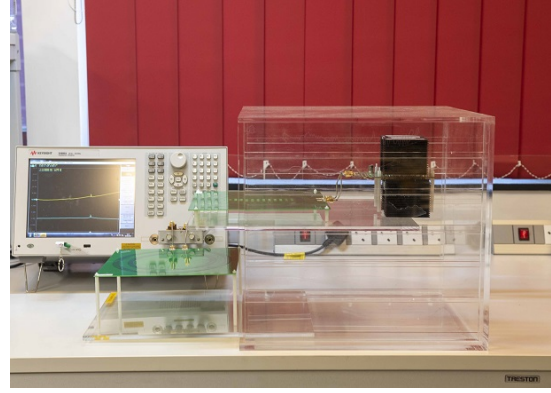
(a)



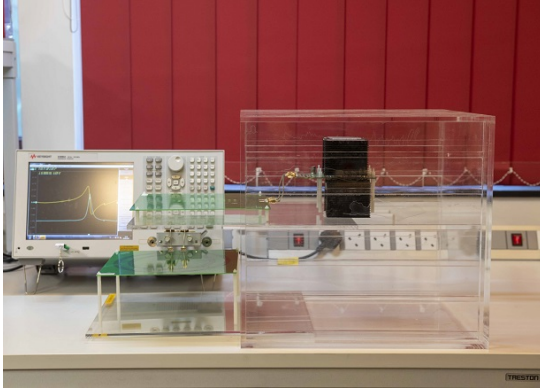
(b)



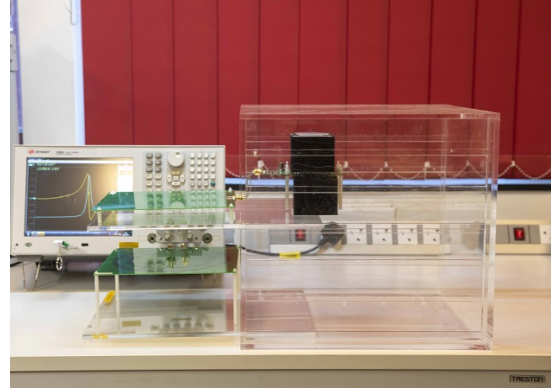
(c)  $k_0$



(d)  $k_3$



(e)  $k_2$



(f)  $k_1$

Figure 4.8: Experimental tuning of the test load.

and that it also corresponds to where the inductance measurement matches the one of when the coils are not coupled.

It should be noted that a receiver with an IPT-rectifier instead of a test load cannot be tuned using this technique as a larger signal would be required to activate the diodes and obtain a representative input impedance of the IPT-rectifier ( $Z_{\text{sec}}$ ), for the fundamental

frequency of the current.

This technique can also be used to obtain  $k$  experimentally using (2.12):

$$k = \sqrt{\frac{R_{\text{eq}}(R_s + R_{\text{sec}})}{4\pi^2 f_{\text{ip}}^2 L_p L_s}}, \quad (4.4)$$

from which every variable is known or can be measured with the impedance analyser. From this, the arbitrary couplings of Fig. 4.8 are calculated as:  $k_1 = 10.3\%$ ,  $k_2 = 8.2\%$  and  $k_3 = 3.9\%$ .

### 4.3.2 Controlled Detuning of the HF-IPT Test Load

In order to provide a useful test load for future experiments, the load was purposely and progressively characterised while it was detuned by modifying  $C_s$  to not only reflect a resistance but also a positive and negative reactance to the transmitter. The load resistance was kept constant as the load was purposely detuned. The test load was characterised as shown in Table 4.2.

Table 4.2: Test load tuning

| Load type  | Resistive | Capacitive (reflects inductance) |       |       | Inductive (reflects capacitance) |       |       |       |
|------------|-----------|----------------------------------|-------|-------|----------------------------------|-------|-------|-------|
| Identifier | R         | RC1                              | RC2   | RC3   | RL1                              | RL2   | RL3   | RL4   |
| $C_s$ [pF] | 109.8     | 108.7                            | 107.4 | 104.8 | 111.0                            | 112.2 | 113.4 | 114.6 |

Different reflected impedance on the transmit side ( $Z_{\text{eq}}$ ) with the same tuning were achieved by changing the separation of the coils (120 mm, 150 mm and 180 mm). A summary of reflected loads is shown in Table 4.3. The construction of loads tuned at different frequencies, where the reflected impedance can be calculated, was done to investigate the effects produced by reflecting positive and negative reactances to candidate inverters, which is necessary considering the reflected reactance on the transmit coil that IPT rectifiers might produce as the load changes.

Table 4.3: Inductively coupled load tuning and measured reflected impedance

| Load | $R_{\text{sec}} [\Omega]$ | $2\pi f_{\text{ip}} L_s [\text{j}\Omega]$<br>$+X_{\text{sec}}$ | Coupling | $R_{\text{eq}} [\Omega]$ | $X_{\text{eq}} [\text{j}\Omega]$ |
|------|---------------------------|----------------------------------------------------------------|----------|--------------------------|----------------------------------|
| R2   | 8.41                      | 0                                                              | 5.3 %    | 3.32                     | 0                                |
| R3   | 8.41                      | 0                                                              | 7.9 %    | 7.37                     | 0                                |
| RC32 | 8.41                      | -5.19                                                          | 5.3 %    | 2.44                     | 1.46                             |
| RC33 | 8.41                      | -5.19                                                          | 7.9 %    | 5.41                     | 3.25                             |
| RL42 | 8.41                      | 4.46                                                           | 5.3 %    | 2.62                     | -1.35                            |
| RL43 | 8.41                      | 4.46                                                           | 7.9 %    | 5.81                     | -3.01                            |

## 4.4 Experiment I: Initial Investigations into the Effects Produced by Reflected Reactances in Candidate 13.56 MHz Inverters

In this experiment, the effects of reflecting different loads on two candidate inverters are investigated. Namely, Inverter I: Class E and Inverter II: Class EF were compared when powered from a 120 V DC input voltage ( $V_{\text{dc}}$ ) while driving the different loads specified in Table. 4.3. The results of these experiments were published in [95].

### 4.4.1 Introduction

The form in which the input impedance of an IPT rectifier changes with the IPT system DC load ( $R_o$ ) and  $k$  depends on the topology and design of the rectifier. Class E rectifiers can be designed to maintain Class E switching under these sorts of conditions (broad changes in  $k$  and load demand) by tuning the passives at a frequency different from the system's operating frequency, and then tuning out the residual reactance of the rectifier in the resonant tank of the receiver [87]. Designing rectifiers in this way, however, relies not only on altering the rectifier's input resistance to regulate power as  $k$  or the load changes, but also on varying the input reactance, which can slightly detune the receiver and therefore reflect a reactive load to the transmitter. This phenomenon is not exclusive to loads that use Class E rectifiers. Other sorts of loads, e.g. inductively

coupled plasma [104], tend to reflect a reactance to the inductive power source. Also, foreign objects reflect reactances to the transmitter.

#### 4.4.2 Experimental Verification

The HF-IPT test load, described in Section 4.3, was tuned as proposed in Table 4.3 to determine the effects produced by changes in the reflected impedance on two candidate inverters: Inverter I: Class E and Inverter II: Class EF; both at power levels lower than 200 W.

The inverters were first tested when feeding a purely resistive load at various couplings, then a capacitive-resistive load and finally an inductive-resistive load were tried. Measurements were taken for all cases in which the temperature of the transistor did not exceed 80 °C. The input power, measured for each of the previously characterised loads, is presented in Table 4.4. The loads RC33 and RL43 could not be powered with Inverter II since the temperature of the inverter was exceeding 80 °C shortly after powering the system.

Table 4.4: Input power at the transmit side with different reflected impedances [W]

|             | Type     | No load | R2   | R3    | RC32 | RC33  | RL42 | RL43 |
|-------------|----------|---------|------|-------|------|-------|------|------|
| Inverter I  | Class E  | 14.4    | 63.6 | 108.0 | 52.8 | 109.2 | 49.2 | 81.6 |
| Inverter II | Class EF | 13.2    | 84.0 | 162.0 | 61.2 | -     | 66.0 | -    |

The drain voltage ( $V_{ds}$ ) and  $i_p$  waveforms for both inverters under resistive and resistive-reactive loads are presented in Fig. 4.9 for the Class E inverter and in Fig. 4.10 for the Class EF. Fig. 4.11 shows the experimental set-up for the measurements.

The effects of varying the resistive loads in Class E inverters and load-independent Class EF inverters have been extensively investigated [64, 65, 73], and the effects of a reflected reactance can be represented as a variation in  $C_3$  (see Figs. 3.3 and 3.7.), i.e. the inverter being detuned. Interestingly, if the receiver has a fixed residual reactance, the reflected resistance and reactance both change as  $k$  changes. Therefore, the effects

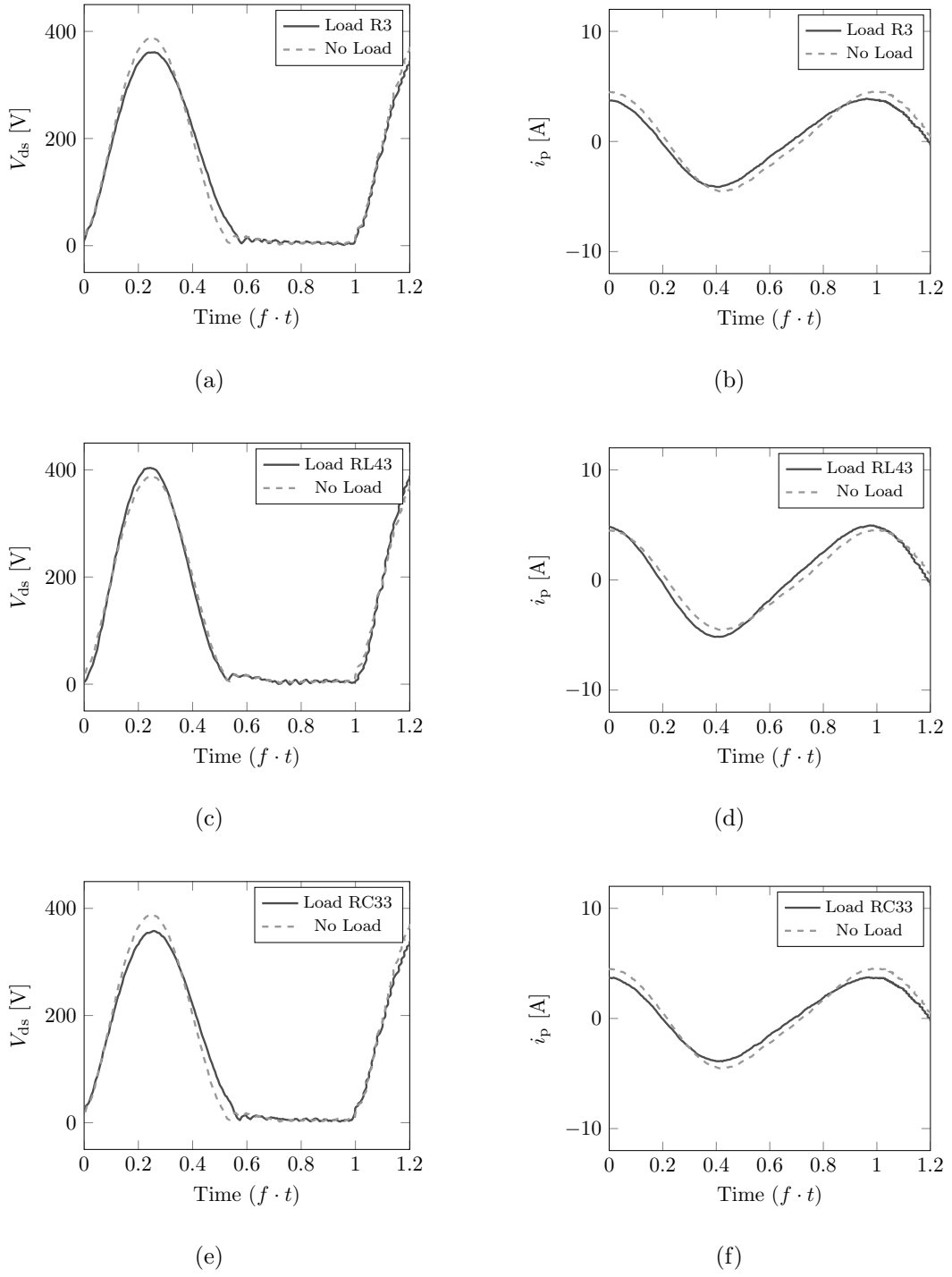


Figure 4.9: Measured waveforms of Inverter I: Class E driving different loads.

of a slightly detuned load moving across a charging pad affects both  $R_{eq}$  and  $X_{eq}$  at the transmit side.

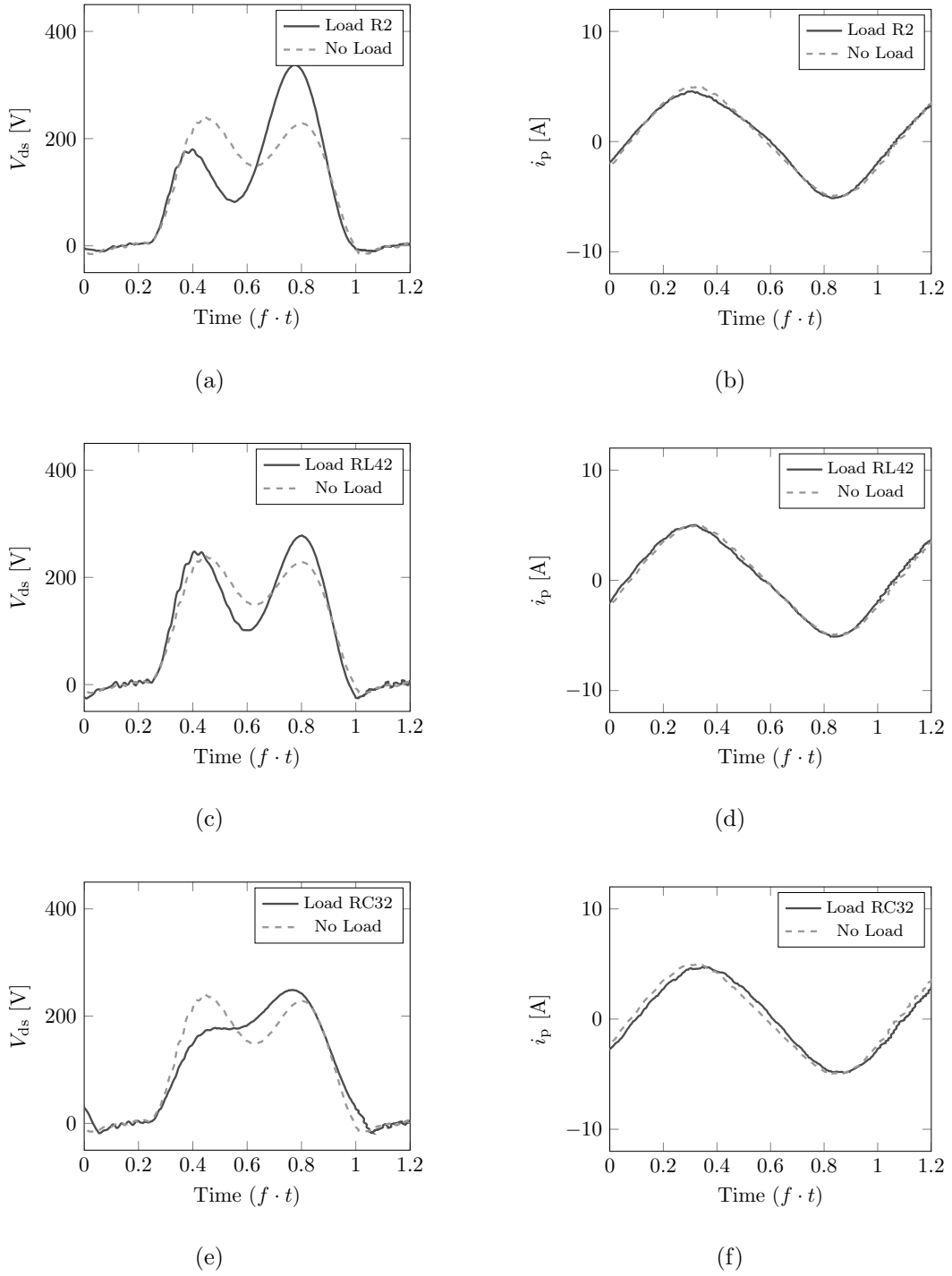


Figure 4.10: Measured waveforms of Inverter II: Class EF driving different loads.

### 4.4.3 Discussion

Inverter II: Class EF was designed for resistive loads from 0 to 6  $\Omega$  and Inverter I: Class E can tolerate a broader resistive load range (0 to 10  $\Omega$ ) but with the disadvantage of having a short period of time with sub-optimum operation (the transistor's anti-parallel

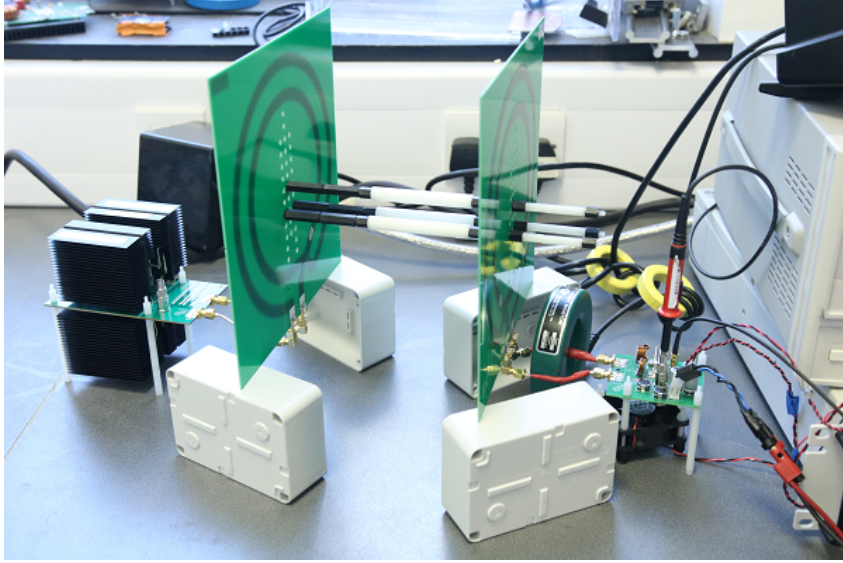


Figure 4.11: Photograph of IPT experimental setup.

diode conducting) when the reflected resistive load is not equal to the maximum load. When a capacitive load is reflected to the transmitter (loads RL42 and RL43) both inverters tend to enter sub-optimum operation. Interestingly, the Class E drain voltage waveform is similar to that of the unloaded condition. When an inductive load was reflected to the transmitter (RC32 and RC33) the load-independent Class EF loses zero-voltage switching (ZVS), while the Class E inverter shows a similar behaviour to that of driving a purely resistive reflected load. This can be seen in Fig. 4.10 at turn-on,  $f \cdot t = 1$ .

## 4.5 Experiment II: HF-IPT with Alternative Media Wireless Gaps

In this experiment, the effects of introducing materials other than air between the coils of a HF-IPT system is investigated. The materials introduced between the coils are specific to a case where a HF-IPT transmitter or receiver is designed to be buried under soil in a park in London.

### 4.5.1 Introduction

Wireless power transfer (WPT) is desirable in applications where the wireless gap is not just comprised of air, but another material. The challenge of performing WPT through alternative media has been addressed before. In [105], for example, a sensor embedded in concrete was charged by the means of electromagnetic waves at gigahertz frequencies. The work in [106], demonstrates IPT to low power devices in or in proximity to water in order to propose the technology as an alternative to power sensors in water treatment environments; and the work in [107], showcases a low frequency IPT system operating inside a tank containing a brine solution.

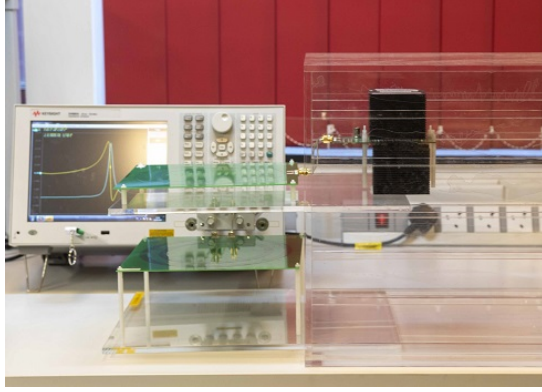
With the purpose of investigating the viability of employing HF-IPT through alternative media, a test rig (Fig. 4.12), consisting of a perspex enclosure with a set of compartments was built. The compartments have capacity for a HF-IPT transmitter, a receiver or a material under test. Each compartment has 50 mm in height without including the perspex sheet, which is 8 mm thick. The width and length are 300 mm and 400 mm respectively.

The materials employed in this experiment are air, crushed brick, wet soil and fresh water. These materials were selected to be representative in cases where either the receiver or the transmitter is buried underground; however, the materials employed are specific to those found in Princes Gardens park in South Kensington, London, and the conclusions from this experiment are specific to those conditions.

### 4.5.2 Inductance and Resistance Measurements

The impedance measurement setup proposed to tune the HF-IPT test load (Section 4.3) was implemented to measure the  $\mathbf{Z}_{eq}$  produced by the test load and a layer of the material under test between the coils. The measurements of equivalent inductance and resistance of the transmit coil, coupled to the HF-IPT test load, with the alternative media between the coils, are presented in Fig. 4.13.

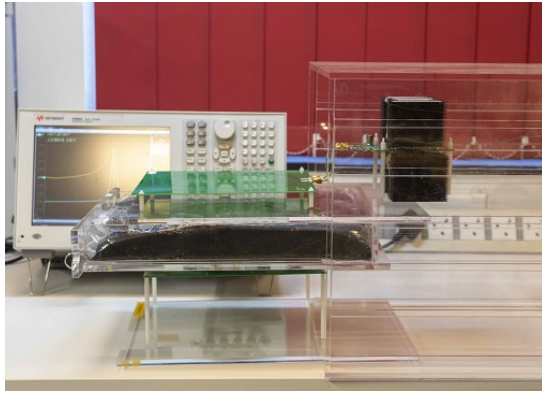




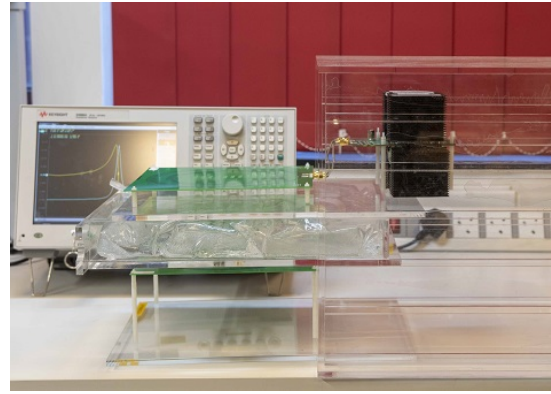
(a) Air



(b) Brick



(c) Soil



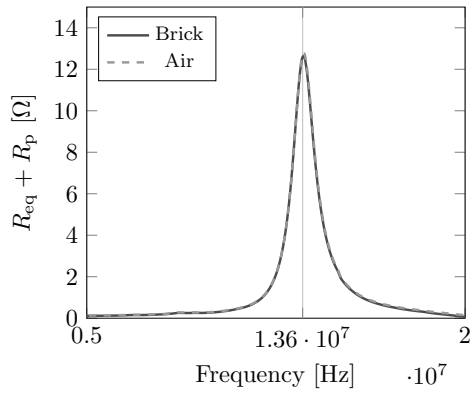
(d) Water

Figure 4.12: Photographs of the test rig developed for the characterisation of an inductive link with alternative media.

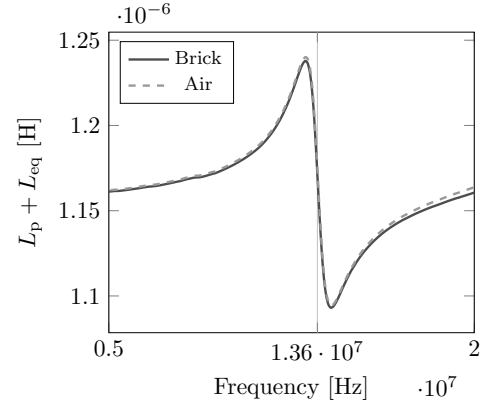
### 4.5.3 HF-IPT End-to-End Experiments at 13.56 MHz

Once the reflected impedance was characterised, the test load was exchanged with a HF-IPT receiver in order to measure output DC power. The test rig in Fig. 4.14 shows the DC to DC HF-IPT system employing Inverter II: Class EF and Rectifier II: VM Class D.

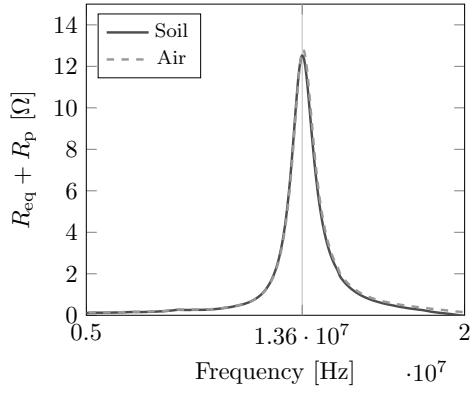
The load-independent Class EF inverter was selected because, as demonstrated in the Experiment I (Section 4.4), the shape of the drain voltage waveform changes with respect to the amplitude and angle of  $\mathbf{Z}_{\text{sec}}$ , and therefore the waveforms could be used to identify large changes on  $\mathbf{Z}_{\text{eq}}$ . At the receive side, the voltage-multiplier Class D rectifier was selected because its large AC to DC gain allows achieving high voltages at low  $k$ , thus making the difference in the induced electromotive force on the receive coil ( $\mathcal{E}_s$ ) more easily



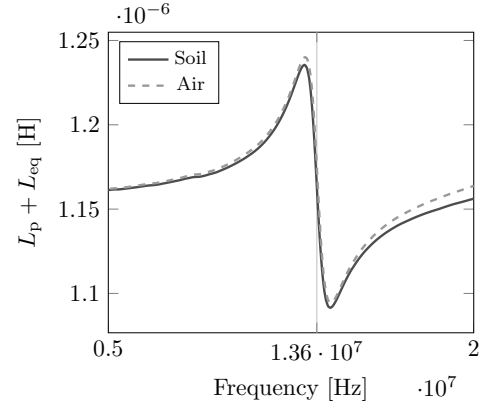
(a)



(b)



(c)



(d)

Figure 4.13: Measured resistance and reactance of the transmit coil coupled to the receivers under test.

identifiable.  $R_o$  ( $124\Omega$ ) and  $k$  (5.3%) were kept constant throughout the experiments. The results are shown in Table 4.5

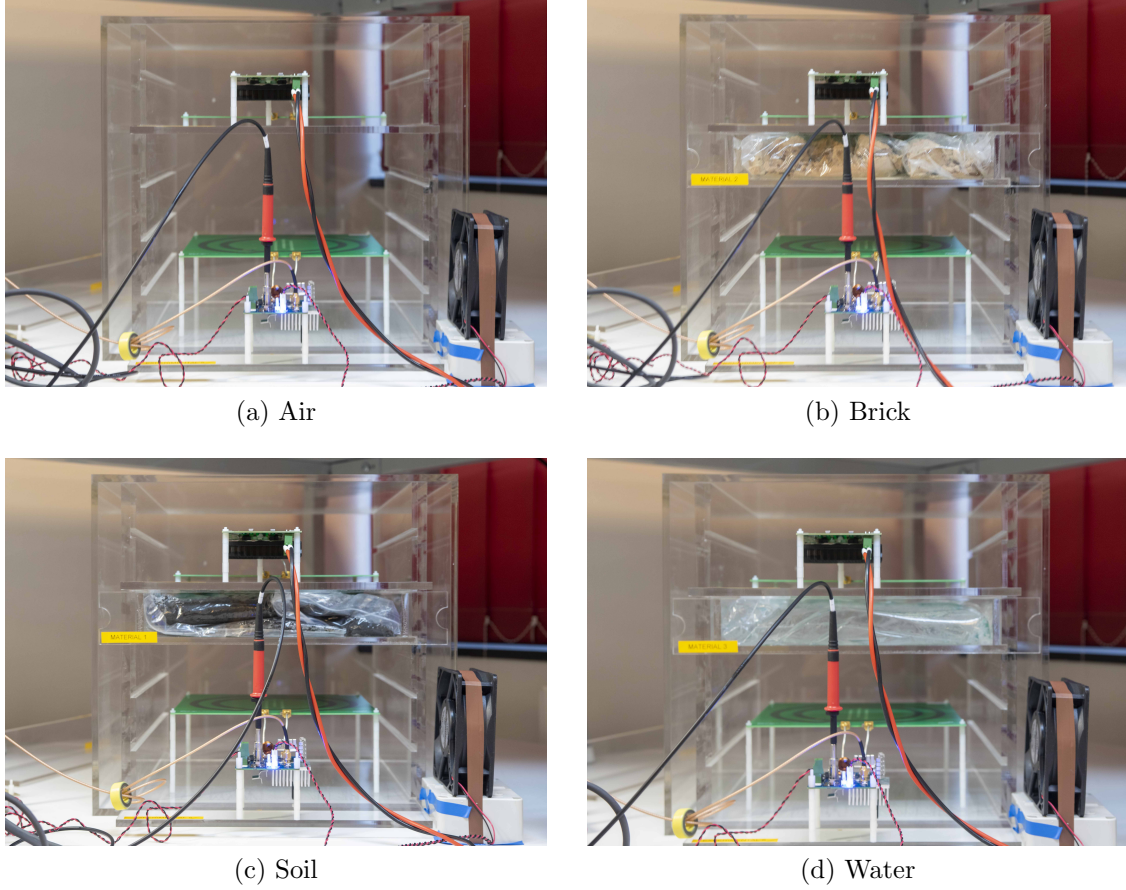


Figure 4.14: DC to DC HF-IPT system performing wireless power transfer through alternative media.

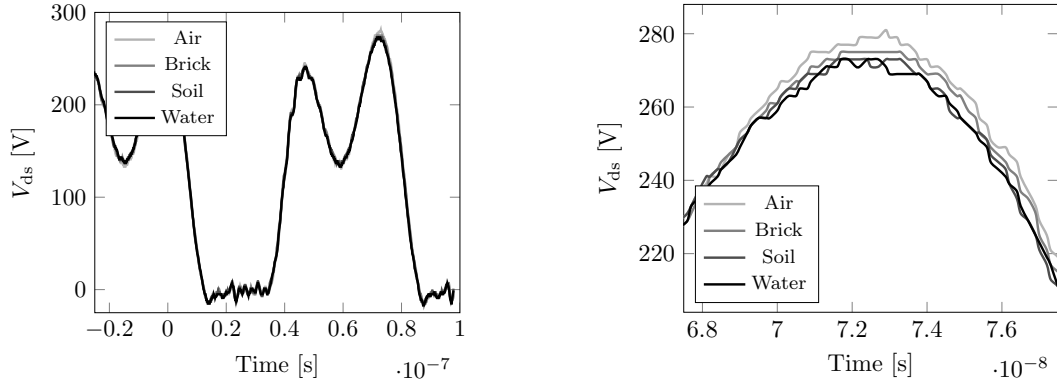
Table 4.5: Experimental results of DC to DC HF-IPT with alternative media wireless gaps

| Media | $P_i$ [W] | $V_o$ [V] | $i_o$ [A] | $P_o$ [W] | Total losses [W] | $\eta_{dc-o}$ |
|-------|-----------|-----------|-----------|-----------|------------------|---------------|
| Air   | 108.9     | 100.8     | 0.813     | 82.0      | 26.9             | 75.2 %        |
| Brick | 105.7     | 98.5      | 0.798     | 78.7      | 27.0             | 74.4 %        |
| Soil  | 101.7     | 95.6      | 0.7645    | 73.1      | 28.6             | 71.8 %        |
| Water | 101.0     | 95.0      | 0.752     | 71.4      | 29.6             | 70.7 %        |

The drain voltage waveforms of the inverter are shown in Fig. 4.15.

#### 4.5.4 Discussion

Minimal changes can be observed in the measured resistance and inductance (Fig. 4.13) of the transmit coil, coupled to the test load with the materials under test between the coils. However, in the DC to DC experiments, a decrease in the reflected resistance on



(a) Drain voltage waveform

(b) Drain voltage waveform (detail)

Figure 4.15: Inverter drain voltage waveforms performing wireless power transfer through alternative media.

the transmit coil ( $R_{eq}$ ) and in the amplitude of  $\mathcal{E}_s$  can be identified. When the layer of water was introduced, the output power decreased by 13 % with respect to air, which demonstrated that  $\mathcal{E}_s$  decreased. In the case of water, since the input power also decreased (7.5 % with respect to air), it can be concluded that  $R_{eq}$  was lower in this case. From the alternative media under test, water was the one with the highest impact on output power and efficiency.

The performance of the circuits at both ends of the system was not affected by the materials under test between the coils. As can be seen in Fig. 4.15, the switching performance of the inverter does not change with the different materials. A slightly higher peak-voltage is reached (Fig. 4.15 (b)), in the case where the link is solely comprised of air, which suggests that when alternative media were placed between the coils, the inverter became slightly less loaded, which is congruent with the measurements of Table 4.5. Furthermore, the measured changes when the materials under test were placed between the coils are relatively minor and no indication of sub-optimum operation, such as an increase in the temperature of the transistor, was observed in the experiments.

In summary, the investigated alternative media do not produce major effects on the operation of the system. These experiments, can therefore be used as an initial validation study for HF-IPT underground devices or to perform WPT from an underground source.

## 4.6 Experiment III: Reflected Impedance Estimation Using Waveforms

In this section, a technique to estimate  $\mathbf{Z}_{\text{eq}}$  was developed by solely observing changes on the drain voltage waveforms of the inverter. The results of this case study were published in [96], and a similar approach was implemented to identify foreign objects within reach of the transmit coil in [98].

### 4.6.1 Introduction

Modelling a HF-IPT system using the equations and equivalent circuits derived in Chapter 2 allows decoupling the transmitter and receiver by representing their respective interaction as  $\mathbf{Z}_{\text{eq}}$  at the transmit side and  $\mathcal{E}_s$  at the receive side. The respective circuits can therefore be designed considering the range of operation for these variables without necessarily knowing the specifics of the link or the other end of the system.

Knowledge of the reflected impedance is key in the design of the inverter for systems that allow for spatial freedom and variable power demand.  $\mathbf{Z}_{\text{eq}}$ , which not always purely resistive has to be confined to the range of tolerance of the inverter, both with respect to resistance and reactance.

### 4.6.2 The Reflected Impedance Estimation Principle

The proposed technique to estimate  $\mathbf{Z}_{\text{eq}}$  consists of gathering signals from the inverter, from simulation or experiments, while power is transferred to known loads, in order to classify the waveforms in the load domain. A model can then be programmed to identify the load based on measurements at the transmit side using the previously collected and classified data.

### 4.6.3 The Reflected Impedance Estimation Technique Using a Class EF Inverter

The experimental waveforms in Fig. 4.9 (Inverter I: Class E) and Fig. 4.10 (Inverter II: Class EF) suggest the possibility of estimating  $\mathbf{Z}_{eq}$  using solely the drain voltage waveforms since changes can be identified when the load changes its magnitude and phase. Most notably, the drain voltage of the Inverter II: Class EF has very distinguishable variations when  $R_{eq}$  and  $X_{eq}$  change. From simulation, the waveforms of this inverter change with load as shown in Fig. 4.16.

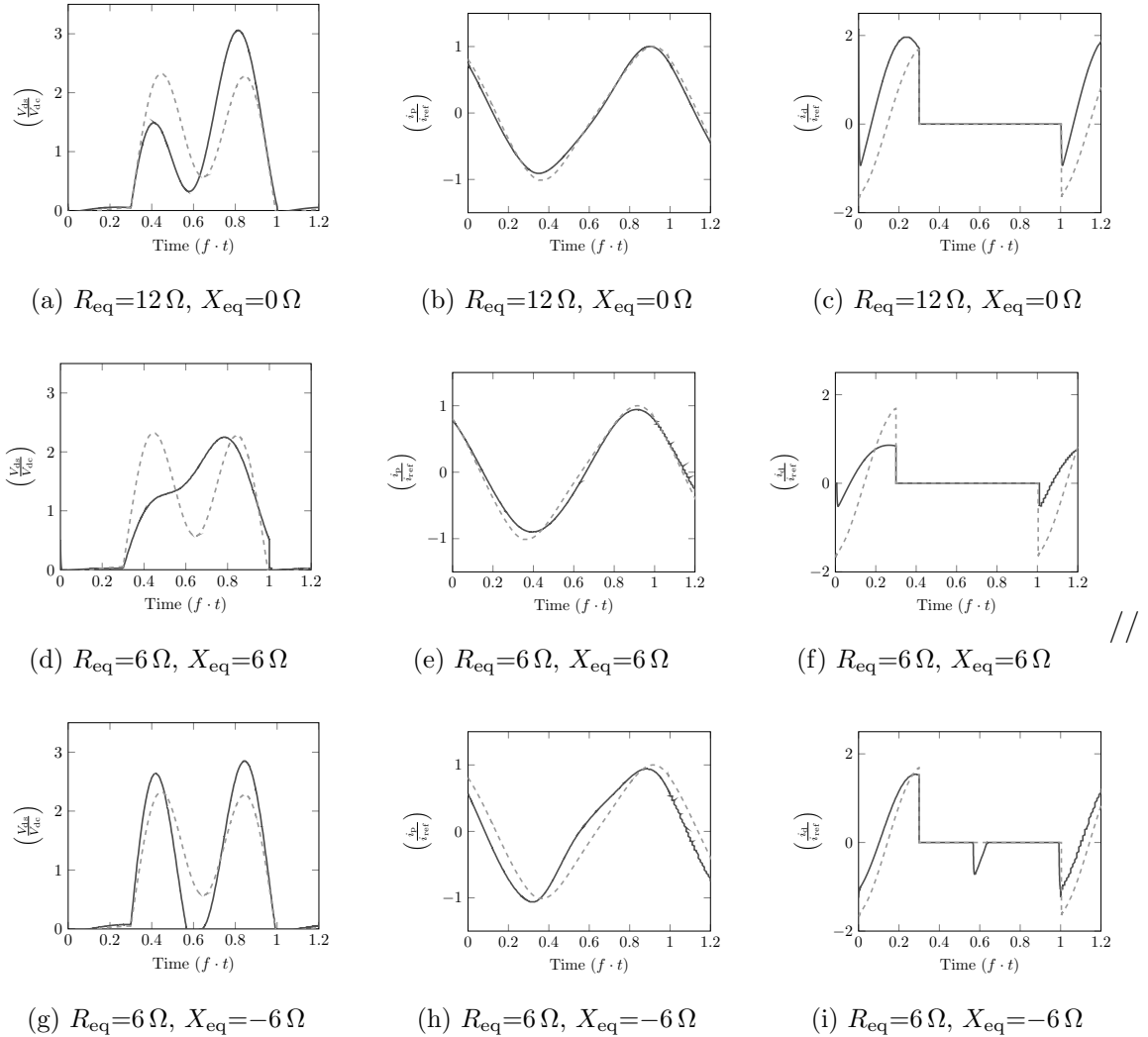


Figure 4.16: Theoretical waveforms of a load independent Class EF inverter with detuned loads.

These theoretical waveforms showcase a range of  $\mathbf{Z}_{eq}$  broader than the one in the practical

experiments in the Experiment I (Section 4.4) as this range is greater than the one tolerated by the inverter without getting damaged. From the waveforms in Fig. 4.16, a clear difference in the shape of the drain voltage waveform ( $V_{ds}$ ) is observed once  $\mathbf{Z}_{eq}$  changes, and since  $V_{ds}$  is relatively easy to measure (as opposed to for example  $i_d$ ) it is well suited as the parameter to estimate the reflected impedance by a candidate receiver.

Once a known load is reflected to the transmitter using the inductive link and the HF-IPT test load developed in Section 4.3, the data necessary to create an impedance estimation model is gathered and classified. Two common machine learning algorithms were considered to train the model, support vector machine (SVM) and convolutional neural network (CNN). SVM, described in [108], is commonly used in classification applications and in regression models, e.g. stock market prediction [109, 110]. SVM can be used to train a model to estimate  $\mathbf{Z}_{eq}$ , however the proposed model should be capable of achieving this regardless of the phase of the drain waveform. Therefore, an SVM-based approach would require data pre-processing before training. In order to avoid the data pre-processing, CNN was selected to train the estimation model, since it can recognise the patterns without being affected by shifts in position [111], i.e. phase. CNN has been used in various classification applications such as medical signal processing [112] and face recognition [113].

The training data are divided into three subsets: 85 % of the training data are used to train the CNN model, 10 % as the validation data set during the training process, and the last 5 % as the final test to evaluate the performance of the trained model. The network layers were configured manually based on the performance of the model.

#### 4.6.4 Verification of the Reflected Impedance Estimation Model

The proposed technique was first verified by simulating Inverter II: Class EF using SPICE, driving different linear loads, and then experimentally. A model was derived to predict the reflected impedance from both simulation and experimental waveforms.

## Circuit Simulation

Inverter II: Class EF was simulated in SPICE, driving a reflected impedance, from which the real part ( $R_{eq}$ ) was set from 0 to 10  $\Omega$ , and the imaginary part ( $X_{eq}$ ) from  $-5$  to 5  $\Omega$ , both with steps of 100 m $\Omega$ . The drain-voltage waveforms were recorded when the system reached steady-state for each reflected load, and then were classified with respect to  $\mathbf{Z}_{eq}$ . These data were then used to train and verify the model. The reflected load prediction using the simulated drain voltage waveform is shown in Fig. 4.17. The estimation accuracy (coefficient of determination) is 0.9942 for the real part of the impedance, and is 0.9905 for the imaginary part.

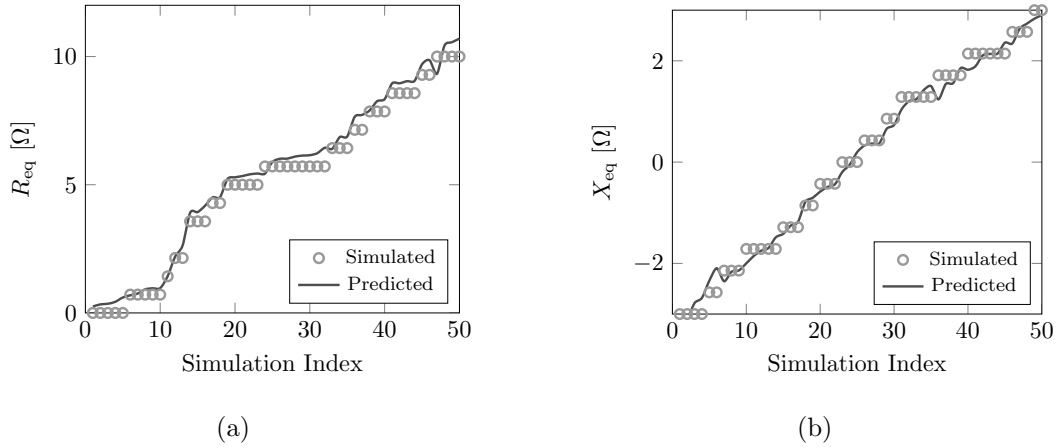


Figure 4.17: Results of the impedance prediction from simulation.

## Experimental Verification

A similar process to gather data was done on the built HF-IPT system under test (Fig. 4.11). The set of loads described in Table 4.3, with the addition of ones obtained from changing the position of the coils from 12 to 20 cm with steps of 1 cm, were used to train the model. Four loads which reflect a negative reactance and three which reflect a positive reactance were used (RL1-RL4 & RC1-RC3 from Table 4.2) at the coil separations mentioned. The reflected reactance allowed by this setup was roughly from  $-2$  to 2  $\Omega$ , and the reflected resistance from 0 to 8  $\Omega$ . The  $\mathbf{Z}_{eq}$  prediction values of the exper-



iments are shown in Fig. 4.18. The estimation accuracy (coefficient of determination) is 0.9899 for the real part of the impedance, and is 0.9743 for the imaginary part. In the plots in Fig. 4.18, the model achieves accurate predictions with the exception of the case when the reflected reactance had a high positive value, suggesting that the model needs further training for such cases.

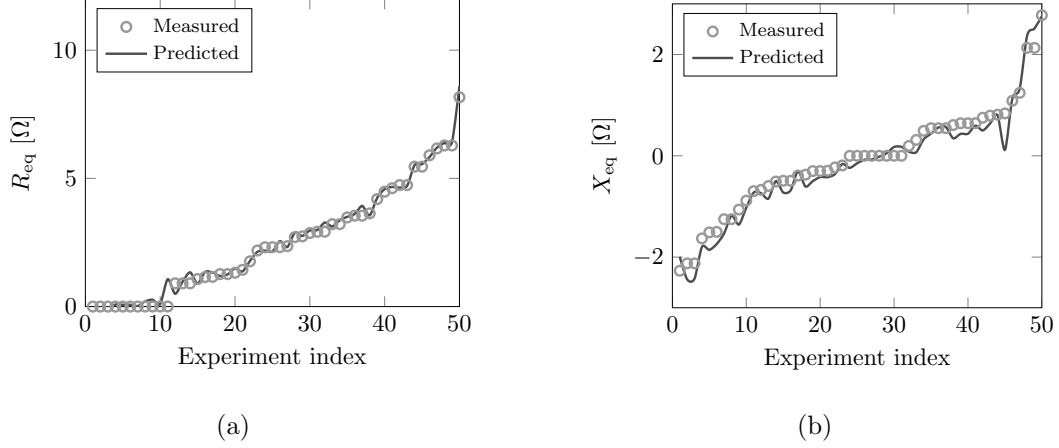


Figure 4.18: Results of the impedance prediction from experiments.

#### 4.6.5 Discussion

These experiments demonstrate an accurate prediction of  $\mathbf{Z}_{eq}$  achieved from solely observing and comparing the transmit side waveforms. This proof of concept was performed on simulation and with experiments separately. Ideally, a model should be generated from simulation, where a much larger set of data can be gathered, without requiring training a built system for all cases which produce changes in  $\mathbf{Z}_{eq}$ . This poses a challenge because in such case, the SPICE model has to represent what is measured by the on-board equipment, and not the signals that would be obtained from ideal measurements. In addition, the experiments were performed using a LeCroy oscilloscope, while a realistic setup for live on-board measurements would likely include an analogue to digital converter with a much lower bandwidth than the one in the oscilloscope, therefore affecting the accuracy of such prediction. These challenges, however, can be tackled, realistically, to perform on-board live  $\mathbf{Z}_{eq}$  measurements. For example, since changes  $\mathbf{Z}_{eq}$  are most likely to be

much slower than the frequency of operation, as these depend of motion or load changes (which are usually coupled to large reactive components), the inverter waveforms can be constructed within a number of periods of the inverter switching signal and before a detectable change in  $Z_{eq}$  can be measured.

The proof of concept of this section can be the basis of a live on-board measurement technique which allows system level decision making in a HF-IPT system integrated in an application, based on the assessment of  $Z_{eq}$  which gives information of the electrical circuit that is coupled to the transmitter.

## 4.7 Experiment IV: Characterisation of Corroded HF-IPT Coils

An additional practical aspect of operating HF-IPT systems in the field, is the possibility of system damage and degradation due to water ingress and/or corrosion. This section describes experiments which investigate the effects of corrosion on air-core coils for HF-IPT. The work presented in this section is a collaboration between members of the Wireless Power Lab and members of the Campbell Group from the Department of Chemical Engineering of Imperial College London. These results were published in [97].

### 4.7.1 Introduction

Limited studies detailing atmospheric corrosion on electronic materials can be found in the literature, e.g. [114], where a review of the effects of corrosion on electronic circuits and PCBs was carried out. Additionally, assessment of corrosion in radio frequency identification (RFID) has also been presented [115]. Corrosion resistance of materials in electronics are influenced by a wide range of environmental variables, e.g. temperature, relative humidity, presence of corrosive gases, as well as fabrication defects and high-voltages [116]. The latter poses a notable concern due to the presence of high voltages in

the coils used in HF-IPT.

The work presented in this section demonstrates the performance of a HF-IPT system after the transmit coil has had exposure to relevant and potentially corrosive conditions. This work includes analysis of changes to the coil during corrosion (using optical and electron microscopy, as well as elemental analysis) to establish the impact on performance metrics.

#### **4.7.2 Association of Power Dissipation on the Coils and Corrosion**

The possible effect of corrosion on the ESR of the coils, and consequently on the Q-factor was raised in [31] where a Q-factor of 2500 was expected from calculations, but it was measured at 950. Authors claim that the unexpected lower value of the Q-factor was probably due to a less conductive copper-oxide layer on the copper conductor that forms the coil, and since the skin effect at these frequencies (9.9 MHz) causes a much higher current density at the edge of the conductor (the skin depth of copper is approximately 21  $\mu\text{m}$  at 9.9 MHz and 18  $\mu\text{m}$  at 13.56 MHz), this layer of oxide would increase the coil ESR. This work aims to corroborate or disprove that hypothesis.

#### **4.7.3 Corrosion of the Immersion Silver Finish of the Coils Under Test**

Since air-core PCB coils allow achieving high Q-factors ( $>500$ ) when operating at the two first industrial, scientific and medical (ISM) bands (Table 4.1), two of the two-turn circular PCB coils described in Section 4.2.2 were selected to conduct the experiments. These coils under test were fabricated on a 1.6 mm thick FR4 substrate PCB with two-ounce-copper, a silver immersion finish of 0.1  $\mu\text{m}$  to 0.2  $\mu\text{m}$ , and no solder mask.

Copper, due to its good electrical conductivity and relatively low cost, is the most

widely used metal for PCBs. However, copper corrodes when exposed to oxidising conditions [117], thereby requiring a further protective layer (which also offers ease for soldering). Most commonly, a surface finish (for lead-free electronics), which was employed in the fabrication of the coils under test, is the immersion silver finish [117]. It is characterised by both high solderability and wettability at a considerably lower cost than most surface finish technologies.

Due to this fabrication method, consideration of corrosion must include: silver, the interface between silver and copper, and copper. The exposure of silver to a chloride-rich atmosphere can form a layer of non-conducting silver chloride ( $\text{AgCl}$ ) on the surface, thereby limiting the shelf-life and solderability [117]. Secondly, due to the contact between silver and copper, galvanic corrosion has been reported to occur. Galvanic corrosion occurs when two metals are in contact and the more noble metal is protected, whilst the more active metal dissolves and corrodes [116]. For PCBs, the active metal is copper and corrosion leads to the evolution of an oxide layer on the surface of copper [118]. The initial phase of the oxide layer is usually cuprite ( $\text{Cu}_2\text{O}$ ) or tenorite ( $\text{CuO}$ ). Thirdly, for copper corrosion in marine environments, the chloride-rich nature results in nantokite ( $\text{CuCl}$ ) and then transforms to paratacamite ( $\text{Cu}_2(\text{OH})_3\text{Cl}$ ) [118].

#### **4.7.4 Experimental Work Part I: Corroding the Coils Under Test**

The coils were corroded in order to establish the performance of a HF-IPT system as compared to one which was untainted. A custom built chamber (Fig. 4.19), was used to simulate a relevant marine-like corrosive environment. The coils were placed in slots and were wetted daily with brine, a 7 wt.%  $\text{NaCl}$  aqueous solution. This was designed to mimic a scenario where the coil was unintentionally splashed on occasion over the course of 30 days. The baseline final Q-factor of an untreated coil, subjected to the same analyses, was measured at 766.9 at 13.56 MHz. By contrast, the worst performing treated coil, described in further detail below, had a Q-factor of 614.1 at 13.56 MHz. It



Figure 4.19: Photograph of the gradually corroded coils in the corrosion chamber.

should be noted that ESR of high-Q coils is difficult to measure accurately [103]; however, these large differences in the measurements indicate that the coils under test (treated and untreated) have different conduction characteristics at the measured frequency.

#### 4.7.5 Experimental Work Part II: DC-DC HF-IPT Testing

An IPT test rig, consisting of Inverter II: Class EF, an IPT-link with two coils (Section 4.2.2) separated by a 15 cm air-gap, and Rectifier II: VM Class D, was designed and built. Fig. 4.20 shows the experimental setup.

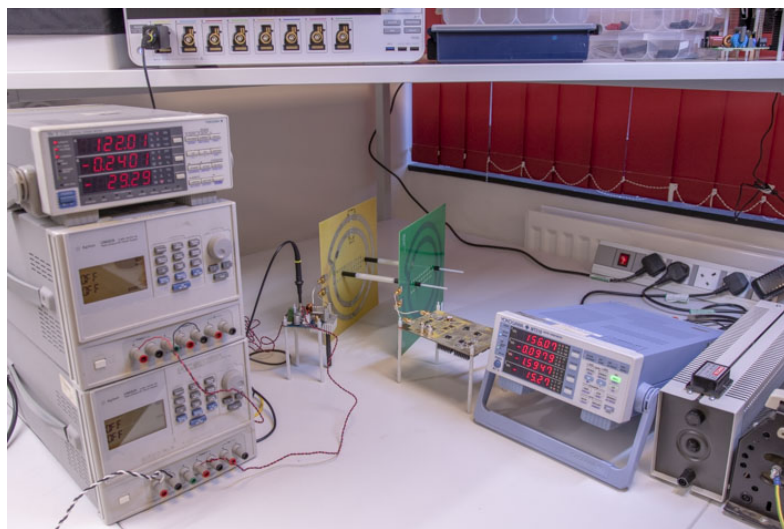


Figure 4.20: Photograph of the experimental test rig for inductive power transfer.

Since Inverter II was designed to be load independent, the coil under test can be placed at the transmit side, where it is expected to drive constant amplitude  $i_p$ . Therefore, at the transmit coil, constant losses are expected throughout the load spectrum. The inverter was set to drive the transmit coil under test with a 13.56 MHz 4.3 A (root mean square (RMS)) sinusoidal current, to induce an electromotive force on the receiver, and perform IPT.

This DC to DC HF-IPT system allows quantifying the losses across the entire system accurately, thereby providing a basis to report total losses in the system when alternating between an untreated and the treated coil at the transmit side.

#### **4.7.6 Experimental Work Part III: Materials Characterisation of the Coils Under Test**

At the conclusion of the experimental study, the conductive materials of the coils can be analysed to investigate the relationship between a reduced Q-factor (and consequently system efficiency) and material changes. The characterisation methods utilised to probe the conductive layer of the coils are X-ray diffraction and Scanning Electron Microscopy (SEM). Elemental analysis was established via Energy Dispersive X-ray (EDX) spectroscopy.

##### **X-ray Diffraction**

The X-ray diffraction (XRD) technique provides insight to the crystal structure of a material by probing with monochromatic X-rays and measuring the diffraction as a function of angle. Each of the peaks in the resulting diffraction pattern stands for a group of crystal planes. The characteristic interplanar spacing ( $d$ ) can be calculated using Bragg's law  $\lambda = 2d \sin(\theta)$ , where  $\lambda$  is the wavelength of the incident X-rays and  $\theta$  the angle between the incident beam and the scattering plane [119]. Each material has a characteristic diffraction pattern dictated by its crystal structure. Therefore, a diffraction pattern can act as a fingerprint to identify a material. The diffraction spectra presented herein were

obtained using a PANalytical Empyrean XRD with a wavelength of  $\lambda=0.15418$  nm. The patterns were collected using a Cu target at 40 kV and 20 mA, while data analysis was performed using the HighScore Plus software.

### **Scanning Electron Microscopy - Energy Dispersive X-ray Spectroscopy**

Scanning electron microscopy (SEM) provides information about the topography and elemental composition of the sample surface. Imaging can be achieved via the interaction of the surface with a high energy beam of electrons (20 keV). The result of such interactions can be inelastic or elastic leading to the generation of secondary or backscattered electrons, respectively [119]. By collecting and counting the electrons leaving the sample surface using appropriate detectors, an image is formed and recorded. Additionally, the elemental composition of the surface can be probed by collecting the X-rays produced when the inner shell electrons of atoms on the surface are excited. These emitted X-rays have a characteristic energy for each element, allowing for identification. The latter technique, termed energy dispersive X-ray spectroscopy (EDX), provides a valuable qualitative analysis of the surface elemental composition.

Imaging and elemental analysis of the PCB samples was performed using 6400 and 6010LA microscopes, both by JEOL, while the image and spectra collection and analysis was performed using INCA software. As non-conductive samples cannot be analysed with the SEM, due to charging, the samples were coated with chromium to enhance electron transfer and improve resolution.

#### **4.7.7 Experimental Results of a HF-IPT system Operating with a Corroded Coil**

For a 13.56 MHz current of 4.3 A, the losses in the coils were calculated to be 2.71 W for the untreated coil and 3.24 W for the treated one in accordance to the measurements of the ESR of the coils. Therefore, since the equivalent-series-resistance of the transmit

coil ( $R_p$ ) is the only variable that changes when the transmit coils are exchanged, the power losses are expected to be higher for the treated coil across the loading range.

Measurements of the total losses of the system (Table 4.6), demonstrate that the system using the treated coil is indeed less efficient than the system using the untreated coil over a broad range of  $R_o$ .

Table 4.6: Experimental results of DC to DC HF-IPT with a corroded and an uncorroded coil

| Coil condition | $R_o$ [ $\Omega$ ] | $P_i$ [W] | $P_o$ [W] | Total losses [W] | $\Delta$ Losses [W] | $\eta_{dc-o}$ |
|----------------|--------------------|-----------|-----------|------------------|---------------------|---------------|
| Untreated      | 1594               | 29.3      | 15.3      | 14.0             |                     | 52.2 %        |
| Treated        | 1591               | 30.1      | 15.2      | 14.9             | +0.9                | 50.4 %        |
| Untreated      | 913                | 39.5      | 25.4      | 14.1             |                     | 64.4 %        |
| Treated        | 913                | 39.9      | 25.1      | 14.8             | +0.7                | 62.8 %        |
| Untreated      | 427                | 65.6      | 50.1      | 15.5             |                     | 76.4 %        |
| Treated        | 426                | 66.1      | 49.5      | 16.6             | +1.1                | 75.8 %        |
| Untreated      | 182                | 96.3      | 73.0      | 23.3             |                     | 75.8 %        |
| Treated        | 183                | 95.7      | 72.1      | 23.6             | +0.3                | 75.4 %        |

In addition, thermal images (Fig. 4.21) of the treated and untreated coils were used to highlight discrepancies which may have arisen from the treatment process. The untreated

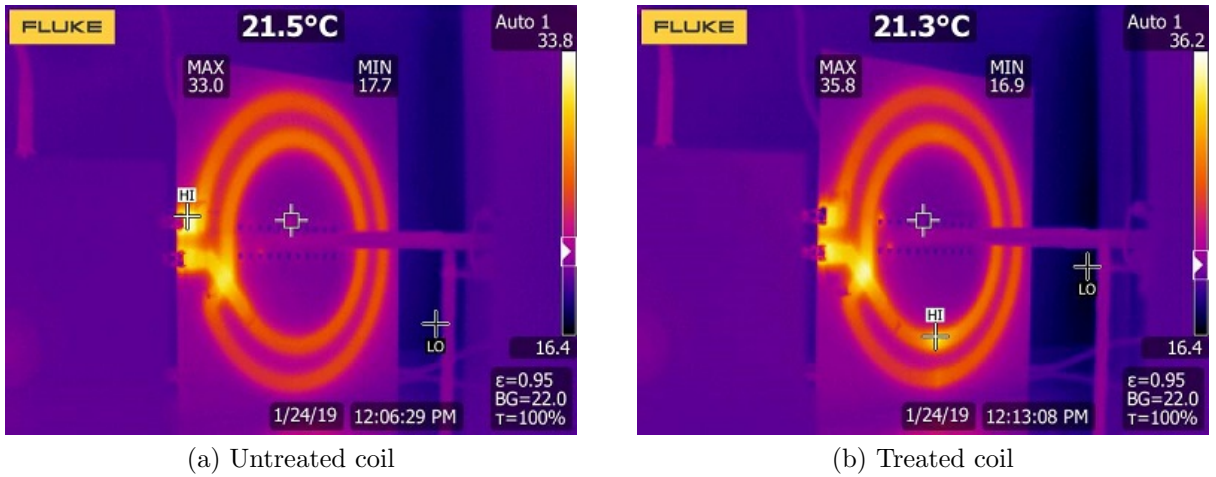


Figure 4.21: Thermal images of the coils under test driving 4.3 A at 13.56 MHz.

coil has a temperature maximum of 33.0°C at the point where the two turns cross and intersect with the inverter, when driving 4.3 A at 13.56 MHz for five minutes. By contrast, the treated coil has a maximum temperature in the section of the coil marked in Fig. 4.22.

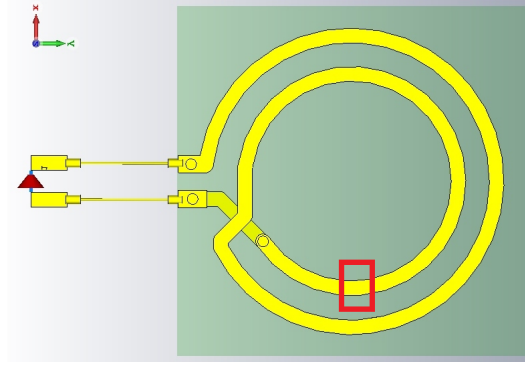


The maximum temperature on the treated coil was measured to be 35.8 °C, thereby also confirming that there is an increase in the  $R_{\text{sec}}$  of the treated coil.

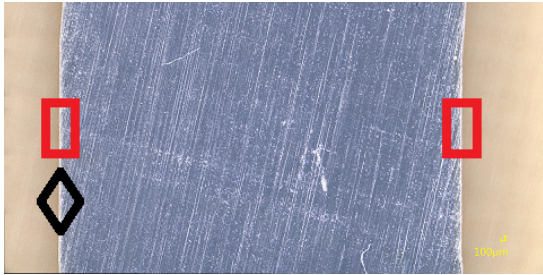
#### 4.7.8 Investigation of the Corrosion Mechanism of the Coils Under Test

The coils used in the HF-IPT experiments were imaged with optical microscopy and SEM. Investigation with XRD was conducted along with elemental analysis using EDX. Untreated coils were established to be uniform across the entirety of the surface. The low magnification optical microscope image (Fig. 4.22 (b)) demonstrates an example of this. However, this is not the case for the treated sample. The most exaggerated example of a treated coil is presented herein, where the thermal imaging (Fig. 4.21 (b)) indicated a maximum temperature inconsistent with the untreated sample. This location (Fig. 4.22 (a)), shows significant surface damage in the optical microscope image (Fig. 4.22 (c)). Whilst the original edges of the coil are apparent, the uniformity of the surface material no longer exists. Moreover, the colour changes to the bulk material are observed across the entirety of the coil. The two distinct regions, each starting at the edge and moving inwards, showed the most severely discoloured and apparently damaged surface observed across the entirety of this coil (Although, other instances of this process commencing could be speculated).

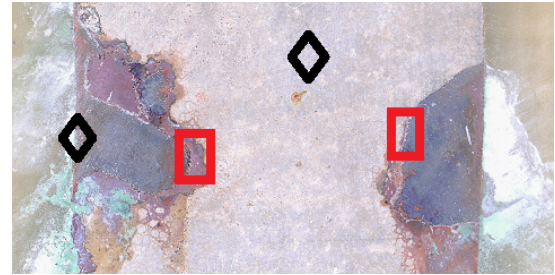
As a baseline, the topography of the untreated coil was obtained using depth profiling (in the red box regions highlighted in Fig. 4.22 (a,b)). The untreated coil has a thickness of *ca.* 65  $\mu\text{m}$ , above the surface of the plastic board (FR4) (Fig. 4.23 (a,b)). This thickness is even across all edges, as is evident from a three-dimensional rendering (Fig. 4.23 (c,d)). Changes to the topography on the damaged coil is evident from both depth profiling (Fig. 4.24 (a,b)) and a three-dimensional rendering (Fig. 4.24 (c,d)). From the original bulk material down to the region, which shows extraordinary damage, an overall reduction in height of *ca.* 30 - 35  $\mu\text{m}$  was recorded. This level of loss implicates not only the removal of the outer silver finish, but also exposure of the underlying copper.



(a) Coil diagram with an indication of the location under test

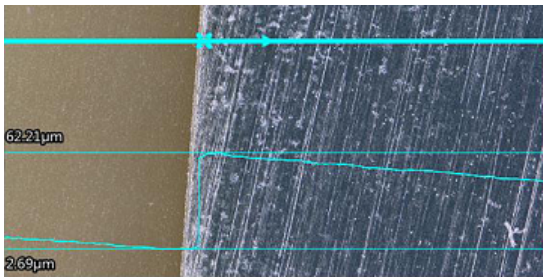


(b) Untreated coil, red boxes and black rhombuses indicate specific location of further optical and electron microscopy analysis respectively

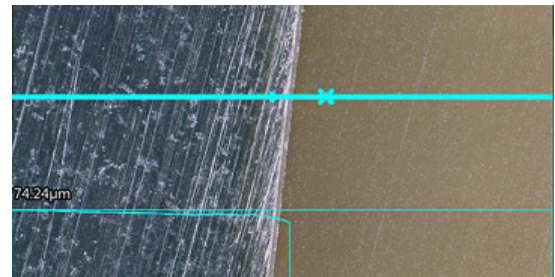


(c) Treated coil, red boxes and black rhombuses indicate specific location of further optical and electron microscopy analysis respectively

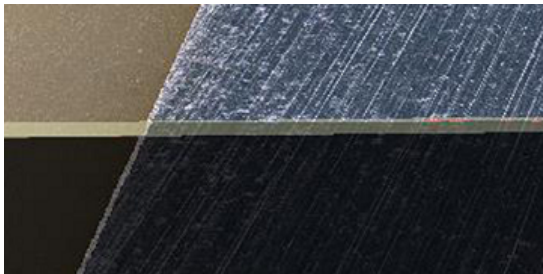
Figure 4.22: Images obtained with a digital microscope (100x).



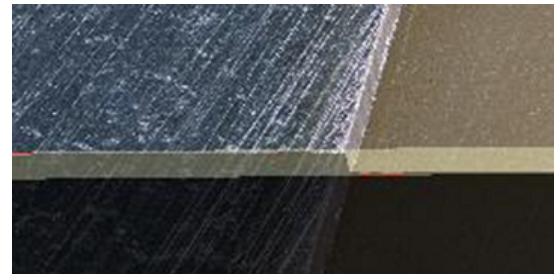
(a) Depth profile of the left edge



(b) Depth profile of the right edge



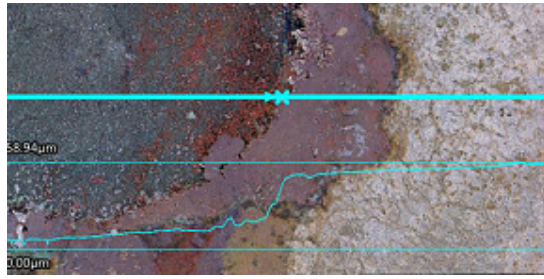
(c) 3D model of the left edge



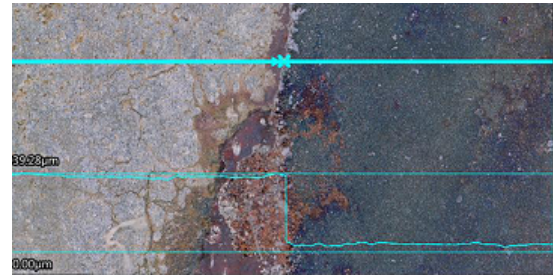
(d) 3D model of the right edge

Figure 4.23: Images of the untreated coil taken with a digital microscope at 500x.

A further assessment of this corrosion with electron microscopy images allows a closer inspection of the removal of the outer layers in the treated coil. For this purpose, the areas



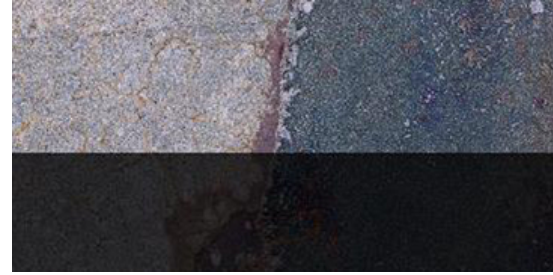
(a) Depth profile of the left edge



(b) Depth profile of the right edge

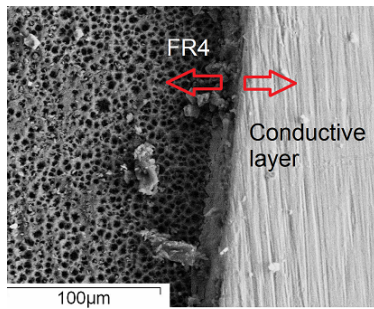


(c) 3D model of the left edge

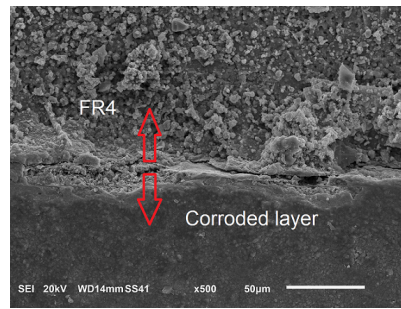


(d) 3D model of the right edge

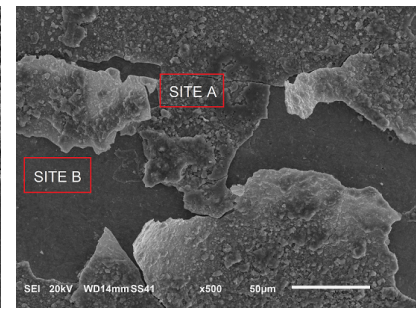
Figure 4.24: Images of the treated coil taken with a digital microscope at 500x.



(a) Untreated - conductive layer edge



(b) Treated - conductive layer edge



(c) Treated - copper trace edge

Figure 4.25: Electron microscopy images.

indicated with black rombi in Fig. 4.22 (b,c) were selected. Imaging of the untreated coil shows a distinct edge between the copper trace and the FR4 (Fig. 4.25 (a)). The edge between the FR4 substrate and the corroded material was probed (Fig. 4.25 (b)). Whilst the border is still evident, the nature of the corroded layer is dramatically different in appearance. A closer inspection of the centre of the coil is also considered, where the smooth layer in the untreated sample is no longer evident. Rather, there is evidence of poorly adhered flakes (Fig. 4.25 (c)). This change is observed across the coil surface, not just local to the area with extreme damage.

Using EDX, a semi-quantitative assessment of the superficial elemental composition of

the coil was made. To ensure the analysis is accurate, multiple measurements were done on each sample and the average values, including the standard deviation are presented in Table 4.7. As a basis of comparison, the untreated coil indicates an elemental composition consistent with the coil fabrication: silver coating on copper is detected (Table 4.7). Whilst the total content of the coil is higher in copper, this is an artefact of the technique. As such, the detection of some copper is a result of the interaction volume of the electron beam and the sample surface, thereby probing not only the surface but some material underneath. Two specific points on the corroded regions were targeted for elemental analysis. On a poorly adhered flake (site A on Fig. 4.25 (c)), only traces of silver remain. The composition is primarily copper and oxygen, with traces of chlorine. This suggests that a copper oxide corrosion product has been formed on the surface. On the underlying material (site B on Fig. 4.25 (c)), below the loosely adhered flakes, 36 and 52 at. % of silver and copper are measured, respectively. Furthermore, oxygen content, at 10 at. %, is markedly lower than on the flakes. Additionally, there is a continuing presence of chlorine across this surface as well.

Table 4.7: Average chemical composition (at.%) of the treated and untreated conductive layer

| Element  | Treated        |                | Untreated      |
|----------|----------------|----------------|----------------|
|          | Site A         | Site B         |                |
| Silver   | $0.1 \pm 0.0$  | $36.2 \pm 4.0$ | $43.3 \pm 1.0$ |
| Copper   | $33.0 \pm 5.6$ | $52.4 \pm 4.3$ | $56.7 \pm 1.0$ |
| Chloride | $0.5 \pm 0.0$  | $1.4 \pm 0.0$  | $0.0 \pm 0.0$  |
| Oxygen   | $66.4 \pm 5.4$ | $10.0 \pm 1.2$ | $0.0 \pm 0.0$  |

Small distinctions between the treated and untreated samples can be noted from the diffraction patterns (Fig.4.26). Whilst notable peaks for copper,  $44^\circ$ ,  $50^\circ$  and  $74^\circ$   $2\theta$ , are consistent, the less intense peaks at lower  $2\theta$  values are of specific interest. The small peaks at  $36^\circ$  and  $38^\circ$  may indicate that  $\text{Cu}_2(\text{OH})_3\text{Cl}$  or  $\text{Cu}_2\text{O}$  is formed on the surface [120], consistent with the previous elemental analysis. Additionally, changes to the nature of the copper crystallinity are suggested as a function of changing peak ratios amongst the three most prominent peaks. Notably, whilst the bulk material changes are not immediately

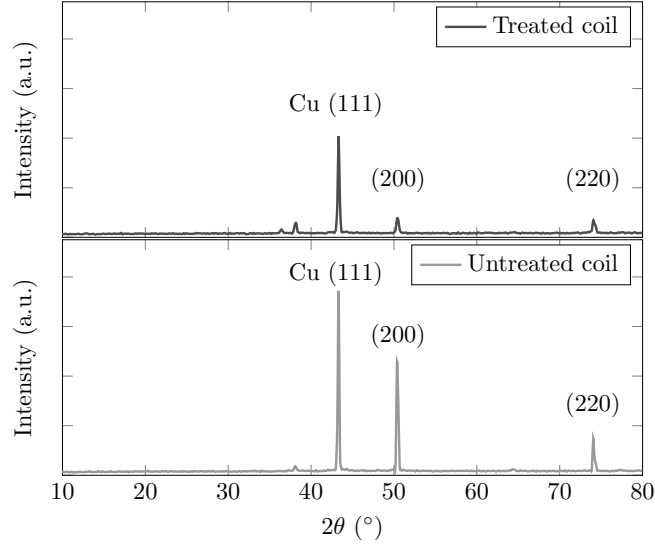


Figure 4.26: X-ray diffraction patterns of the untreated and treated coils.

responsible for reduced performance of the coil, it likely provides a foundation upon which more significant, and impactful, corrosion can occur.

#### 4.7.9 Conclusion

Impedance measurements and thermal images indicate that the treated coil, with the lowest Q-factor, has an additional source of losses at a local region of the copper trace. This subsequently compromises the overall conduction path of the coil. Observations of this part of the trace using optical microscopy indicate that the trace was heavily corroded, thereby significantly reducing the conduction path. As such, the local corrosion damage is attributed with a 20 % decrease in Q-factor on the treated coil, as compared to the untreated. Notably, the oxide layer and damage across the surface more widely is not implicated in these performance losses. As such, this work demonstrates that corrosion can significantly influence the Q-factor of IPT coils, and consequently the end to end efficiency of a HF-IPT system.

## 4.8 Summary

This chapter presents work in the characterisation of HF-IPT systems in scenarios where practical aspects, such as detuning, and system deterioration are taken into account in the design of the transmitter and receiver, which can be investigated independently in accordance to the formulation proposed for the reflected impedance on the transmit side and the induced electromotive force on the receive coil described in Chapter 2.

The design and construction of a HF-IPT test load, and the methodology to tune it precisely using the impedance analyser was a fundamental element in characterising the effects of different reflected impedances on the transmit side. The characterisation of an inductive link with air-core coils allowed performing diverse but related experiments. First, the inverters were tested against detuned loads; second, the effects of introducing three alternative media between the coils instead of air were analysed. A reflected impedance estimation technique using the transmit side waveforms was investigated, and finally, the effects of corrosion on air-core coils were experimentally verified.

The research presented in this chapter has been exploratory in nature, and opens up possibilities of future research projects. The experiments help better understand fundamental principles of HF-IPT concerning different possible situations that may arise as the systems are integrated into commercial products.



# Chapter 5

## Simplified HF-IPT Systems

### AC-Mains Interface

The content in this chapter focuses on original work involving feeding a high frequency inductive power transfer (HF-IPT) system with low frequency (50 or 60 Hz) alternating current (AC), emulating a mains voltage, instead of the traditional approach of relying on a direct current (DC) source to feed the system. Part of the work discussed in this chapter has been published in an IEEE Conference Proceedings, [121] and the work involving driving the proposed Class E based inverters from an AC source has been filed as an international patent [122]. The proposed system configuration allows one power electronic stage to be eliminated, thus increasing the end-to-end system efficiency.

#### 5.1 Introduction

Most of the electronic appliances which require a DC voltage source but are powered or charged from the mains, require power factor correction (PFC) circuitry so that the nonlinear load comprised of a mains-rectification stage and a DC load appears to be resistive. Emulating a resistive load, in principle, is useful because when the power factor is unity (i.e. the load is purely resistive) the highest output power is achieved for a given

line total root mean square (RMS) current, minimising the line conduction losses and maximising the utilisation of the feeding conductor. The power factor can be calculated as:

$$PF = \frac{P_{\text{avg}}}{V_{\text{rms}} I_{\text{rms}}}, \quad (5.1)$$

where  $P_{\text{avg}}$  is the real power (i.e. the average power over one period of time) and  $V_{\text{rms}}$  and  $I_{\text{rms}}$  are the RMS voltage and current respectively. Since the voltage and current are periodic signals, they can be represented as a Fourier series, thus allowing the real power to be calculated from the voltages and currents that operate at the same frequency (otherwise the average power is zero) and at the same phase (if the signals are expressed as the sum of two signals, one that is in phase and one that is  $90^\circ$  out of phase).

In practical terms, if the voltage is assumed an ideal sine-wave, the power factor is calculated as:

$$PF = \frac{\cos(\theta)}{\sqrt{1 + THD^2}}, \quad (5.2)$$

where  $\theta$  is the phase angle between the fundamental voltage and current, and the current's total harmonic distortion (THD) is defined as:

$$THD = \frac{\sqrt{\sum_{n=2}^{\infty} I_n^2}}{I_1}, \quad (5.3)$$

where  $I_1$  is the current at the fundamental frequency in RMS and  $I_n$  is the current at each harmonic frequency in RMS. The THD that a nonlinear load is allowed to generate is regulated by standards (e.g. [123]), for which passive and active current shaping circuits are employed, typically at the front end of the converter connected to the power supply. Most commonly, switched-mode power supplies using topologies based on the boost converter [124] are used to achieve a resistance emulation given that the boost-converter is a current-source DC-DC converter, which is a convenient feature for input current shaping.

The traditional design theory of high frequency inverters assumes an ideal DC voltage



source (e.g. [125]), and proposes the design of the passives, switching frequency and duty cycle accordingly. A typical HF-IPT system, which has a high frequency inverter at the transmit side, is also commonly designed from an ideal DC voltage source. Similarly, at the receive side a DC output voltage is expected. Therefore, for such type of designs, a front-end circuit that performs PFC and regulates the output voltage which feeds the inverter is required.

In the following sections of this chapter, an alternative HF-IPT system design which takes advantage of some of the inherent properties of the HF-IPT circuits introduced in Chapter 3 is proposed. In this system the power conversion stage from mains AC is integrated in the design of the inductive power transfer (IPT) system with the purpose of simplification and convenience for integration into commercial applications, and since one power processing stage is removed, the new system configuration can potentially achieve higher efficiencies.

## 5.2 Power Factor Correction and Power Throughput Control at the Receiver

Fig. 5.1 shows a block diagram of the main power conversion stages for HF-IPT, assuming the coupling factor ( $k$ ) to be constant. If the transfer function of the three stages illustrated in the diagram in Fig. 5.1 are accounted as constant gains, the amplitude of



Figure 5.1: DC to DC HF-IPT system diagram with constant  $k$ .

the transmit-coil current ( $i_p$ ) depends strictly on the DC input voltage ( $V_{dc}$ ) and the voltage-to-current gain of the inverter, the amplitude of the induced electromotive force on the receive coil  $\mathcal{E}_s$  depends on  $i_p$  and the current-to-voltage gain of the IPT-link, and the IPT-rectifier output voltage ( $V_o$ ) depends on  $\mathcal{E}_s$  and the voltage gain of the rectifier.

The end-to-end HF-IPT system could therefore be represented as a single gain, corresponding to the product of the gains of the inverter, the IPT-link and the rectifier, and as such, the voltage waveform at the output is expected to be a scaled version of the input voltage. Conversely, the impedance seen at the input of the system can also be modelled as the product of the load impedance and the inverse of the system's gain, if the end-to-end efficiency is assumed unity.

If at the transmit side of a typical HF-IPT system, the mains connection is considered instead of assuming an ideal DC source, the mains-to-DC power conversion can be represented as in Fig. 5.2, where a single stage accounts for the power conversion of the HF-IPT system. Interestingly, in principle, the system diagram shown in Fig. 5.2 can be



Figure 5.2: Mains to DC simplified typical HF-IPT system diagram.

reorganised as proposed in Fig. 5.3 where  $V_{\text{rect}}$  and  $V_o$ , instead of being purely DC, are shaped as the absolute value of a low frequency sine-wave. The functionality of a HF-IPT



Figure 5.3: Mains to DC simplified HF-IPT system diagram with the PFC circuitry on the receiver.

system employing the system architecture proposed in Fig. 5.2 and the one in Fig. 5.3 should be equivalent, however the advantage of the latter is that the system's output voltage can be regulated by the PFC stage allowing combining the PFC capabilities of that stage with an output voltage regulation stage (shown in the system architecture of Fig. 5.4), which is commonly implemented in this type of system.

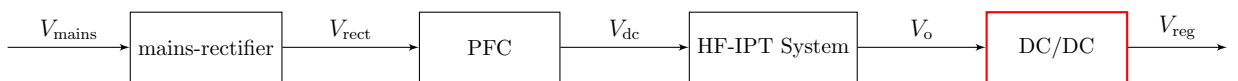


Figure 5.4: Mains to DC simplified typical HF-IPT system diagram with a regulated output.

Performing PFC at the receive side of an IPT system was first proposed in [126]. That system was designed to operate at 20 kHz, and be capable of supplying 1 kW to the load. The main purpose of feeding the inverter from a rectified mains voltage, as opposed to a regulated DC source, was to reduce the size of the input DC link capacitance of the inverter, consequently improving the power density and reliability of the system, as it is well known that large capacitor banks tend to significantly compromise the reliability of a power converter [127]. One of the challenges addressed in [126] was therefore to operate the inverter from a fluctuating input voltage, shaped as the absolute value of a single-phase AC voltage source, and perform PFC at the receive side with a boost-like current shaping circuit. Improvements of the type of system first proposed in [126] can be found in the development of bidirectional wireless power transfer (WPT) systems in [58, 128]. As expected of a system that is fed from a rectified AC voltage, it is noted in [129] that feeding an inverter from this type of source results in reaching peak voltages and currents that are higher than those required for systems that operate in DC.

Mains to DC converters are not commonly accounted for in the design of HF-IPT systems since such solutions have mostly been proposed for low power applications. However, with efficient higher power ( $>1$  kW) HF-IPT systems demonstrated [130, 131], solutions integrating the mains to DC power conversion stage become more relevant. An example of such concern has been addressed for example in [132], where a 100 W mains-connected 6.78 MHz full-bridge inverter combined with a totem-pole rectifier was proposed. In that work the two power conversion stages are unified, hence reducing the number of components and allowing to drive an IPT transmit coil with the inverter connected directly from the mains, however the system architecture used is a traditional one which requires the additional power conversion stage for output voltage regulation highlighted in Fig. 5.4.

### 5.3 Description of the Proposed System Architecture

The main reason to operate a HF-IPT system with the PFC stage at the receive side is different to that mentioned for systems that operate at kilohertz frequencies, mostly because the current-driven resonant inverters (i.e. Class E and Class EF) proposed for HF-IPT have a large input inductance as part of the topology, thus not necessarily requiring a large DC link capacitance. However, this technique allows the system to be simplified by reducing the number of power conversion stages, hence reducing size, cost and losses associated with that power conversion stage.

A typical system architecture with a DC-DC converter after the rectifier for power regulation is shown in Fig. 5.5 (a). This system architecture includes a PFC stage at the transmit side which is not always accounted for but is conventionally required to produce a DC voltage supply from the mains. The voltage regulation feature of the PFC stage is not fully exploited since the output voltage regulation is done by the DC-DC converter at the receive side. The system can be simplified by either eliminating the DC-DC converter at the receive side, for which a control loop from the system output voltage to the PFC stage would be required, or by eliminating the PFC stage from the transmit side, and perform PFC at the receive side DC-DC converter. The latter is the proposed system architecture (Fig. 5.5(b)).

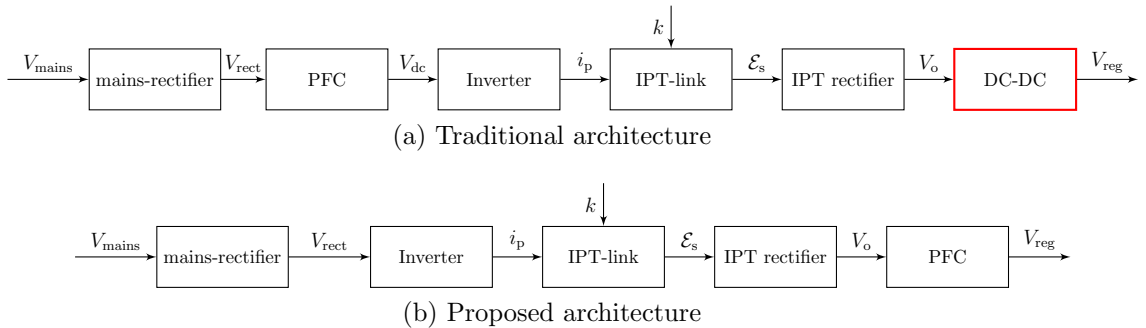


Figure 5.5: Mains to DC typical and proposed HF-IPT system architecture.

The power conversion stages of the system are further illustrated with the transmit side

waveforms in Fig. 5.6, and the receive side waveforms in Fig. 5.7.

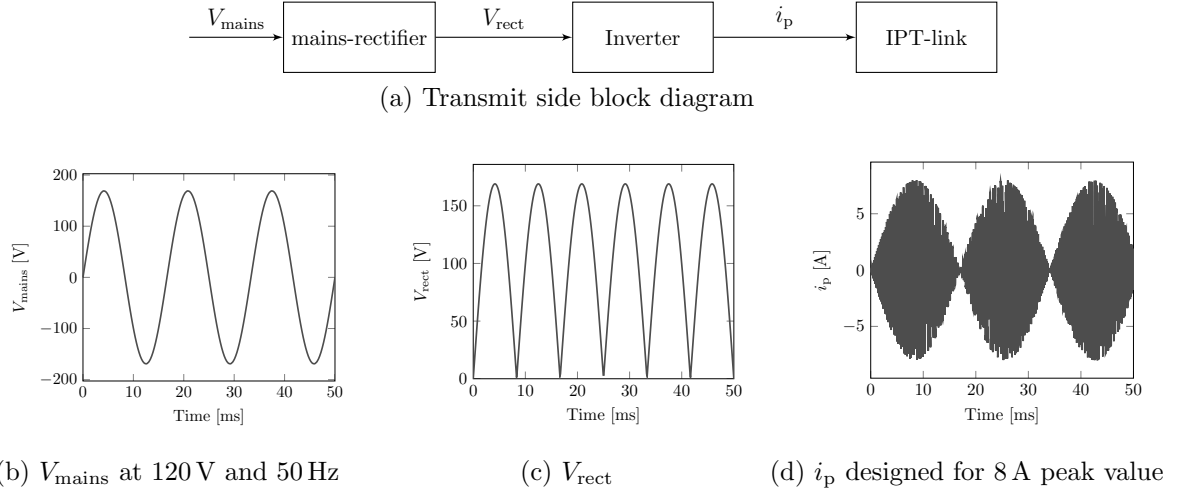


Figure 5.6: Theoretical waveforms at the transmit side of a HF-IPT system using the proposed system architecture.

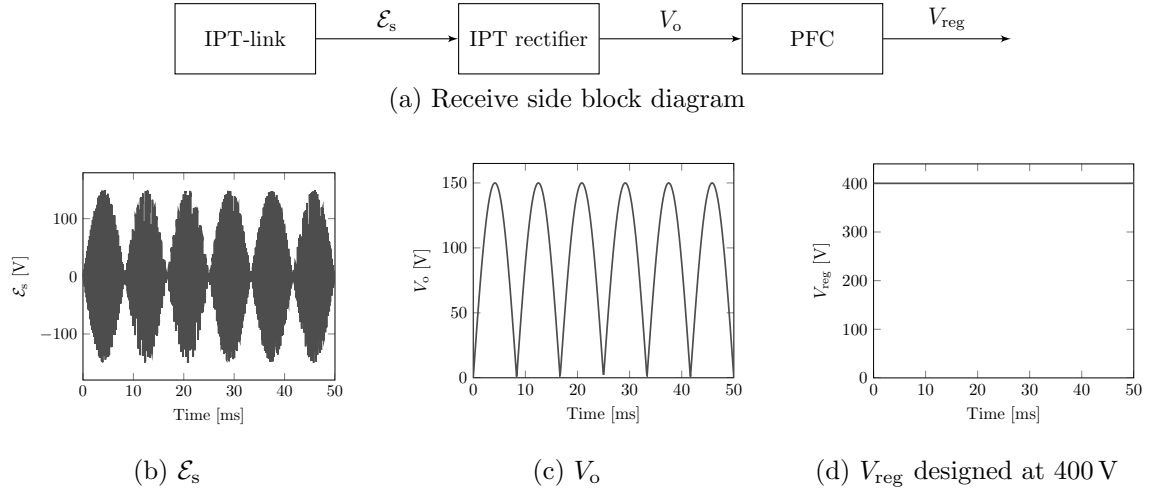


Figure 5.7: Theoretical waveforms at the receive side of a HF-IPT system using the proposed system architecture.

The proposed system architecture for HF-IPT systems operates as follows: a single-phase mains voltage ( $V_{\text{mains}}$ ) source is rectified by the mains-rectifier in Fig. 5.6 (a), with theoretical voltage waveforms shown in Fig. 5.6 (b),(c). The output voltage of this stage,  $V_{\text{rect}}$ , is shaped as the absolute value of  $V_{\text{mains}}$  minus the voltage drop on the diodes (Fig. 5.6 (c)). This source is then connected to a Class E or a Class EF inverter. The amplitude of the output current on the inverter depends of  $V_{\text{rect}}$ ; therefore, the transmit coil is driven by a high frequency (the frequency of operation of the inverter) output current ( $i_p$ ), with a modulated amplitude determined by  $V_{\text{rect}}$  (Fig. 5.6 (d)). The

switching frequency and duty cycle of the inverter are kept constant. The current on the transmit coil ( $i_p$ ) produces an induced electromotive force ( $\mathcal{E}_s$ ) on the receive coil, as specified in (2.7), and shown in Fig. 5.7 (b). An IPT-rectifier (this block also includes the capacitance typically used to resonate the receive coil) is then integrated to rectify the high frequency  $\mathcal{E}_s$  and filter out the high frequency component, but not the lower (twice the mains) frequency. The voltage waveform of  $V_o$  is therefore DC with a modulated amplitude determined by  $V_{\text{rect}}$  (Fig. 5.7 (c)) and the gain of the rectifier. The final power conversion stage is a DC-DC converter, which not only shapes the system's input current ( $i_{\text{mains}}$ ) to emulate a resistive load, but can also regulate the output voltage ( $V_{\text{reg}}$ ). A single closed-loop switch-mode power supply at the receive side of the system is utilised for output voltage regulation (i.e. power throughput control) and PFC.

Since the PFC stage emulates a resistive load, with a constant resistance regardless of the instantaneous changes on  $V_o$ , a constant resistance is also reflected back to the mains power source. Fig. 5.8 illustrates the voltage and current waveforms with the same shape and phase henceforth emulating a constant resistance.

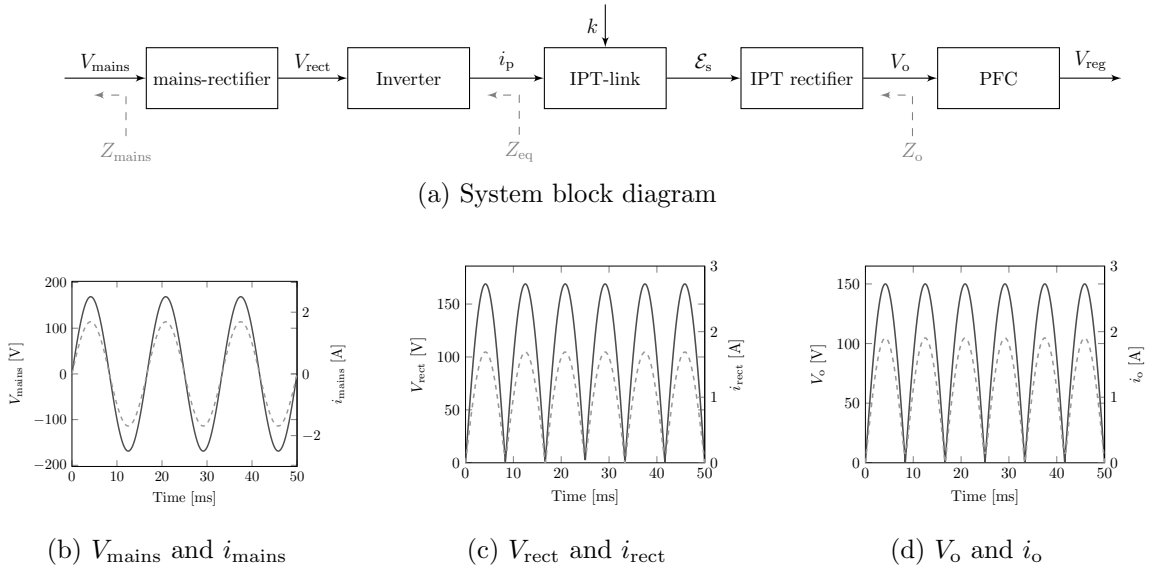


Figure 5.8: Theoretical waveforms of voltages and currents at key stages of the system. The current is plotted as a dotted line.

As in any PFC stage, as the load resistance changes, the input resistance of the converter also changes. However, since the current shaping control of this type of system relies on

the shape of its input voltage, the current, regardless of its average value, should follow the form and phase of the voltage (Fig. 5.9). Similarly, in an IPT system with constant

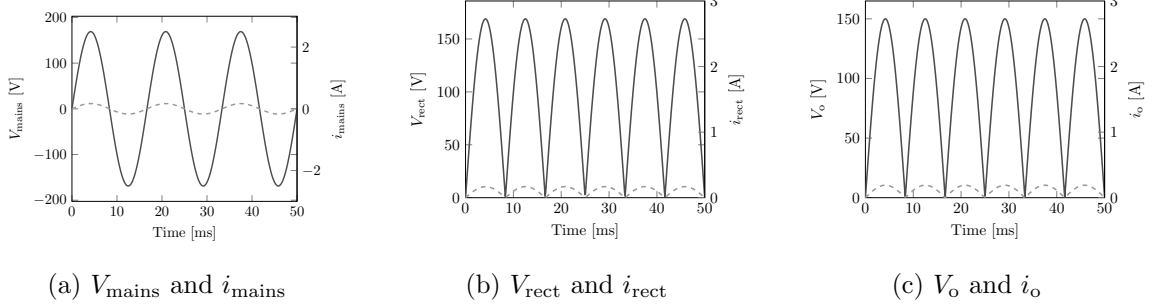


Figure 5.9: Theoretical waveforms of voltages and currents at key stages of the system driving a lighter load. The current is plotted as a dotted line.

$i_p$ , when  $k$  changes, the amplitude of  $V_o$  changes accordingly, and as long as these changes are within the admissible range of the PFC stage, it will draw more (or less) current from  $V_o$  to maintain the constant IPT system regulated output voltage ( $V_{\text{reg}}$ ), as illustrated in Fig. 5.10.

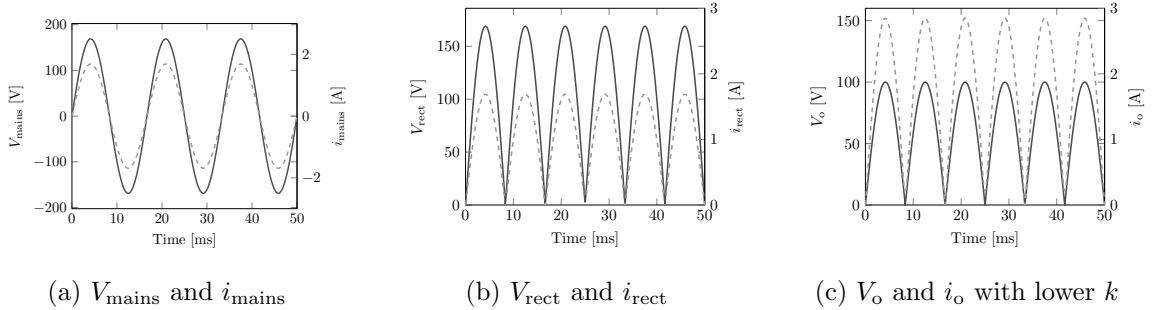


Figure 5.10: Theoretical waveforms of voltages and currents at key stages of the system with lower  $k$ . The current is plotted as a dotted line.

## 5.4 Experimental Verification

A set of experiments were performed to verify the practicality of the proposed system architecture. A 150 W 13.56 MHz HF-IPT system was built, using the proposed inverters and rectifiers of Chapter 3, and the printed circuit board (PCB) coils of Chapter 4.

The test setup (Fig. 5.11) and the equipment used for these experiments are specified in Section 3.7.

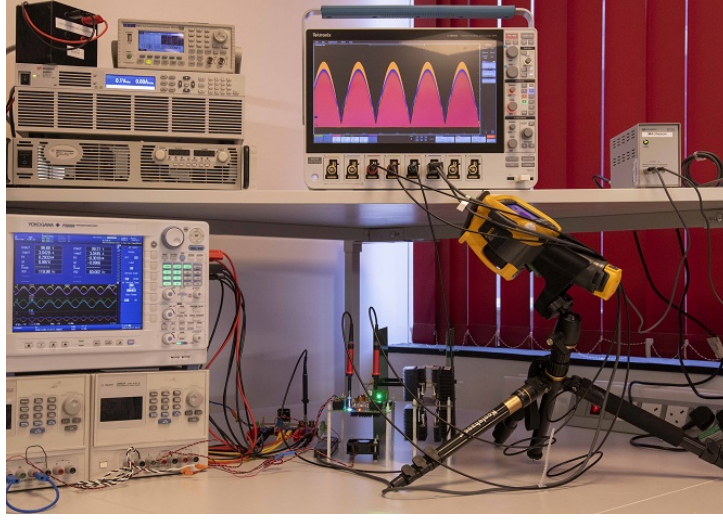


Figure 5.11: 60 Hz AC to regulated DC HF-IPT systems experimental setup.

The experiments presented herein are organised as follows: first, the three candidate inverters proposed in Chapter 3 were powered from a rectified 60 Hz voltage source, and these were tested powering the inductively coupled test load developed in Section 4.3. A change in the reflected load was verified. Afterwards, a HF-IPT receiver was fed from the transmitter instead of the test load, first to verify that the PFC stage at the receiver performs PFC at the mains source, and afterwards to demonstrate the proposed system architecture operating from the proposed current-source inverters.

#### 5.4.1 Case Study I: Driving Class E and Class EF Inverters from a 60 Hz Sinusoidal Input Voltage

The inverters built and described in Chapter 3 were tested at different DC input voltages (Fig. 3.24) in order to verify the correct operation throughout the voltage range, and the inverters experimental voltage-to-current gain. Inverter I: Class E and Inverter II: Class EF were tuned considering a maximum voltage of 170 V ( $\approx 120\sqrt{2}$  V); however their voltage-to-current gain is slightly different. Fig. 5.12 shows the respective measurements



of the amplitude of  $i_p$  in function of  $V_{dc}$ . The gain of Inverter I: Class E is  $0.0447 \text{ A V}^{-1}$ , and the gain of Inverter II: Class EF is  $0.0506 \text{ A V}^{-1}$ .

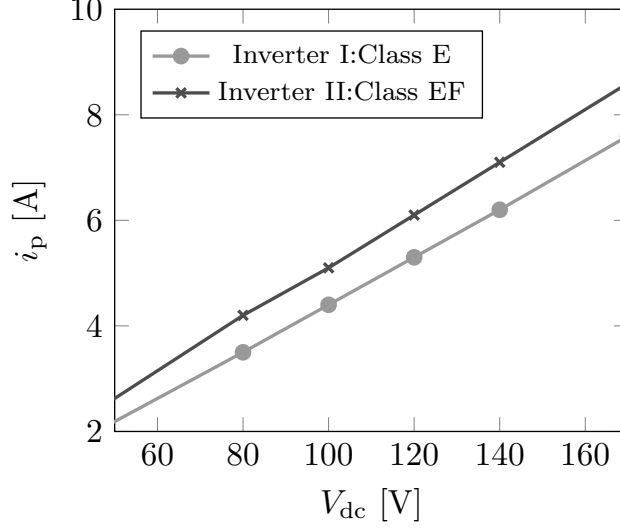


Figure 5.12: Experimental verification of the output current amplitude as a function of input voltage of inverters I and II.

It should be noted that in the circuit topologies proposed for this system architecture, there is a capacitance (Denoted as  $C_1$  in Fig. 3.3 and 3.7) in parallel with the transistor, and since the output capacitance of a transistor ( $C_{oss}$ ) is voltage dependent, as the input voltage changes, the tuning might be compromised. This effect, however, was taken into account when the inverters were designed (Chapter 3), by proposing solutions where  $C_1$  is much larger than  $C_{oss}$ , and the constant inverters gain (Fig. 5.12) demonstrates that the inverters can be fed from a broad range of input voltage, as is the case in the proposed system architecture.

The inverters were powered from an AC power source at 120 V and 60 Hz, which is the nominal single-phase commercial and residential mains voltage in countries like Costa Rica or the United States. This voltage specification was selected, as opposed to 240 V and 50 Hz, which is the nominal mains voltage in the UK, since the devices voltage ratings would have been surpassed with these topologies. The circuit topologies of Inverter I: Class E and Inverter II: Class EF driven from a low-frequency AC source are shown in

Fig. 5.13.

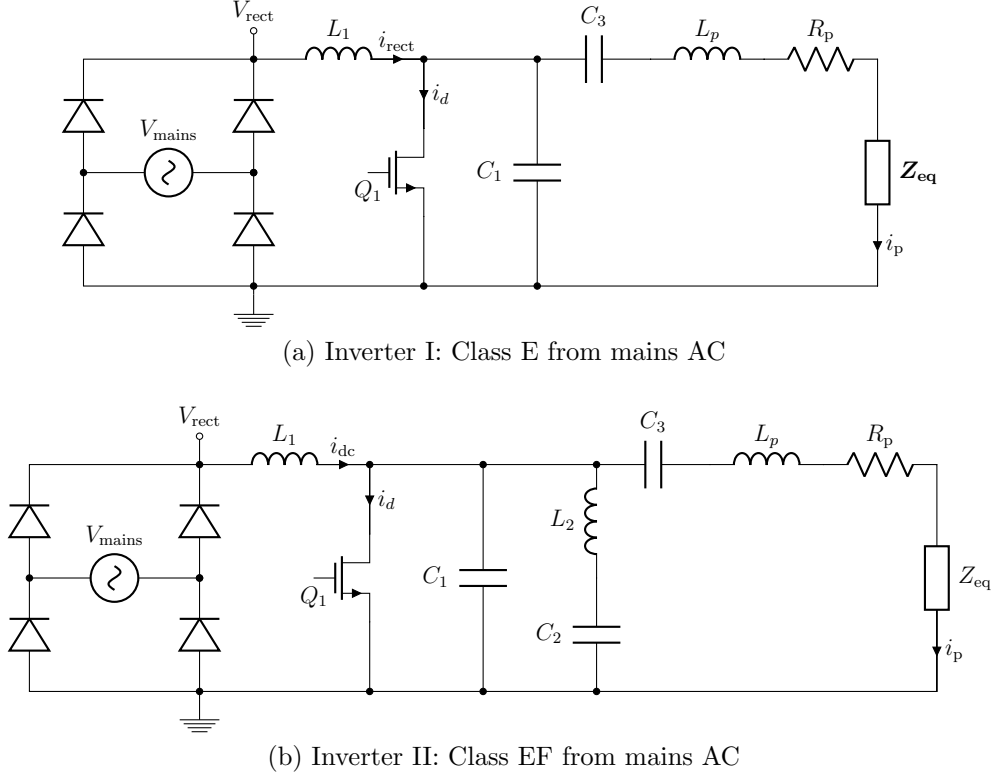


Figure 5.13: Circuit topologies of Inverters I and II from mains AC.

Oscilloscope traces verify that the shape of rectified-mains voltage ( $V_{\text{rect}}$ ) and  $i_{\text{rect}}$  are illustrated in Fig. 5.14. These traces were captured for an input power of 150 W, and the power factor in both cases was  $>0.995$ . The coils were aligned at a separation of 11 cm for Inverter I and 12 cm for Inverter II. This to achieve the same power level using the test-load developed in Section 4.3. The difference in coupling for the two cases is due to Inverter I having a lower voltage-to-current gain than Inverter II (Fig. 5.12), and since both inverters are powered from the same power supply, they require different reflected loads to reach the same power throughput. The only detectable difference in the waveforms of the system operating from these two candidate inverters is that the drain-voltage of Inverter I is higher than the one of Inverter II, which is expected.

These experiments demonstrate that a resistive-tuned reflected impedance on the transmit side ( $Z_{\text{eq}}$ ) in a HF-IPT system using the current-source inverters proposed in Chapter 3 achieve unity power factor when powered from a rectified low frequency AC voltage.

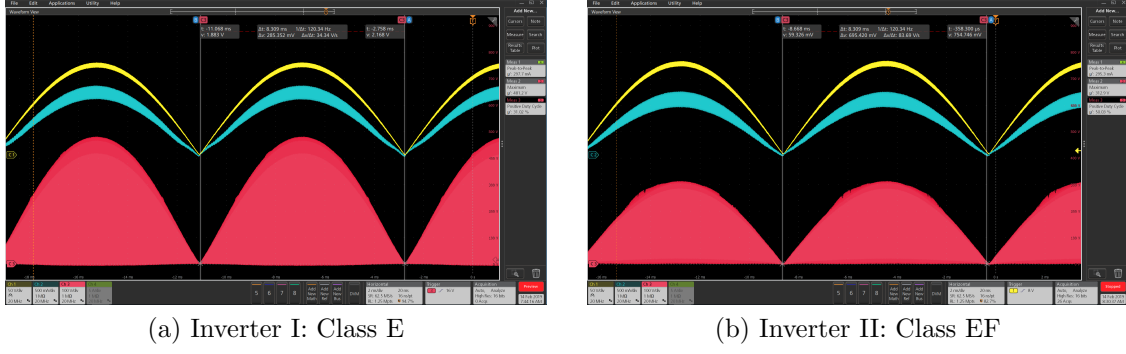


Figure 5.14: Oscilloscope Traces of  $V_{\text{rect}}$  (Channel 1),  $i_{\text{rect}}$  (Channel 2) and  $V_{ds}$  (Channel 3).

### 5.4.2 Case Study II: Changing the Reflected Load to a Push-Pull Class EF Inverter When Powered From a 60 Hz Sinusoidal Input Voltage

Inverter III: Push-Pull Class EF was also configured to be powered from a low frequency AC source. The circuit diagram is shown in Fig. 5.15.

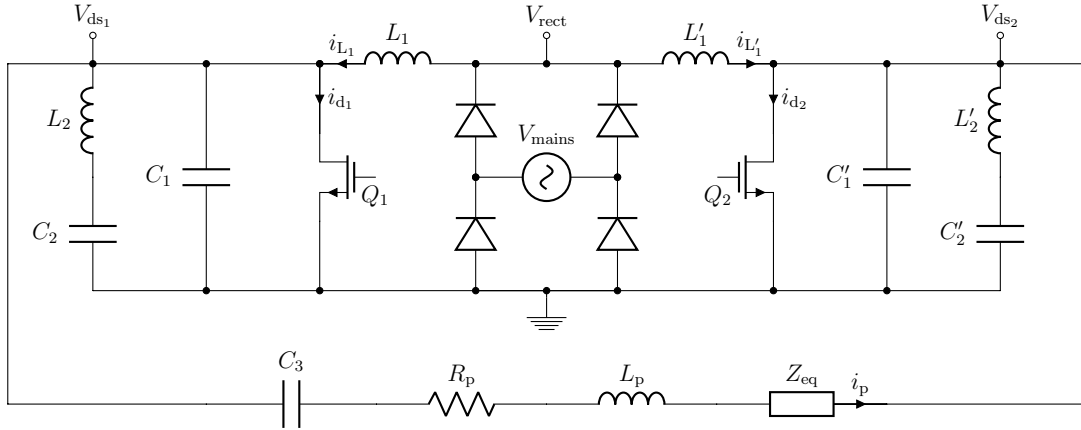


Figure 5.15: Circuit topology of Inverter III from mains AC.

This design has a significantly higher voltage-to-current gain than the one of Inverters I and II. The gain of this inverter was measured at  $0.098 \text{ A V}^{-1}$  as is shown in Fig. 5.16. The higher gain feature, which is inherent of the push-pull topology, allows driving the coil with higher current amplitude, thus making it more suitable for higher power applications. For this reason, it was also built using higher rated transistors, as specified in Table 3.4.

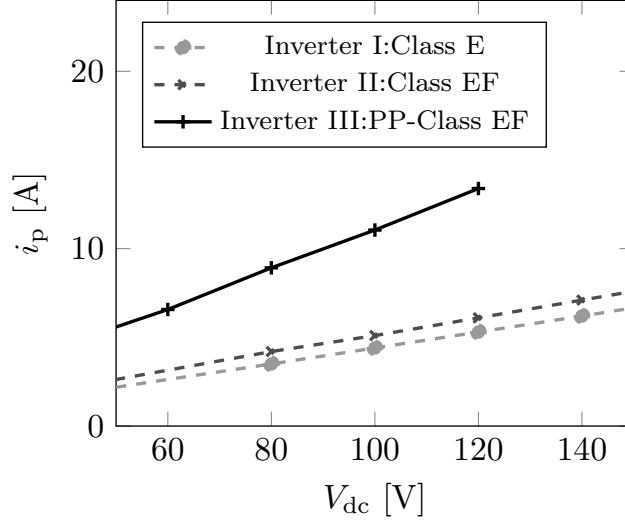


Figure 5.16: Experimental verification of the output current amplitude as a function of input voltage of Inverter III.

For these experiments, the inverter was powered from a 100 V 60 Hz voltage source as opposed to the typical mains voltage of 120 V used in the experiments using Inverters I and II. Fig. 5.17 shows the experimental setup using Inverter III: Push-Pull Class EF and the HF-IPT test load developed in Section 4.3.

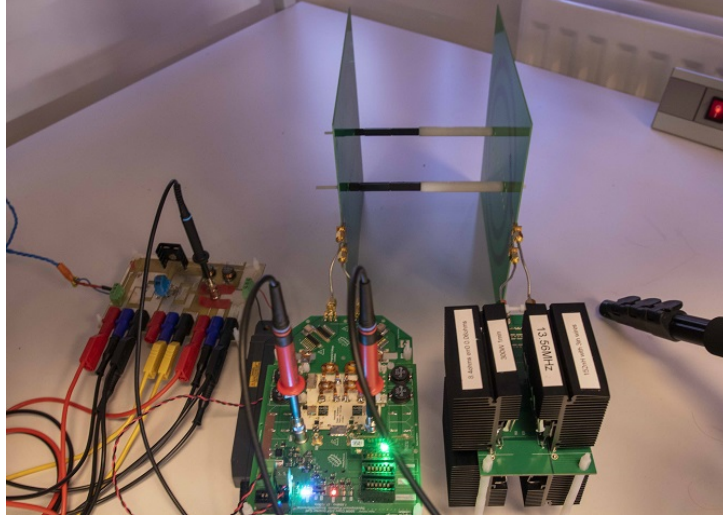


Figure 5.17: Experimental setup using Inverter III: Push-Pull Class EF from a low frequency AC voltage source.

After demonstrating that this type of inverter is also suitable to be powered from a low frequency AC voltage source, the test load coil position was stepped to change the

reflected resistance on the transmit coil ( $R_{eq}$ ) and verify that as long as it is maintained resistive, the system should achieve unity power factor. Fig. 5.18 shows the waveforms of  $V_{rect}$  and  $i_{rect}$  with different  $R_{eq}$  corresponding to different coupling values with the test load.

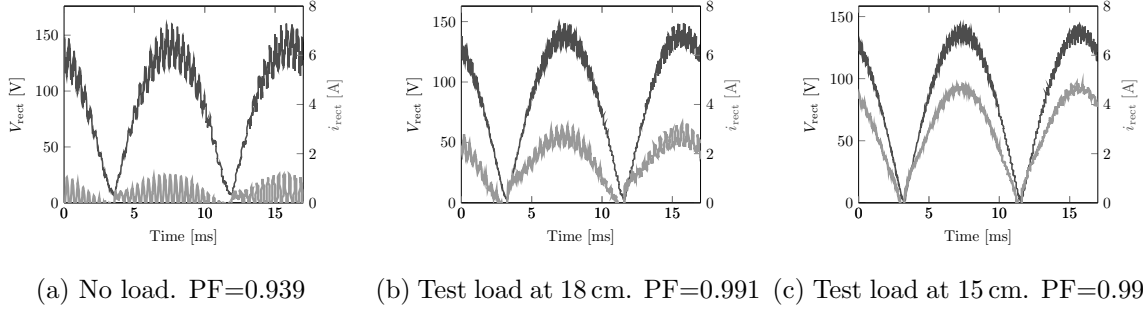


Figure 5.18: Experimental waveforms of Inverter III powered from a 60 Hz voltage source.

When the system was unloaded, the standby power consumption was approximately 40 W, and the power factor 0.939. As the transmitter was loaded, first with the load at 18 cm distance and afterwards at 15 cm, the power factor improved to 0.997. At this point, the system's input power was measured at 293 W. It should be noted that  $Z_{eq}$  for these experiments is much lower than in the experiments of the previous case study due to the higher gain of the inverter. However, unity power factor is also achieved for the lower  $Z_{eq}$  of this case study.

### 5.4.3 Case Study III: Operating the Proposed Architecture Using a Large DC-Link Capacitor After the IPT-Rectifier Instead of the PFC Stage

Once it was demonstrated that unity power factor is achieved when the receiver is a resistive load, a receiver with a rectifier was integrated. The rectifier in these experiments is Rectifier I: RCN Class D, described in Section 3.5.

Fig. 5.19 (a) shows a photograph of the experimental setup for this experiment and Fig. 5.19 (b) shows a detail of the PCB of the rectifier, with the addition of three 100  $\mu$ H

electrolytic capacitors in parallel with the output capacitance of the rectifier to filter out not only the 13.56 MHz signal but also, to an extent, the 120 Hz from  $V_{\text{rect}}$ . In that case, instead of being resistive,  $\mathbf{Z}_{\text{eq}}$  depends of  $V_{\text{rect}}$ , and thus is non-linear.

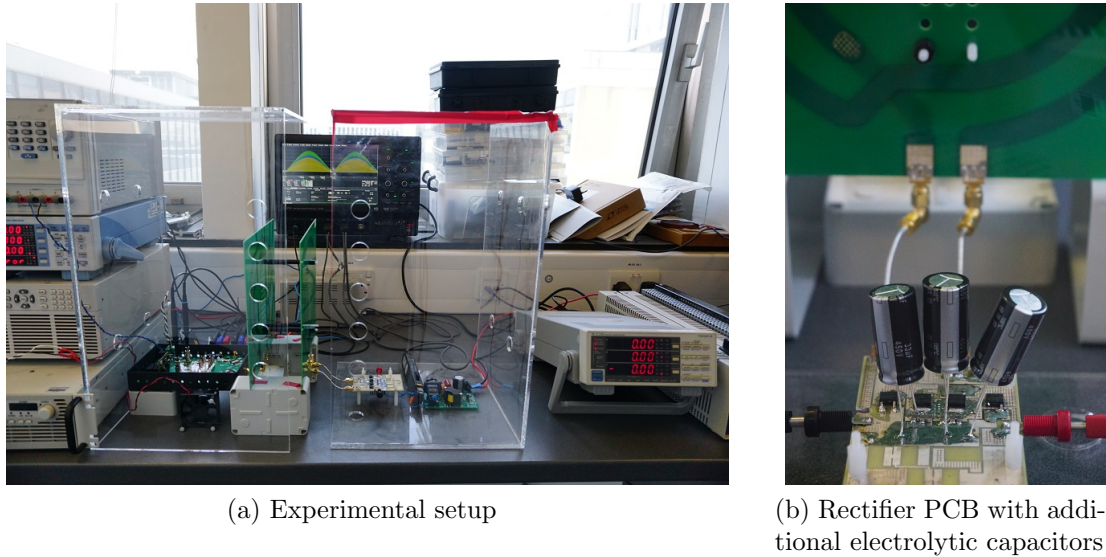


Figure 5.19: Experimental setup to reflect a non-linear load to the transmitter.

The first experiment consisted of powering the system without modifying Rectifier I, with the designed output capacitance of 1  $\mu\text{H}$  (Table 3.5), which would filter out the high frequencies of the inverter but not the low frequencies of the mains voltage. After the rectifier, the PFC circuit, also specified in Section 3.5, which emulates a resistive load, hence achieving a constant  $\mathbf{Z}_{\text{eq}}$  was integrated. The system operated similarly as the one verified in the first case study. The circuit waveforms are shown in Fig. 5.20 (a). A second experiment was conducted where three electrolytic capacitors of 100  $\mu\text{H}$  (Fig. 5.19 (b)) were included in parallel with the designed output capacitance of the rectifier (of 1  $\mu\text{H}$ ) to verify that once  $\mathbf{Z}_{\text{eq}}$  is non linear (the added capacitors make it behave as an impedance that depends of  $V_{\text{rect}}$ ), unity power factor is no longer achieved. The waveforms of this second experiment are shown in Fig. 5.20 (b).

These two experiments were conducted to demonstrate the necessity of having a resistance emulation circuit after the rectifier in order to achieve unity power factor, similarly as it is required in a traditional PFC circuitry from the mains.

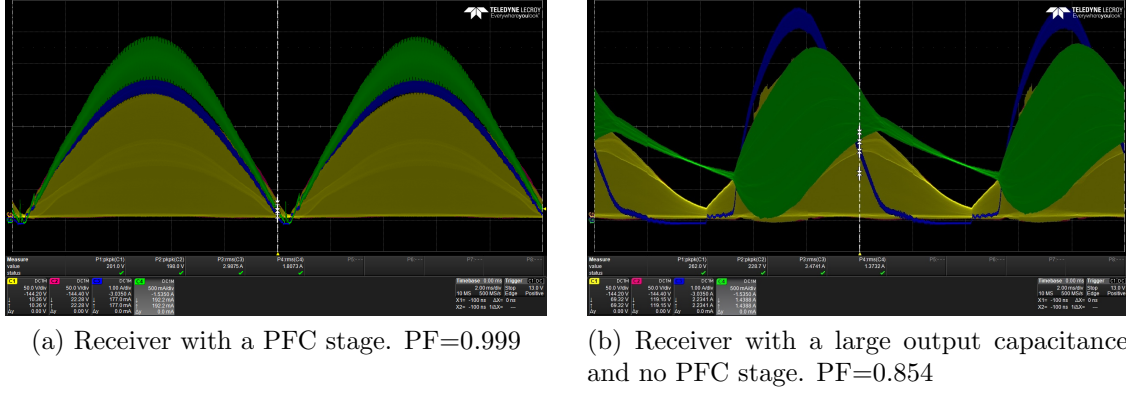


Figure 5.20: Oscilloscope traces of  $V_{ds}$  (Channels 1, 2),  $i_o$  (Channel 3) and  $i_{rect}$  (Channel 4).

#### 5.4.4 Case Study IV: Driving the Inverters from a 60 Hz Input Voltage and Feeding the Load with a Regulated DC Output Voltage

A 150 W end-to-end system using the proposed system architecture was built. The circuits which were implemented were Inverter III: PP Class EF and Rectifier II: VM Class D. The additional boards (the PFC stage and the mains-rectifier) are detailed in Section 3.7.3.

The system was tested at different air-gaps from 15 cm to 12 cm corresponding to a coupling range of 3 to 8 % and different loads. At lower coupling ( $<4\%$ ), the system could only power up loads lower than 100 W, since as the load was increased, the PFC stage stopped regulating the output voltage at 400 V. The minimum coupling at which the system operated for the entire load range was at 5.3 %, when the coil separation was set at 13 cm. The results and waveforms of the system operating at this coil separation are shown in Fig. 5.21.

The load was stepped from 50 to 150 W in order to assess the system at different loading conditions. At full load, the system reached an end-to-end efficiency of 73.3 % and the power factor was measured at 0.993.

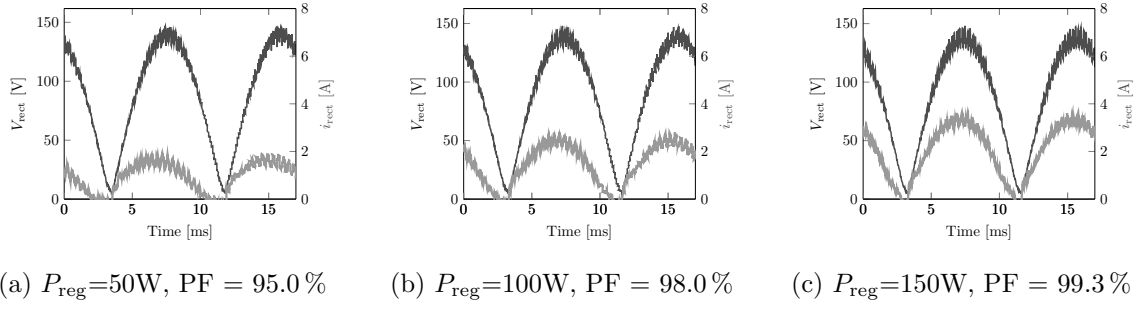


Figure 5.21: Experimental waveforms of the input voltage and current of the inverter when fed from an 60 Hz AC source, and loaded with the proposed rectifier followed by a PFC stage at the receiver and a DC load.

## 5.5 Summary

This chapter proposes an alternative HF-IPT system architecture feeding the Class E or Class EF inverter directly from the mains voltage, where in order to achieve unity power factor, a PFC stage was included at the receive side to also regulate the output voltage, taking the place of the regular DC-DC converter that is often present on the receiver. The proposed architecture was tested using the three proposed current source inverters, taking advantage of the inherently continuous input current of these single switch resonant inverters, which help filter out high frequencies at the source. A PFC stage based on the boost topology at the receive side, after the rectifier, was used to shape the input current of the system in order to achieve unity power factor.

The operation of an end-to-end prototype using the proposed system architecture was demonstrated experimentally at 13.56 MHz and 150 W. Results show an end-to-end efficiency of 73.2 % and a power factor of 0.993 when powering a 150 W load.

The system architecture proposed in this chapter was also implemented in Chapter 7, where the detailed design of a wirelessly charged electric scooter is described from the mains to the battery.



# Chapter 6

## Analysis and Design for Dynamic HF-IPT Systems

The content in this chapter focuses on original work involving characterisation of high frequency inductive power transfer (HF-IPT) systems with movement between the transmit and receive sides using probability theory. Part of the work discussed in this chapter has been published in IEEE Transactions on Power Electronics: [53] and an IEEE Conference Proceedings: [133].

### 6.1 Introduction

Near-field wireless power transfer (WPT) technology can potentially change how users interact with electronic devices. In applications as broad as medical implants (charged at several microwatts up to a few watts [134, 135, 136]) up to electric vehicles and trains (requiring from a few kilowatts to megawatts [34, 137]) wireless charging is beneficial, and in applications, such as autonomous vehicles and robots, it becomes truly disruptive.

The size and configuration of the energised zones in which the devices can be charged and how freely these can move within the boundaries of the energised zones, are fundamental

design challenges concerning the user experience with the technology, which is essential for its commercial success.

There are different design approaches for inductive power transfer (IPT) depending on the application. For example, solutions for charging electric vehicles [57], where the system relies on specific and precise link properties, and thus are not designed to allow for significant tolerance to misalignment or movement, can achieve very high end-to-end efficiencies. In contrast, systems can be designed for broader ranges in coupling factor ( $k$ ) allowing tolerance to misalignment or movement, for example in dynamic electric vehicle charging [56, 138], where the design slightly compromises operating at the peak efficiency point all the time to exploit the spatial freedom that the technology provides. Making the system more robust with respect to misalignment and tuning has been an important topic of research in IPT [139, 140].

In HF-IPT with air core coils the technical challenges of allowing for coil misalignment and movement are different than those found for kilohertz IPT in applications such as electric vehicle charging, because not only are the circuits used to drive the coils different, but also the ability to shape the flux is only dependent on the shape of the electrical conductor that forms the coil.

In this chapter, the characterisation of a dynamic charging environment is proposed to allow optimising the system using probability theory, and therefore account for typical desirable features that a dynamic inductive charger must have. The model not only takes into account the distribution of  $k$  throughout the charging space, but also represents the movement of the device being charged as a probability distribution, thus allowing the formulation to be applicable to systems in which the movement is either predictable or unpredictable. The model can be useful for two main design considerations. First, to identify at which range of  $k$  the circuits at both ends can operate and at which  $k$  they should achieve the peak efficiency so that the energy-efficiency throughout the charging time is maximised. And second, the model gives a general expression which can be used to evaluate the performance of an IPT system as  $k$  changes, and therefore despite the system

tolerating a range of  $k$ , it could be suggested, based on the movement characterisation of the device, to bound it to a more restrictive range of  $k$ .

## 6.2 HF-IPT Systems with Variable Coupling

As the coils of an IPT system move,  $k$ , and the induced voltages on the mutually coupled inductances, change. As discussed in previous chapters, and notwithstanding that there are techniques that allow regulating power at the inverter or the rectifier, the HF-IPT systems proposed in this thesis consider the inverter and the rectifier as open loop power converters with constant gains. The inverter operates with constant duty cycle and frequency, and the amplitude of transmit-coil current ( $i_p$ ) is linearly dependent on the DC input voltage ( $V_{dc}$ ). Similarly, the amplitude of the non-synchronous IPT-rectifier output voltage ( $V_o$ ) can be formulated as linearly dependent of the induced electromotive force on the receive coil ( $\mathcal{E}_s$ ). Fig. 6.1 illustrates a block diagram that accounts for changes in  $k$ , where the link is formulated as in (2.7).  $\mathcal{E}_s$  depends on the product of  $i_p$  and  $k$ . Interestingly, with this block diagram  $k$  can be conceived as a modulating signal.

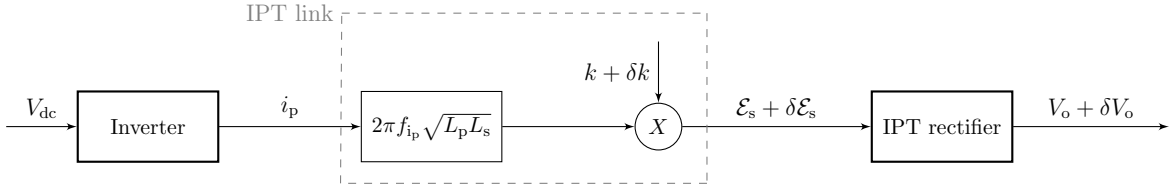


Figure 6.1: Block diagram of a HF-IPT system with variable coupling.

The formulation illustrated in Fig. 6.1 allows the inverter and the rectifier to be designed for a range of  $k$ , and considering abrupt disturbances in  $k$  within its range.

As mentioned in Chapter 3, a controller can be included before the inverter or after the rectifier to regulate the power throughput against changes in  $k$ . However, in addition to system-level power throughput control, the tolerance to variable  $k$  needs to be addressed in the design of the inverter and the rectifier.

The optimal load as expressed in (2.17) depends of  $k$ . Thus, if the link is designed so that the input impedance of the rectifier is equal to the optimal load, a power converter after the rectifier would be required to emulate the  $k$ -dependent optimal load. Since the power converter after the rectifier is in charge of emulating the optimal load, an additional DC-DC converter at the transmit side would be required to change the amplitude of  $i_p$  and in this way deliver the required output power. This system operation, which was tried in [141], requires two power converters, one at each end of the system. Therefore the additional losses introduced by these stages should be lower than the losses introduced in the link when the coils are aligned at a point of non-optimal  $k$ , for it to be reasonable to implement over the simpler solution of using a single power converter at the receive side for power throughput control and operating the transmitter at constant  $i_p$  amplitude.

Notwithstanding that solutions as the one in [141] might be reasonable in some cases, investigating the variability profile of  $k$  for a given link is necessary to assess this option in comparison to a system that tolerates operating with the input impedance of the rectifier different than the optimal load.

### 6.2.1 Formulating the HF-IPT System End-to-End Efficiency, Input Power and Output Power

In order to account for the effects produced by changing  $k$  on a HF-IPT system, the formulation for the input power, output power and the IPT system DC to DC efficiency ( $\eta_{dc-o}$ ) need to be formulated. First,  $\eta_{dc-o}$  can be calculated as the product of the different stages of the system, which depends on how the stages are separated:

$$\begin{aligned}\eta_{dc-o} &= \eta_{inv}\eta_{link}\eta_{rect} \\ &= \eta_{dc-\mathcal{E}}\eta_{\mathcal{E}-o}.\end{aligned}\tag{6.1}$$

The IPT link efficiency ( $\eta_{link}$ ) considers the losses in the coils, and  $\eta_{inv}$  and  $\eta_{rect}$  are the efficiencies of the inverter and the rectifier respectively. Alternatively, losses can

be distributed with the IPT transmitter DC to AC efficiency ( $\eta_{\text{dc}-\mathcal{E}}$ ) which accounts for the losses in the inverter and the transmit coil, and the IPT receiver AC to DC efficiency ( $\eta_{\mathcal{E}-\text{o}}$ ) which considers the losses in the receiving coil and the rectifier.

The output power can be calculated at the output of the rectifier as:

$$P_o = \frac{V_o^2}{R_o}, \quad (6.2)$$

but also at the transmit side (Fig. 6.2) as:

$$P_o = \frac{i_p^2 R_{\text{eq}} \eta_{\mathcal{E}-\text{o}}}{2}, \quad (6.3)$$

where  $R_{\text{eq}}$  is the real part of the reflected impedance on the transmit side ( $\mathbf{Z}_{\text{eq}}$ ).

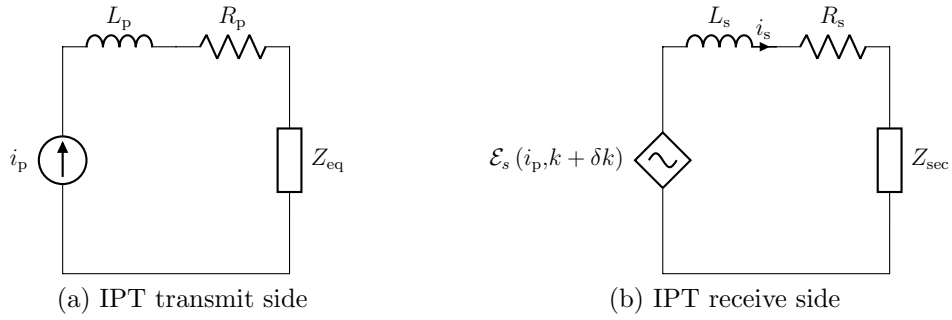


Figure 6.2: General equivalent circuits of an IPT link with a reflected impedance considering changes in  $k$ .

Similarly, the input power can be calculated at the input of the inverter as:

$$P_i = V_{\text{dc}} i_{\text{dc}}, \quad (6.4)$$

or at the receive side as:

$$P_i = \frac{i_s^2 (R_{\text{sec}} + R_s)}{2\eta_{\text{dc}-\mathcal{E}}}, \quad (6.5)$$

where  $R_{\text{sec}}$  is the real part of input impedance of the IPT-rectifier ( $\mathbf{Z}_{\text{sec}}$ ), and  $R_s$  is the equivalent series resistance (ESR) of the receive coil. Many other formulations are

possible taking into account the circuit model in Fig. 6.2 and the expressions in (6.1).

### 6.2.2 Considering Time in the Formulation of the System Efficiency

When  $k$  is considered variable with respect to time, the system can be optimised calculating not the instantaneous power efficiency but the energy-efficiency in a period of time. The energy delivered to the load for the charging time of  $t_0$  to  $t_f$ , is defined by integrating power throughout the charging time:

$$E_o \Big|_{t_0}^{t_f} = \int_{t_0}^{t_f} P_o (R_{eq}, i_p, \eta_{\mathcal{E}-o}) dt, \quad (6.6)$$

and since  $R_{eq}$ , depends on  $k$  and  $R_{sec}$ :

$$E_o \Big|_{t_0}^{t_f} = \int_{t_0}^{t_f} P_o (k, R_{sec}, i_p, \eta_{\mathcal{E}-o}) dt. \quad (6.7)$$

A similar expression can be found for the energy extracted from the DC source,

$$E_i \Big|_{t_0}^{t_f} = \int_{t_0}^{t_f} P_i (k, R_{sec}, i_p, \eta_{dc-o}) dt, \quad (6.8)$$

to calculate the energy-efficiency as:

$$\eta_{dc-o_{avg}} = \frac{E_o}{E_i}. \quad (6.9)$$

The DC-DC energy-efficiency is equivalent to the weighted-average DC-DC power-efficiency of the system throughout the charging-time.

If the variables that define  $P_o$  and  $P_i$  do not change with time,  $E_o$  and  $E_i$  are calculated as the product of power and charging time. However, if  $k$  changes with time, it will affect how  $P_o$  and  $P_i$  are distributed and  $E_o$  and  $E_i$  should be calculated using the general expressions of (6.7) and (6.8) respectively. The variables, other than  $k$ , that define  $P_o$

might also change to compensate for the changes in  $k$  which would also affect how  $P_o$  and  $P_i$  are distributed in time. For example, the inverter can be designed to achieve constant amplitude in  $i_p$  or alternatively with a control loop, it can be designed to adapt the amplitude of  $i_p$  as  $k$  changes. In the latter case, in the formulation of (6.7) and (6.8),  $i_p$  can be written as a function of  $k$ . Similarly,  $R_{sec}$  can be modified with a power converter after the rectifier which would, for a series-tuned IPT system, decrease the value of  $R_{sec}$  when  $k$  decreases, in which case it can also be written as a function of  $k$ . The other parameters that define  $P_o$ , such as the inductance on the coils ( $L_p$ ,  $L_s$ ) are considered constants. It should be noted that (6.2.1) considers tuned circuits; hence, the input reactance of the rectifier ( $X_{sec}$ ) is assumed to be zero at the frequency of operation. However, the losses produced by changes in  $X_{sec}$  can be included in the model, as showed in the second case study of this chapter.

### 6.2.3 Characterising the Coupling Distribution in Space

In dynamic environments where  $k$  changes as the relative position of the coils change it is key to characterise  $k$  as a function of the geometry of the link considering the possible relative positions of the coils for a charging environment. Several mathematical models to characterise  $k$  as a function of distance and misalignment can be found in the literature for different coil designs, for example [101] can be used for circular coils and [142] for planar coils.

An alternative, introduced in this work, is to describe  $k$  as a probability density function or a probability mass function. These distributions can be obtained by curve fitting data from simulations, measurements or analytical expressions.

### 6.2.4 Characterising the Coupling Distribution in Time

The probability distribution of the relative position of the coils,  $g(x_1, \dots, x_n)$ , can be defined from a motion profile as a general expression for a defined charging time as:

$$\int_{x_{1\min}}^{x_{1\max}} \dots \int_{x_{n\min}}^{x_{n\max}} g(x_1, \dots, x_n) dx_1 \dots dx_n = 1, \quad (6.10)$$

where,  $x_1, \dots, x_n$  are the variables that determine the relative position of the coils, and consequently affect  $k$  (e.g. linear misalignment, distance, angular misalignment),  $n$  represents the number of variables taken into account and  $x_{n\min}$  to  $x_{n\max}$  represents the range considered for the variable  $x_n$ .

The probability-density-function of (6.10), which maps the probability distribution of the position of the coils throughout the geometry of the link, contains part of the information required to obtain the expected value of  $k$  for a period of charging time. The expected value of  $k$  is a useful parameter to know because it can help determine the values or ranges of  $k$  which are more likely to take place, and thus have a higher effect on the energy-efficiency. In order to formulate a probability distribution which not only accounts for the distribution of the position of the coils but also the distribution of  $k$  in space, a probability-density-function  $f(k)$  is proposed.  $f(k)$  is obtained by performing a probability-density-function transformation of the probability-density-function of the position of the coils with respect to time (determined by motion) with the probability-distribution of  $k$  in space. The first case study (Section 6.3.2) shows an example of how  $f(k)$  is obtained from the coupling distribution in space, and a known movement profile.

Interestingly,  $f(k)$  can be obtained not only for predictable movement patterns, as the one in the first case study in this chapter (Section 6.3), but also for unpredictable movement patterns, as the one in the second case study (Section 6.4). Moreover, if the relative position and the misalignment are predictable for the entire charging-time,  $k$  can be described as a function of time, and the energy-efficiency could be calculated directly from (6.9) in the time domain. This approach, however, does not consider spatial autonomy,



where it is often impossible to predict the exact position and misalignment of the coils, and therefore describe  $k$  as a function of time. Therefore, treating the motion of the coils as a probability distribution deals with the unpredictability of  $k$  by treating it as a stochastic variable.

The expressions in (6.7) and (6.8), assuming that  $R_{\text{sec}}$ ,  $i_p$ ,  $\eta_{\mathcal{E}-o}$  and  $\eta_{\text{dc}-o}$  are functions of  $k$  (or constant), can be reformulated as the product of charging time and the expected value of  $P_o$  and  $P_i$ :

$$E_o \Big|_{t_0}^{t_f} = (t_f - t_0) \int_{k_{\min}}^{k_{\max}} P_o(k) f(k) dk, \quad (6.11)$$

$$E_i \Big|_{t_0}^{t_f} = (t_f - t_0) \int_{k_{\min}}^{k_{\max}} P_i(k) f(k) dk. \quad (6.12)$$

The energy-efficiency can be therefore calculated independent of the charging time:

$$\eta_{\text{dc}-o_{\text{avg}}} = \frac{\int_{k_{\min}}^{k_{\max}} P_o(k) f(k) dk}{\int_{k_{\min}}^{k_{\max}} P_i(k) f(k) dk}. \quad (6.13)$$

This formulation allows to design and evaluate the system at different values of  $k$ , where each value is weighted by its probability using the probability-density-function  $f(k)$ .

### 6.2.5 Optimising the Link for Variable Coupling Environments

The proposed energy-efficiency formulation using  $f(k)$  points out key parameters that can be optimised for energy-efficiency. First, the range of  $k$  at which the system transfers power can be selected not only based on the capabilities of the system but on the evaluation of the performance at different values of  $k$  and considering how probable these cases are. Second, the coils can be designed to achieve desirable features in  $f(k)$  for a particular motion profile. For example, the shape of the conductors of the coils can be

optimised for maximum  $k$  at the most probable position, or a shape can be found where the change of  $k$  is minimised for the given movement profile. And third, the model can help determine at which  $k$  it is best to optimise the inverter, the rectifier and the link.

### 6.3 Case Study I: Selecting the Alignment Range of Charging for a System With a Known Motion Profile and Without Power Throughput Control

A 150 W HF-IPT system using Inverter II: Class EF and Rectifier I: RCN Class D was mounted on a linear slide to change the alignment of the coils and emulate a HF-IPT charger for a load moving across the charging pad at constant velocity, maintaining the same separation as the coils move. The experiments in this case study were performed with copper-pipe coils with 178 mm diameter with similar inductance to the printed circuit board (PCB) coils proposed in Chapter 4. Fig. 6.3 shows a photograph of the experimental setup and Fig. 6.4 shows photographs of the system with the coils at different alignments.

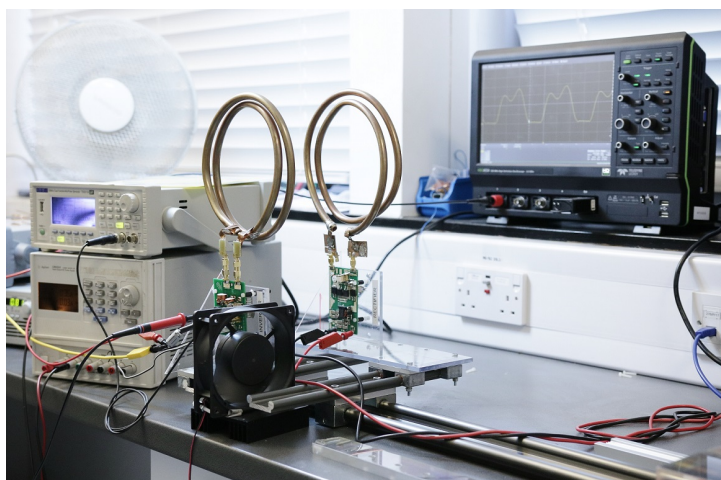


Figure 6.3: Experimental setup for variable  $k$ . Coils separation at 17.8 cm.

Two designs are presented: one for a frequency of operation of 6.78 MHz and another one for 13.56 MHz. Both designs use the same coils, the same PCB layouts and the same

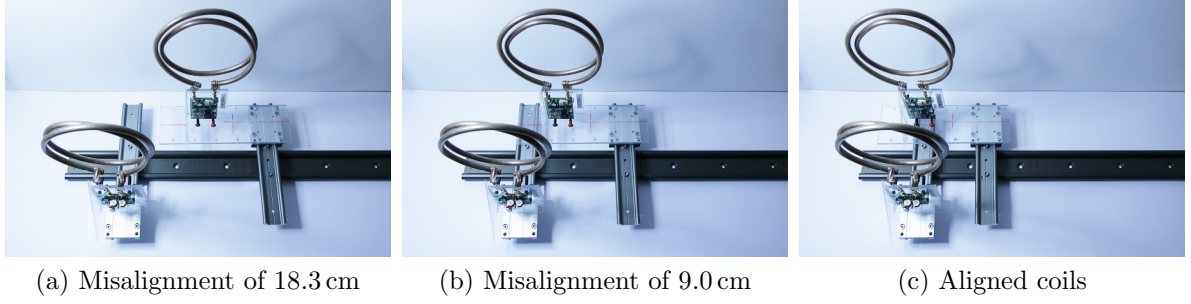


Figure 6.4: Photograph of the linear slide used to change the alignment of the coils.

switching devices but with different tuning; however, in order to obtain a similar range in output power, the coil-separation was set at 11 cm for the experiment at 6.78 MHz, and at 17.8 cm for the one at 13.56 MHz. This corresponds to a range of  $k$  of 4 to 10.5 % for the system at 6.78 MHz and 2 to 4.9 % for the system at 13.56 MHz. The distribution of  $k$  with respect to misalignment is shown in Fig. 6.6. The components values (Table 6.1) were slightly altered from the designs presented in Chapter 3. The peak efficiency achieved

Table 6.1: List of components of the circuits used in the Case Study I

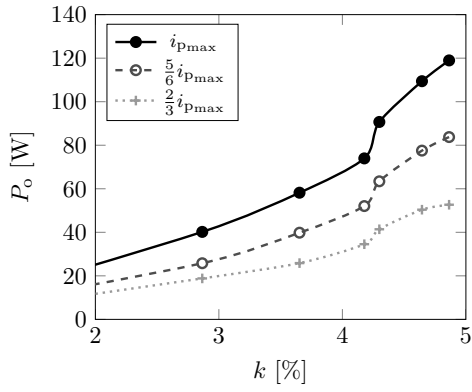
| Component             | Value (6.78 MHz)                        | Value (13.56 MHz) | Description            |
|-----------------------|-----------------------------------------|-------------------|------------------------|
| $L_p$ (nH)            | 980                                     | 989               |                        |
| $R_p$ ( $\Omega$ )    | 0.053                                   | 0.054             |                        |
| $L_s$ (nH)            | 984                                     | 990               |                        |
| $R_s$ ( $\Omega$ )    | 0.051                                   | 0.052             |                        |
| $C_1$ (pF)            | $469 + C_{oss}$                         | $180 + C_{oss}$   | Vishay QUAD HIFREQ     |
| $C_2$ (pF)            | 418                                     | 180               | Vishay QUAD HIFREQ     |
| $C_3$ (pF)            | 1003                                    | 190               | Vishay QUAD HIFREQ     |
| $L_1$ ( $\mu$ H)      | 88                                      | 88                | Würth Elektronik WE-PD |
| $L_2$ (nH)            | 404                                     | 270               | Coilcraft 2014VS       |
| $R_{eq}$ ( $\Omega$ ) | up to 6.00                              | up to 5.00        | reflected resistance   |
| $C_s$ (pF)            | 560                                     | 147               | Vishay QUAD HIFREQ     |
| $C_{rcn}$ (nF)        | 7.06                                    | 3.11              | Johanson Technology E  |
| $L_{rcn}$ (nH)        | 78                                      | 42                | Coilcraft 2014VS       |
| $R_o$ ( $\Omega$ )    | 5 to 120                                | 5 to 120          | dc-load resistance     |
| $C_o$ ( $\mu$ F)      | 1                                       | 1                 | Murata GRM Series      |
| $Q_1$                 | GS66504B (650 V, 15 A) GaN FET          |                   |                        |
| $D$                   | 10 A, 120 V Schottky diodes (FSV10120V) |                   |                        |

in both experiments was 88 % at 6.78 MHz and 83 % at 13.56 MHz. These efficiencies are to the author's knowledge, the highest ever achieved at these frequencies and coil separations. The load in the experiments was a rheostat. The IPT system DC load ( $R_o$ )

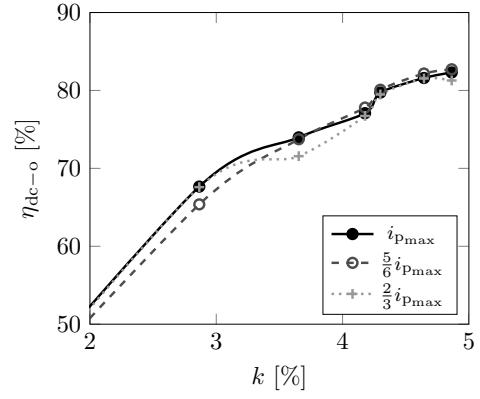
was set at  $40\ \Omega$  for the experiments at 13.56 MHz, and at  $45\ \Omega$  for the experiments at 6.78 MHz, as these achieved the highest efficiencies.

### 6.3.1 Characterising Performance with Respect to Coupling

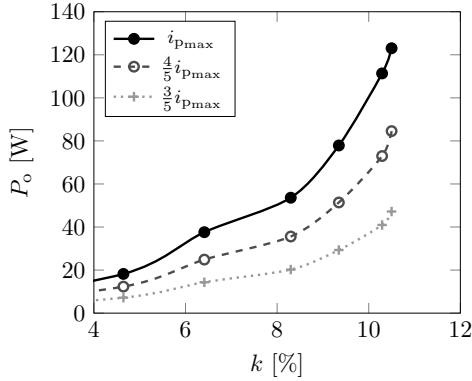
Seven coil positions were chosen to characterise  $P_o$  and  $\eta_{dc-o}$  with respect to  $k$ . Measurements are shown in Fig. 6.5. The plots show that although the efficiency deteriorates



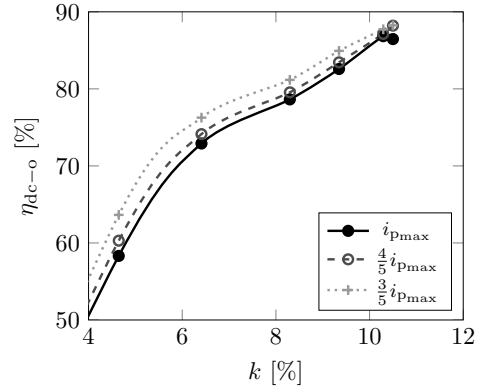
(a) Output power against  $k$  at 13.56 MHz and an  $R_o$  of  $40\ \Omega$



(b) End-to-end efficiency against  $k$  at 13.56 MHz and an  $R_o$  of  $40\ \Omega$



(c) Output power against  $k$  at 6.78 MHz and an  $R_o$  of  $45\ \Omega$



(d) End-to-end efficiency against  $k$  at 6.78 MHz and an  $R_o$  of  $45\ \Omega$

Figure 6.5: Power and efficiency measurements with respect to coupling.

as  $k$  decreases, it is fairly maintained due to the properties of the chosen circuit topologies. For example, in the experiment at 6.78 MHz, when  $i_p$  was set at a 60 % of its maximum amplitude,  $\eta_{dc-o}$  ranged from 88-70 % (20 % drop) for a 11-5.5 % (50 % drop)

coupling. The compressed dependence between  $\eta_{\text{dc-o}}$  and  $k$  is consistent throughout the  $i_{\text{p}}$  amplitudes tried at both frequencies. This is expected because the link-efficiency is independent of the amplitude of  $i_{\text{p}}$ .

A slight effect in the tuning of the inverter was perceived at 6.78 MHz when  $V_{\text{dc}}$  was changed. The range of  $V_{\text{dc}}$  on the 6.78 MHz inverter was not only larger than the one at 13.56 MHz (40 % against 33 %) but also the range of values was set at lower voltage (90-150 V against 120-180 V). These changes had an effect on  $C_1$  since the output capacitance of a transistor ( $C_{\text{oss}}$ ) of the gallium nitride (GaN) transistors are higher from 90 to 150 V than from 120 to 180 V [143]. At 6.78 MHz however, the values of the passive components are higher than those at 13.56 MHz and therefore less vulnerable to parasitic capacitances and inductances, which allows a broader range of  $V_{\text{dc}}$ .

### 6.3.2 Obtaining the Probability Density Function Based on a Known Motion Profile

In order to calculate the output-energy and the energy-efficiency of the proposed case study, in which the load travels at constant velocity over a charging pad, a probability-density-function  $g(x)$  describes the relative motion of the coils throughout the range of motion of the system ( $x_{\text{min}}$  to  $x_{\text{max}}$ ):

$$\int_{x_{\text{min}}}^{x_{\text{max}}} g(x) dx = 1. \quad (6.14)$$

The constant velocity ( $v$ ) from  $x_{\text{min}}$  to  $x_{\text{max}}$  in the time interval  $t_{\text{f}}$  to  $t_0$  is then,

$$v = \frac{\partial x}{\partial t} = \frac{x_{\text{max}} - x_{\text{min}}}{t_{\text{f}} - t_0}, \quad (6.15)$$

and therefore the relative position should be represented as a uniform distribution:

$$g(x) = \begin{cases} \frac{1}{(x_{\max} - x_{\min})}, & x_{\min} \leq x \leq x_{\max} \\ 0, & x_{\min} > x > x_{\max}. \end{cases} \quad (6.16)$$

A variable transformation  $k(x)$  should then be performed to model the probability-distribution of  $k$ , as opposed to the probability distribution of  $x$ , therefore (6.14) can be rewritten as:

$$\int_{k_{\min}}^{k_{\max}} -g(k(x)) \frac{dx}{dk} dk = 1, \quad (6.17)$$

where  $k_{\min}$  corresponds to  $k(x_{\max})$  and  $k_{\max}$  to  $k(x_{\min})$ . The probability-density-function of  $k$  is then defined as:

$$f(k) = -g(k(x)) \frac{dx}{dk}. \quad (6.18)$$

The distribution of  $k$  in space,  $k(x)$ , was found from measurements fitted as a first order Gaussian function (Fig. 6.6):

$$k(x) = k_{\max} e^{-\frac{(x-b)^2}{c^2}}, \quad (6.19)$$

where  $b$  represents the centre of the position of the peak (expected to be zero in this case study), and  $c$  represents the standard deviation ( $c^2 = 2\sigma^2$ ).

The function  $x(k)$  is then calculated considering the mathematical restrictions for the variables and constants ( $0 < k < k_{\max}$ ,  $c \neq 0$ ,  $x_{\min} < x < x_{\max}$ ):

$$x(k) = b + c \sqrt{\ln \left( \frac{k_{\max}}{k} \right)}, \quad (6.20)$$

then by deriving  $x(k)$  with respect to  $k$  and substituting it in (6.18), the probability-

density-function of  $k$  is represented as:

$$f(k) = \left( \frac{1}{x_{\max} - x_{\min}} \right) \frac{c}{2k \sqrt{\ln \left( \frac{k_{\max}}{k} \right)}}. \quad (6.21)$$

Therefore, the output-energy for this particular case study can be calculated as:

$$E_o = \frac{c}{2v} \int_{k_{\min}}^{k_{\max}} \frac{P_o(k)}{k \sqrt{\ln \left( \frac{k_{\max}}{k} \right)}} dk, \quad (6.22)$$

and the energy-efficiency as:

$$\eta_{\text{avg}} = \frac{\int_{k_{\min}}^{k_{\max}} \frac{P_o(k)}{k \sqrt{\ln \left( \frac{k_{\max}}{k} \right)}} dk}{\int_{k_{\min}}^{k_{\max}} \frac{P_i(k)}{k \sqrt{\ln \left( \frac{k_{\max}}{k} \right)}} dk}. \quad (6.23)$$

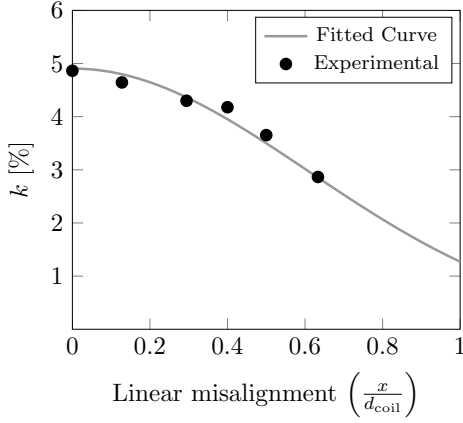
It should be noted that the proposed model can also be used if  $P_o$  is regulated against  $k$ . For example, if  $P_o$  is constant (by controlling the amplitude of  $\mathbf{i}_p$  or  $R_{\text{sec}}$ ), the input power would still change with  $k$  as previously defined, and therefore energy-efficiency would be calculated as:

$$\eta_{\text{avg}} \Big|_{P_o=\text{constant}} = \frac{(t_f - t_0) P_o}{\frac{c}{2v} \int_{k_{\min}}^{k_{\max}} \frac{P_i(k)}{k \sqrt{\ln \left( \frac{k_{\max}}{k} \right)}} dk}. \quad (6.24)$$

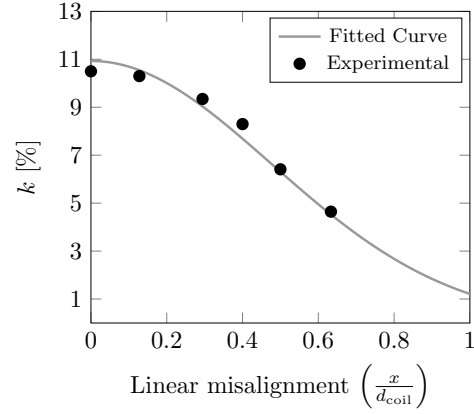
The distribution of  $k$  for both designs is shown in Fig. 6.6. This plot shows  $k(x)$  from experimental verification and the fitted curve described by the Gaussian functions using the values from Table 6.2. The measurements of  $k$  were done with the impedance analyser.

Table 6.2: Constants values for the probability density function

| Data from    | Switching frequency | $k_{\max}$ | $b$                      | $c$    |
|--------------|---------------------|------------|--------------------------|--------|
| Measurements | 6.78 MHz            | 0.1093     | $-1.84 \times 10^{-12}$  | 0.1214 |
| Measurements | 13.56 MHz           | 0.0491     | $-4.184 \times 10^{-13}$ | 0.1549 |



(a) Coil separation of 17.8 cm.



(b) Coil separation of 11 cm.

Figure 6.6:  $k$  against linear misalignment (normalised to a coil diameter distance) for coil separations of 11 cm and 17.8 cm.

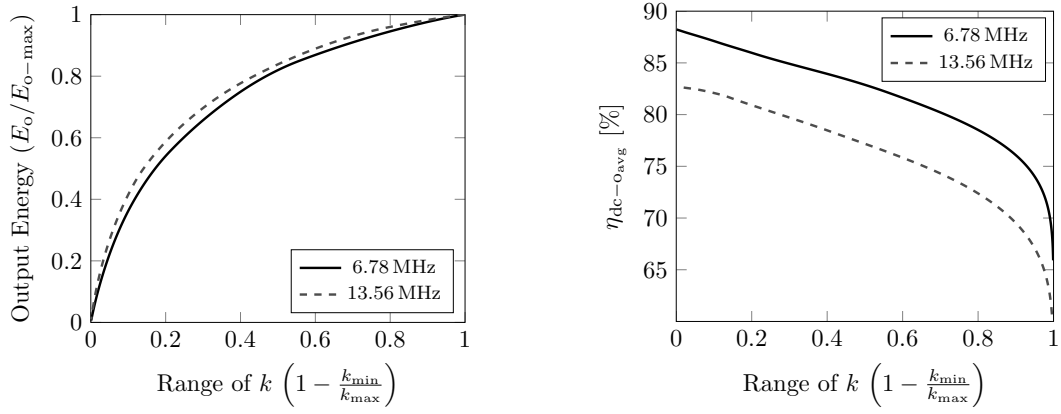
### 6.3.3 Energy-Efficiency and Output Energy with Respect to Misalignment Range

The output-energy and the energy-efficiency were calculated using (6.22) and (6.23), leaving the minimum coupling as a variable so that the allowed misalignment at which power is transferred is different to that which the system is capable of achieving.  $P_o$  and  $P_i$  are expressed as functions of  $k$  from measurements.

In order to evaluate how the relative motion of the coils affects the system energy efficiency and energy throughput, both the energy-efficiency and the output energy were plotted against the variable range of  $k$  (Fig. 6.7).  $k_{\min}$ , in this case the variable to design, represents the minimum  $k$  at which the system transfers power and  $k_{\max}$  corresponds to  $k$  when the coils are aligned.

Primarily, in a particular design, the range of motion (or the tolerance to misalignment  $k_{\min}$ ) can be selected to deliver the required amount of energy to the load with an associated energy-efficiency, which decreases as the allowed range of  $k$  increases. For example, this model can be used by the device picking up charge from a charging pad at constant velocity, by selecting when to start and stop receiving power taking into account that it can choose to narrow down the tolerance to misalignment, therefore receiving less energy





(a) Calculated output energy (normalised to the case with the highest  $k$  tolerance) with respect to range of motion

(b) Calculated energy-efficiency with respect to range of motion

Figure 6.7: Calculation of output-energy and energy-efficiency with respect to range of  $k$ .

at higher weighted average efficiency, or conversely, receive more energy with a higher tolerance to misalignment, and thus receiving power for a longer period of time but at a lower weighted average efficiency. With the proposed circuit design, at 6.78 MHz, an 80 % energy-efficiency is achieved for a tolerance to misalignment of  $k_{\min} = 0.3k_{\max}$ , and at 13.56 MHz, a 74 % energy-efficiency is achieved for  $k_{\min} = 0.3k_{\max}$ . In both cases the energy-efficiency at this point is 90 % of the maximum power efficiency. It should be noted that the tolerance to misalignment of  $0.3k_{\max}$  at 6.78 MHz represents a misalignment of 12.5 cm and at 13.56 MHz a misalignment of 17.25 cm, which demonstrates the intrinsic increase in freedom of motion (larger wireless air-gaps and tolerance to misalignment) at higher frequencies if the coils used are useful at both frequencies.

## 6.4 Case Study II: Optimising a HF-IPT System with Power Throughput Control and Unpredictable Motion

This case study demonstrates how the proposed probability-based analysis can be used to increase the end-to-end efficiency of a HF-IPT system in which the position, and therefore  $k$ , cannot be predicted as a function of time. Defining  $k$  as a stochastic variable allows to model it as a probability distribution from empirical evidence. Therefore, notwithstanding that the exact position of the device is not predicted in time, those values of  $k$  with a higher probability of occurring can be identified and the tuning of the inverter and the rectifier can be done accordingly. The system can be tuned to achieve the highest efficiency at the coupling range with the highest probability, to therefore achieve the highest charging energy-efficiency.

The experimental work presented consists of comparing a system using two rectifier designs using the hybrid Class E topology. The rectifiers are tuned for the link's optimal load at different coupling values in order to verify which version achieves the highest charging efficiency. The load in the experiments is a inductively powered drone without a battery hovering randomly over the charging pad, and the range of motion is set by a nylon string tether. The experiments show lower energy consumption when the rectifier is tuned to present the optimal load of the link at the coupling value with the highest probability, as opposed to the first design iteration, which was designed to present the optimal load of the link at minimum  $k$ .

### 6.4.1 An Overview of the Wirelessly Powered Drone as an Experimental Platform

A lightweight 13.56 MHz 15 W IPT system was designed and built to wirelessly power a drone without a battery (Fig. 6.8), to use it as a demonstrator of the lightweight and

tolerance to misalignment capabilities of HF-IPT technology. In this case study, because of its unpredictable motion, it is used as an experimental platform. An overview of the specifications of the system are presented herein, however the design and development of the system is covered in Chapter 7.

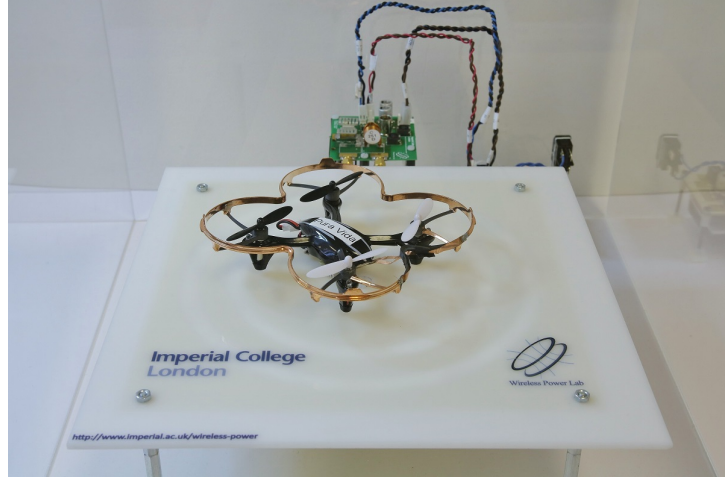


Figure 6.8: Photograph of the wirelessly powered drone.

The system was designed so that the drone could operate from the critical point of maximum power at minimum  $k$ , up to the point of maximum  $k$  and minimum power, which is when the drone is sitting on top of the charging pad and the propellers are not spinning. This is achieved by designing the inverter to be load independent using Inverter II: Class EF, and tuning the rectifier to allow for large changes on the  $\mathcal{E}_s$ . After the rectifier a DC-DC converter was integrated to feed the drone with 4.3 V. The system diagram is shown in Fig. 6.9.

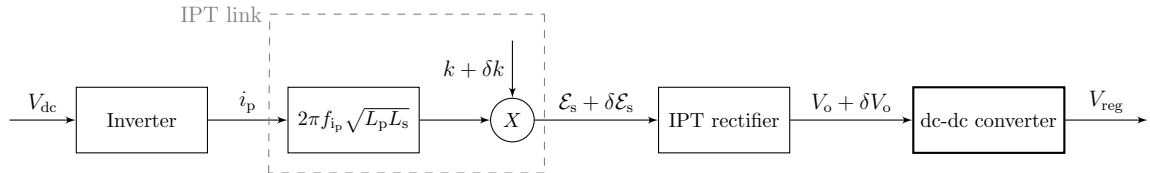


Figure 6.9: Block diagram of a HF-IPT system with variable coupling and an output voltage regulation stage.

The only control loop of the system is that of the DC-DC converter which regulates the output voltage. The output voltage regulation stage changes the real part of  $\mathbf{Z}_{\text{sec}}$  to regulate power throughput; nonetheless, it can also produce changes on the imaginary part

which would detune the receiver. The control loop of the DC-DC converter compensates for changes on the load and in  $\mathcal{E}_s$ , which are produced by changes in  $k$  since the amplitude of  $i_p$  was designed to be constant.

The drone can hover at any position within its range of motion, set by a 7.5 cm nylon string tether connected between the bottom of the drone and the centre of the transmit coil. The drone is powered from a 13.56 MHz IPT system using Inverter II: Class EF, and a hybrid Class E rectifier at the receive side. The DC-DC converter used is a LMZ14203 by Texas Instruments. Since at the receive side the battery was removed, the energy storage system on the drone consists of solely two 100  $\mu$ H ceramic capacitor in parallel at the output of the DC-DC converter. The average power consumption of the drone at full power was measured at 13 W with a feeding voltage of 4.3 V.

### 6.4.2 Characterising the Range of Motion of the Wirelessly Powered Drone

The range of motion of the system was limited by anchoring the drone to the centre of the charging pad with a nylon string tether. The maximum coupling,  $k_{\max}$ , takes place when the drone sits on top of the transmitting coil, and the minimum coupling  $k_{\min}$  when the drone hovers far from the transmitter at maximum separation. Coupling depends on the length of the tether,  $r$  in Fig. 6.10, and the angle  $\theta$ , which determines misalignment.

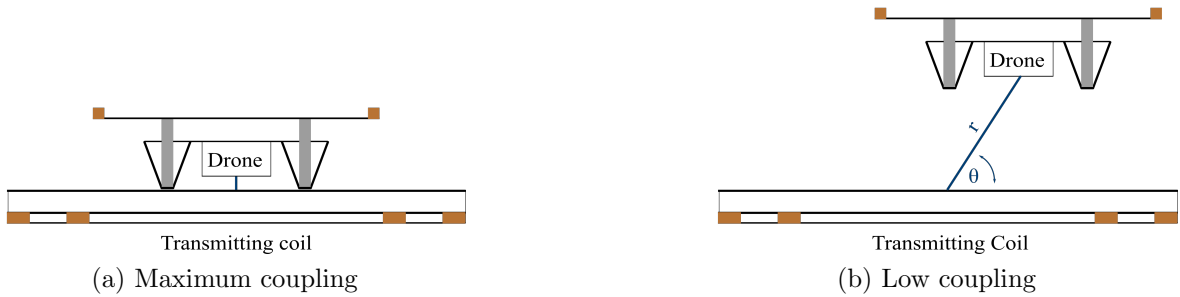


Figure 6.10: Cross section of the IPT-link describing the range of motion of the system.

The maximum simulated coupling was 23.7 %; however, the maximum measured coupling

did not exceed 20 % at any point throughout the range of motion. The length of the tether was set at 7.5 cm in order to have a minimum coupling higher than 5 %. The minimum coupling was simulated and measured at 5.8 % when  $r = 7.5$  cm and  $\theta = 0^\circ$ . The distribution of  $k$  in space is shown as a colour map in Fig. 6.11, obtained from electromagnetic simulations using CST Microwave Studio.

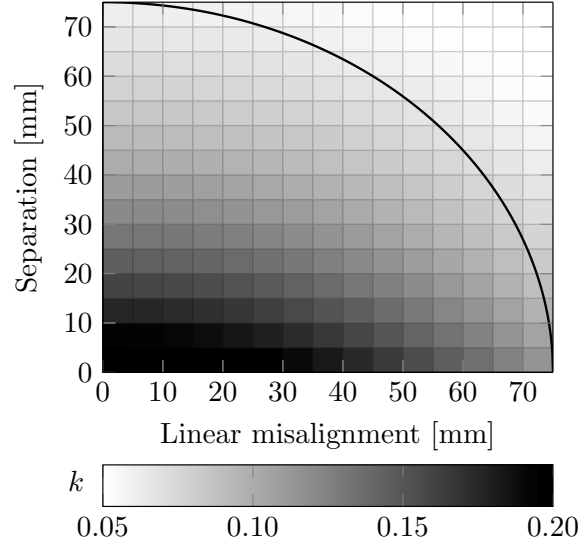


Figure 6.11: Wirelessly powered drone coupling distribution in space.

### 6.4.3 A First Iteration Design of the Rectifier for the Wirelessly Powered Drone

The receiving-end of the system consists of a hybrid Class E rectifier [87] and a DC-DC converter to regulate the output voltage. The hybrid Class E topology is shown in Fig. 3.13 (b). Class E rectifiers in HF-IPT systems can be designed at a resonant frequency different than the frequency of operation (The specific design of this rectifier is done in Section 7.2.8), given the possibility of compensating the residual reactance of the rectifier in the resonant tank of the receiving coil. This feature grants various degrees of freedom in the design, regarding among other variables, the input impedance of the rectifier and its deviation with DC-load. These rectifiers can therefore be tuned

to achieve a certain input impedance for a combination of  $k$  and output power by solely designing the passive components of the topology. The hybrid Class E rectifier grants yet another degree of freedom, which is used in this design to better delimit the output voltage range against coupling and therefore avoid surpassing the maximum or minimum voltage at the input of the DC-DC converter.

The combination of the Class E rectifier and the DC-DC converter allows designing the maximum efficiency point of the system for a particular combination of  $k$  and  $P_o$  by selecting the reflected impedance of the receiving-end to be completely resistive (i.e. the secondary side being perfectly tuned at the frequency of operation) and at the same time matching the input resistance of the rectifier to the optimal load at that particular combination of  $k$  and  $P_o$ .

The initial challenges for designing the rectifier are: being able operate efficiently at a highly variable coupling factor (5 to 20 %) and a broad range in power (from maximum to zero power at any coupling), without leaving the operational range of input voltage of the DC-DC converter that follows the rectifier. In addition, the reflected impedance to the transmitting-end cannot have a significant reactance because that would detune the inverter.

These specifications were achieved by selecting, from the possible designs proposed in [87], a hybrid Class E rectifier with a resonant frequency lower than the frequency of operation. The resonant frequency of the rectifier relative to the operating frequency:

$$A_r = \frac{f_{\text{rect}}}{f_{\text{ip}}}, \quad (6.25)$$

chosen was 0.3, which might vary slightly depending on the available commercial values for the inductance  $L_E$  in Fig. 3.13 (b).

The rectifier was designed so that the highest  $\eta_{\text{link}}$  was met at minimum  $k$  (5 %) and maximum power, considering as an assumption that the drone would mostly operate at the furthest distance from the charging pad.

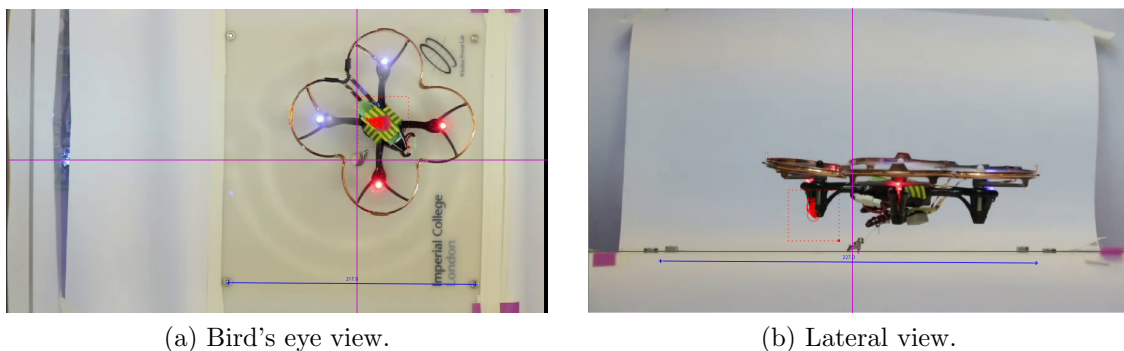
#### 6.4.4 Obtaining the Probability Distribution of a Wirelessly Powered Drone from Observation

If no external control over the flight operation of the drone is performed, then its position in time is unpredictable, and cannot be described analytically. This is the case for the demo that was built. It is operated at maximum throttle with no external control over its direction, which tends to vary unpredictably. The drone's position can however be analysed as a stochastic variable and characterised using statistical data.

Fig. 6.13 describes the distribution of  $k$  as a histogram with equal size bins of 0.5% coupling, when the drone is at maximum throttle and has no external flight control. Characterising motion as a probability distribution is useful because it allows optimising for the link and the circuits at both ends of the system for maximum overall energy efficiency.

Once the initial design is built and operating, statistical data can be gathered on the relative position of the coils, and the dynamic behaviour of the system can be characterised to formulate coupling as a probability distribution.

The relative position of the coils was monitored for one minute (3600 samples) using two video cameras, one at the top and one at the side of the IPT system under test; this to measure the linear misalignment and separation of the coils. One synchronised frame from both cameras is shown in Fig. 6.12.



(a) Bird's eye view.

(b) Lateral view.

Figure 6.12: Capture of a key frame for position tracking of the drone.

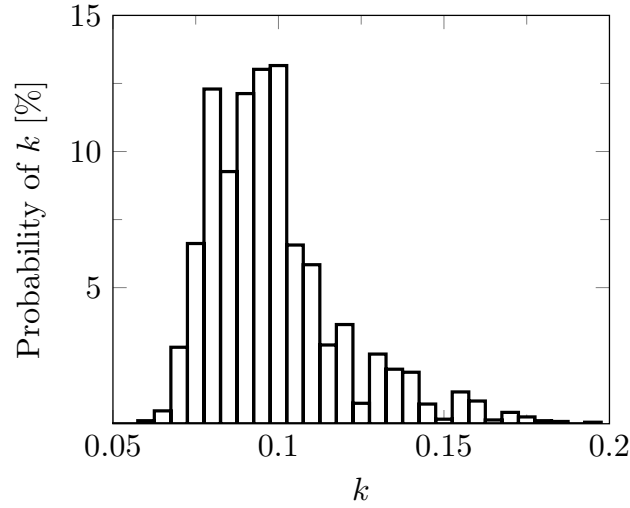


Figure 6.13: Histogram of coupling obtained from 3600 samples in one minute.

The coupling factor at each position within the range of motion of the system was calculated from simulation, and several points including the maximum and minimum coupling were verified experimentally using the impedance analyser. The distribution of  $k$  was obtained by recording the position of the drone and performing a probability mass function transformation from linear misalignment and distance to coupling.

The probability distribution is shown as a histogram in Fig. 6.13. The histogram was constructed with equal size bins of 0.5% coupling, which for this case study was determined to be the minimum step change that affects the optimum rectifier design using commercial off-the-shelf components.

#### 6.4.5 Optimised Design of the Rectifier for the Wirelessly Powered Drone

The histogram in Fig. 6.13 points out that the highest probability of coupling occurs between 8 and 10% coupling. Therefore, the initial design that assumes that the most probable value of  $k$  was 5%, operates with an input resistance of the rectifier different than the optimal load of the IPT link most of the time. Furthermore, most of the energy dissipated by the load in the one minute experiment was transferred wirelessly in



suboptimal tuning of the IPT link.

An improved version of the first rectifier, which also includes the same capabilities as the first, was built with the values of the passives as shown in Table 6.3, in order to tune the

Table 6.3: Design parameters and components values of the rectifiers

| Component            | Design 1            | Design 2             | Description              |
|----------------------|---------------------|----------------------|--------------------------|
| $k_{\text{opt}}$ (%) | 5                   | 10                   |                          |
| $A_r$                | 0.3                 | 0.3                  |                          |
| $L_s$ (nH)           | 360                 | 360                  |                          |
| $C_p$ (pF)           | 62                  | 54                   | Vishay QUAD HIFREQ       |
| $C_{hs}$ (pF)        | 343                 | 420                  | Vishay QUAD HIFREQ       |
| $C_{hp}$ (pF)        | $16+C_{\text{oss}}$ | $343+C_{\text{oss}}$ | Vishay QUAD HIFREQ       |
| $L_h$ ( $\mu$ H)     | 3.3                 | 2.2                  | Coilcraft XAL40xx Series |
| $C_o$ ( $\mu$ F)     | 20                  | 20                   | AVX X7R                  |
| $D$                  |                     | 2x Cree C3D1P7060Q   |                          |

input resistance of the rectifier to meet the optimal load at the value of coupling with the highest probability, as opposed to the first design, which was tuned to achieve the optimal load at minimum coupling.

The two designs were compared by measuring the energy consumption at the direct current (DC) source of the transmitter. The results are shown in Table 6.4.

Table 6.4: Average power consumption at the source with the different rectifier designs

| Inverter input voltage | No load | Initial Design | Optimised Design |
|------------------------|---------|----------------|------------------|
| 74 V                   | 6.6 W   | 26.67 W        | 21.78 W          |

A detailed description of the design of this rectifier is presented in Section 7.2.8.

## 6.5 Summary

A probability based model is introduced to analyse the energy-efficiency and the output-energy in terms of the range of motion of the dynamic charging system. The basis of the model is the function  $f(k)$ , a probability-density-function that models the distribution

of  $k$  throughout the range of motion of the system for a given charging time, which represents the motion behaviour of the device being charged and the coupling distribution of the system. The main advantages of calculating the output-energy and the energy-efficiency using this model is that primarily, for a particular design, the selection of the range of motion can be determined by considering the throughput energy and the energy-efficiency associated to the charging process. Also, by analysing the system as proposed, further optimisation, considering the mobility of the device being charged, can be achieved by tuning the circuits to a more accurate value of  $k$  which would improve the energy-efficiency, or by improving the design of the IPT-link to obtain a probability-density-function associated to a higher energy-efficiency or a broader range of motion. This work was used extensively in the following chapter where a detailed design of a wirelessly powered drone is described.

# Chapter 7

## End-to-End Application

## Demonstrations of HF-IPT Systems

The content in this chapter focuses on original work involving design and construction of high frequency inductive power transfer (HF-IPT) systems to demonstrate the capabilities of the technology while integrated into applications. The demonstrators build on work performed in the previous chapters, in particular Chapter 5 (where a system topology for HF-IPT includes the mains interface), and Chapter 6 (where a probability-based model is introduced to design dynamic systems). Part of the work discussed in this chapter has been published in IEEE Transactions on Power Electronics [84] and an IEEE Conference Proceedings [144]. Previous versions of one of the demonstrators have been published in [83, 133, 145], and results of a related project have been published in [146, 147].

### 7.1 Introduction

This chapter covers the design and construction of two end-to-end HF-IPT systems designed to be integrated into applications. The design examples presented herein take into account the characterisation and optimisation techniques proposed in the previous chapters of the thesis. The two HF-IPT systems developed are a 13.56 MHz, 15 W system

to wirelessly power an unmanned aerial vehicle (UAV) while it hovers close to a transmit pad; and a 6.78 MHz, 100 W system to wirelessly charge an electric scooter. Both are ongoing projects. The design solution to wirelessly power a drone without a battery is now being implemented to charge the on-board battery of a Matrice 100 drone by DJI at 100 W, and the design principles used to charge the electric scooter are now being employed to charge a Govecs S2.4 Electric Scooter at 600 W.

## 7.2 The Wirelessly Powered Drone

This section demonstrates the design and construction of a HF-IPT system designed to power a drone without a battery which can hover freely in proximity to a 13.56 MHz transmitter. The section covers the complete system level design which provides uninterrupted power flow efficiently while allowing for variable power demand and coupling factor ( $k$ ). It includes the design of lightweight air-core coils that can achieve a high-enough inductance and  $k$  for the required power level at 13.56 MHz, without degrading the aerodynamics of the drone. The design of the resonant power converters at both ends of the system is also included. At the transmit side, a load-independent Class EF inverter based on Inverter II: Class EF was selected, and at the receive side a hybrid Class E rectifier, which allows tuning for large changes in  $k$  and power demand was used.

For the demo, and as covered in section 6.4, the range of motion of the drone was limited by a 7.5 cm nylon string tether, connected between the centre of the transmit coil and the bottom of the drone. The experimental results and specific practical considerations required for the physical implementation are provided. An average end-to-end efficiency of 60 % was achieved for a coupling range of 23 % to 5.8 %. Relevant simulations concerning human exposure to electromagnetic fields are also included to assure that the demonstrator is safe according to the relevant guidelines.

The demonstrator has been shown at the Applied Power Electronics Conference (APEC) 2017, Centre for Power Electronics UK Annual Conference 2017, 2018 and 2019, PCIM

Europe 2017 and IEEE Wireless Power Week 2018 and 2019. A video of the system in operation can be found in [148].

The work presented in this section is also relevant to a project where a drone is employed to feed power and obtain data from a sensor at the same time. Some of the results of that project have been published in [146, 147].

### **7.2.1 Introduction to Wireless Power for UAVs**

Considering the vast range of applications for UAVs [149], in this work a demonstrator of a HF-IPT system that can wirelessly power a drone without a battery while it hovers close to a charging pad was developed to showcase features that are beneficial, and in some cases necessary, to wirelessly power or charge dynamic loads such as a hovering UAV.

One of the biggest challenges addressed in the demonstrator of a wirelessly powered drone by William C. Brown in 1969 [13], also pointed out in more recent wireless power transfer (WPT) systems proposed to wirelessly charge UAVs [150], is dealing with wireless power transfer and variable position. This considering either dynamic charging or tolerance to misalignment as the drone lands. There are wireless power solutions which have recently been introduced into the UAV market [151]; however, the tolerance to misalignment of such system is very limited, and it is known that precision landing to the degree required by these systems can be challenging and in cases not possible [152].

### **7.2.2 Designed System Overview**

The UAV model implemented is a Hubsan H107L X4 quadcopter. This commercially available drone comes with a lithium polymer battery that outputs a voltage of 3.7 to 4.2 V, depending on the state of charge. At full throttle the motors of the drone draw approximately 13.0 W from the source. The battery was replaced with an IPT receiver,

enabling the drone to operate without an energy storage system in proximity to the charging pad, this to demonstrate the capabilities of the technology handling a variable load with variable position. Fig. 7.1 shows a photograph of the demo, including the Hubsan H107L X4 quadcopter, with an on-board inductive power transfer (IPT) receiver instead of a battery.

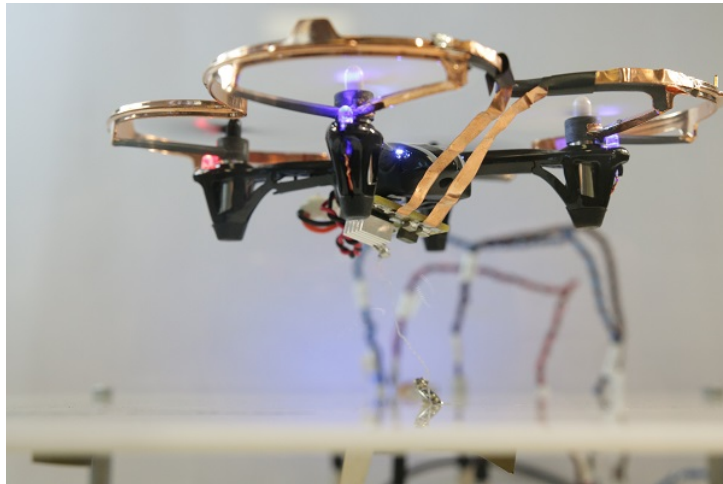


Figure 7.1: Photograph of the wirelessly powered drone hovering over a charging pad.

The selected frequency of operation was 13.56 MHz, this because the design of the coils (size, shape and weight), especially the one on the drone, is very restrictive. The option of including magnetically permeable materials was dismissed, since these would add weight and most likely interfere with the air-flow of the propellers. Also, a planar coil with multiple turns would produce similar disadvantages at the receive side. The benefit of operating at 13.56 MHz is that a single-turn air-core coil of roughly 10 cm diameter produces a sufficiently large reactance to load the transmitter, and therefore supply the power demand of the drone at minimum  $k$ , which for this experiment was 5.8 %.

### 7.2.3 The HF-IPT System Architecture and Power Throughput Control for the UAV

The system architecture implemented is shown in Fig. 7.2. The inverter was operated in open loop (constant duty cycle and constant frequency as proposed in Section 3.2.4) and

the DC input voltage ( $V_{dc}$ ) was set at 74 V to output a transmit-coil current ( $i_p$ ) with an amplitude of approximately 3.8 A. The power throughput control was performed by varying input impedance of the IPT-rectifier ( $Z_{sec}$ ). To vary  $Z_{sec}$ , a pulse-width mod-

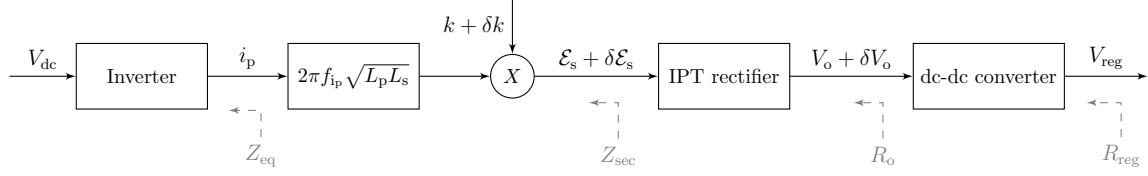


Figure 7.2: Block diagram of a HF-IPT system (including reflected impedances) with variable coupling and an output voltage regulation stage.

ulation (PWM)-based DC-DC converter is integrated as the last power conversion stage in the receiver, as can be seen in the diagram in Fig. 7.2, to regulate the output voltage with duty cycle ( $\delta_{dc-dc}$ ). As a consequence, the DC-DC converter's input resistance ( $R_o$ ) changes with its input voltage (or the rectifier's output voltage,  $V_o$ ) and its load resistance ( $R_{reg}$ ):

$$R_o = \eta_{dc-dc} \left( \frac{V_o}{V_{reg}(\delta_{dc-dc})} \right)^2 R_{reg}, \quad (7.1)$$

where  $\eta_{dc-dc}$  is the efficiency of the DC-DC converter and  $V_{reg}$  is the regulated output voltage.

This raises challenges for the rectifier design because it must operate at highly variable induced electromotive force on the receive coil ( $\mathcal{E}_s$ ) (2.7) and at variable load resistances (7.1). In addition, the range of the IPT-rectifier output voltage ( $V_o$ ) has to be limited not only by the capabilities of the components of the rectifier but also by the range of input voltages supported by the DC-DC converter.

## Electric Dynamics of the System

The dynamics can be analysed at both ends of the system separately when the inverter is load independent, given that this feature detaches the design of both ends of the system. However, at the transmit side, the reflected resistance on the transmit coil ( $R_{eq}$ ) and

the reflected reactance on the transmit coil ( $X_{eq}$ ) can reach values beyond the range of steady-state operation as the system enters a transient, which may temporarily overload the inverter and produce unwanted changes in the behaviour of  $i_p$ , or in some cases surpass the ratings of the devices in the inverter. Since there is no feedback-loop at the transmit side to regulate power throughput, the dynamics of the transmitter have to be considered solely for the design of the input filter and the maximum ratings of the devices for step changes in  $k$  and  $R_L$ .

At the receive side, the response of  $V_o$  has to be considered as a consequence of a step change in  $i_p$ ,  $k$  (or  $\mathcal{E}_s$  representing both) and the load. The behaviour of the DC-DC converter and the design of the filters determine whether the system remains stable over the entire operating range of  $k$  and load.

## 7.2.4 Design of an Inductive Link for Wireless Charging of UAVs

The receive coil in this application is one of the most restrictive elements in the design. It is required to be light-weight and the electric conductors and magnetic cores (if any) which comprise the coil should not interfere with the aerodynamics of the drone. The size and shape of the coils also determine the coupling distribution in space, confining a range of motion for a given range of  $k$ . The design of the transmit coil is less restrictive since weight and aerodynamics are not as crucial as in the receive coil. Nonetheless, a system in which the transmitter is mounted on a drone is discussed in Section 7.2.13.

The receiving air-core coil was fabricated by copper-plating a plastic structure with the shape of the propeller guard of the drone as shown in Fig. 7.3, and Fig. 7.4. This minimises the effect of the coil on the aerodynamics of the drone, and at the same time achieves high coupling ( $k > 5\%$ ) over a broad range in distance and misalignment, due to the large loop area of the coil.

It was verified in previous experiments on the unmodified drone that the presence of the



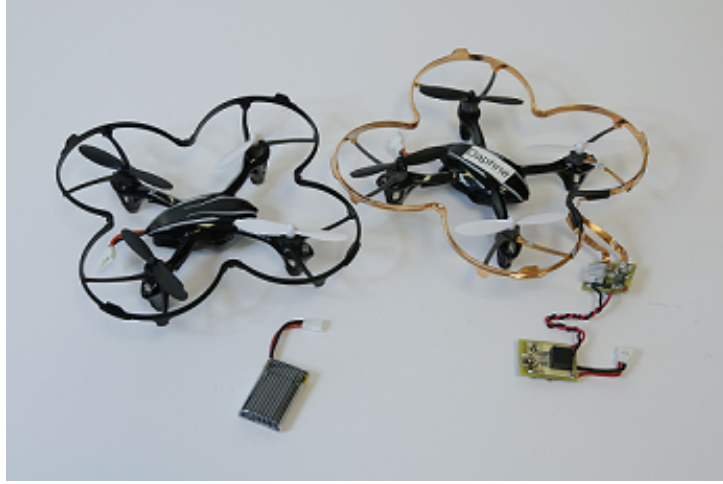
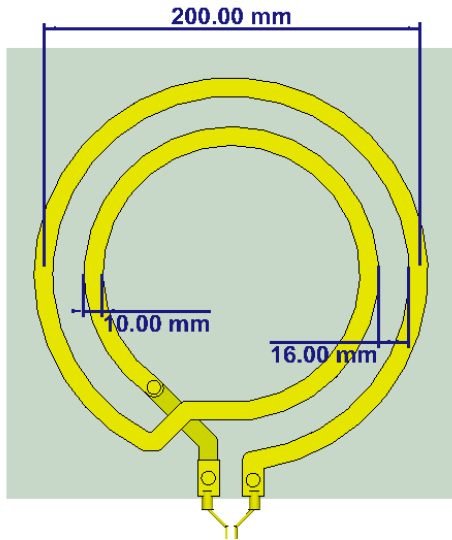
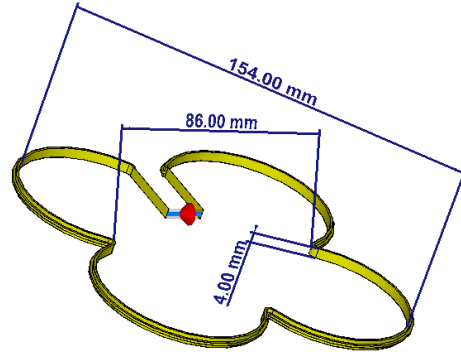


Figure 7.3: Photograph of the off-the-shelf drone with its battery (left) next to the modified drone with a plated copper coil, the rectifier board and the DC-DC converter (right).



(a) Transmit coil



(b) Receive coil

Figure 7.4: Coils diagrams for the wirelessly powered drone demonstrator.

magnetic field generated by the proposed HF-IPT transmitter had no effect on the flying controls and also did not produce heating anywhere in the drone.

The transmit coil consists of a two turn circular printed circuit board (PCB) coil with an external diameter of 200 mm and no magnetically permeable materials. The specifications of this coil, described in Chapter 4, can be seen in Fig. 7.4 (a). Both coils were simulated using CST and measured with the impedance analyser. It should be noted, however, that the equivalent series resistance (ESR) of high-Q coils is very difficult to

measure accurately [102, 103], and therefore the values from simulations are preferred over measurements for the design of the inverter and the rectifier. The specifications of the coils are presented in Table 7.1, and the coils diagrams are shown in Fig. 7.4.

Table 7.1: Specifications of the coils for the wirelessly powered drone demonstrator at 13.56 MHz

|                                    | Transmit coil   | Receive coil            |
|------------------------------------|-----------------|-------------------------|
| Shape                              | Circular planar | Drone's propeller guard |
| External Diameter [cm]             | 20              | 9-16                    |
| Number of turns                    | 2               | 1                       |
| Measured inductance [nH]           | 1181.0          | 379.9                   |
| Simulated inductance [nH]          | 1155.7          | 392.0                   |
| Measured resistance [ $m\Omega$ ]  | 202.7           | 103.9                   |
| Simulated resistance [ $m\Omega$ ] | 191.2           | 89.0                    |
| Measured unloaded Q                | 496             | 312                     |
| Simulated unloaded Q               | 515             | 375                     |

### 7.2.5 Range of Motion of the Wirelessly Powered Drone

The range of motion of the drone was limited by anchoring it to the centre of the charging pad with a 7.5 cm nylon string tether. The coupling distribution in space and the motion characterisation are presented in Section 6.4.

### 7.2.6 Simulations of Electromagnetic Exposure

Simulations of the specific absorption rate (SAR) were performed using 3D electromagnetic simulation software at the critical points of operation in terms of exposure of the wirelessly powered drone demonstrator, which was designed to be enclosed in a cubic Perspex box of 40 cm wide. The simulations results at the two most critical points are shown in Fig. 7.5 considering the amplitude of  $i_p$  equal to 5.5 A, which is the worst case among the viable solutions to power the drone efficiently with the proposed solution.

The simulations show that the highest localised SAR value is  $0.001\,990\,47\text{ W kg}^{-1}$ , which is more than a thousand times lower than the SAR limit for the head and trunk, for general

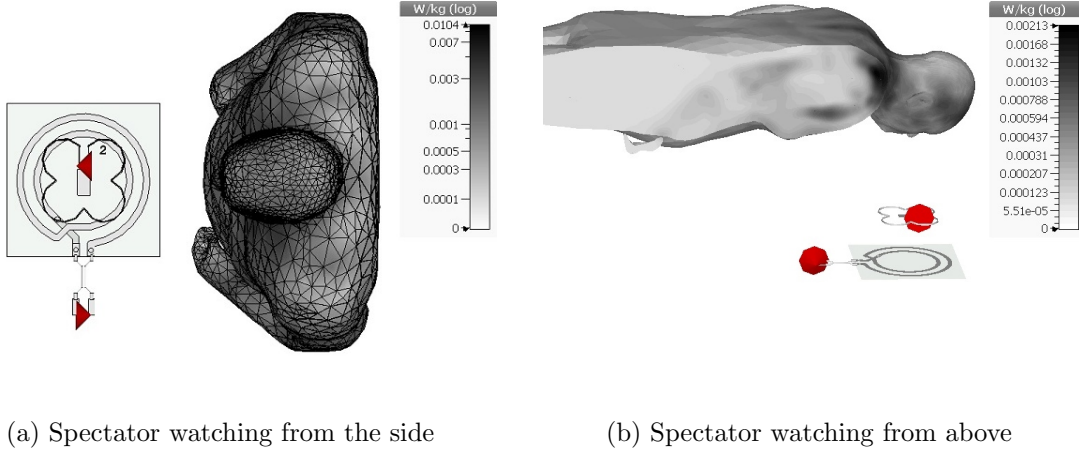


Figure 7.5: SAR simulations with an  $i_p$  of 5.5 A.

public exposure according to the International Commission on Non-Ionizing Radiation Protection (ICNIRP) [21, 22].

### 7.2.7 The Inverter Designed to Wirelessly Power a Drone

Inverter II: Class EF (Section 3.2.2) was chosen, primarily because achieving high efficiency independent of the reflected load is very convenient when dealing with an IPT system in which  $k$  varies. Also, it allows the design of the transmitter and receiver to be done separately. The load-independent feature detaches the effects of the receiver on the transmitter, simplifying the dynamics of the system and allowing power throughput control to be implemented only at the receive side. Moreover,  $i_p$  can be considered load-independent, as does  $\mathcal{E}_s(k, i_p)$ . The transmitter's purpose in this design is to generate a load-independent alternating magnetic field allowing the receiver to cope with power throughput control as  $k$  or power demand change.

Fig 7.6 shows experimental drain-voltage waveforms of the switch when the drone was operating at different throttle values. It can be seen that, as expected, the drain voltage is zero at turn-on ( $f_{ip}t = 1$ ) for all cases.

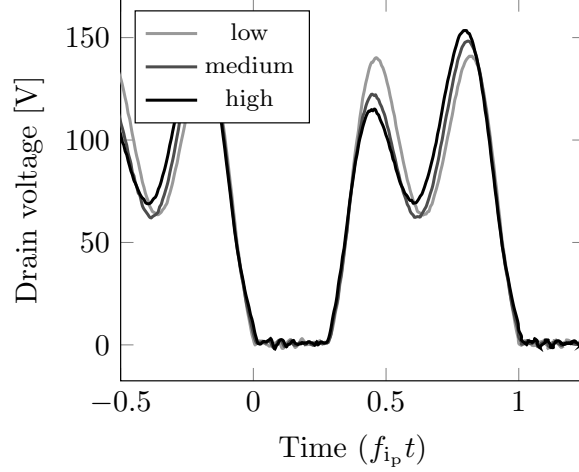


Figure 7.6: Experimental waveforms of the drain voltage of Inverter II: Class EF while wirelessly powering the drone at three throttle values.

### 7.2.8 The Rectifier Designed to Wirelessly Power a Drone

Various topologies were considered when selecting the rectifier for this system. A half-wave current-driven Class D rectifier was implemented in the first version of the demonstrator [145]. However, the drone needed to remain fairly distant from the charging pad not to exceed the maximum input voltage of the DC-DC converter. In order to allow operating at broader ranges in  $k$ , a non-synchronous hybrid Class E rectifier that can be designed for low input reactance deviation as the load changes was built. An alternative solution based on a synchronous Class E rectifier is also mentioned in Section 7.2.12. The hybrid Class E rectifier topology is shown in Fig. 7.7. The benefits of this topology in comparison to other non-synchronous half-wave Class E rectifiers were discussed in [87], and the equations presented here were derived from that work.

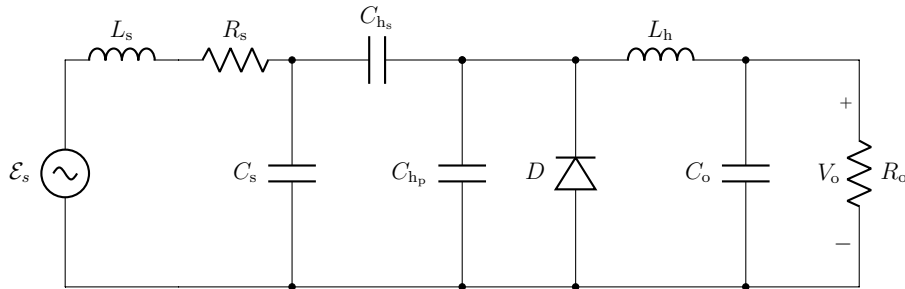


Figure 7.7: Circuit topology of the implemented hybrid Class E rectifier.

The major restriction in the design of the rectifier for this application is the range of the output voltage of the rectifier, dictated by the tolerance of the components and most importantly by the range of input voltages of the following stage. The DC-DC converter chosen for this design, the LMZ14203 by Texas Instruments, allows input voltages from 6 V to 42 V. Therefore,  $V_o$  should remain within that range considering the domain of  $\mathcal{E}_s$  (which can be calculated from (2.7)) to assure uninterrupted power flow and avoid damaging the DC-DC converter.

Some of the design principles, and the criteria to select some of the design variables in presented in Section 6.4. The design process of the rectifier begins by selecting at which point of  $k$  to optimise the link: ( $k_{\text{opt}}$ ), and calculating the optimal load of the link at that point. For this parallel-tuned IPT system the optimal load can be calculated as:

$$R_{\text{acopt}} = \frac{1}{2\pi f_{\text{ip}} C_s - \frac{1}{X_{\text{in}}}} \left( \frac{Q_p}{\sqrt{1 + k_{\text{opt}}^2 Q_p Q_s}} \right). \quad (7.2)$$

$Q_p$  and  $Q_s$  are the Q-factors of the transmit and receive coils respectively and  $X_{\text{in}}$  is the input reactance (positive if inductive and negative if capacitive) of the rectifier. Equation (7.2) assumes the circuit in Fig. 7.8 to be at resonance.

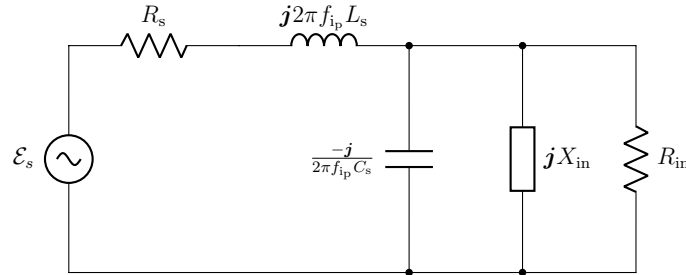


Figure 7.8: Equivalent impedances of the receiver of an IPT system with the proposed impedance model of the hybrid Class E rectifier.

$k_{\text{opt}}$  was selected using the probability distribution presented in Fig. 6.13 in order to optimise the system in terms of efficiency at the most probable  $k$ . At this point it is important to verify that the range in  $R_{\text{eq}}$  and  $X_{\text{eq}}$  are within the range tolerated by the inverter for the entire range of  $k$  and  $P_o$ . After optimising the link, the amplitude of  $i_p$

should be defined to provide the power required by the load  $P_o$ .

The design of the rectifier is relevant to achieve operation at the optimal load since the rectifier's AC-DC gain ( $M_v$ ) can be determined by the passive components of the topology. The rectifier's input resistance ( $R_{in}$ ), which should be equal to the optimal load at  $k_{opt}$ , can be described as a function of IPT system DC load ( $R_o$ ), which is the DC-DC converter's input resistance, as:

$$R_{ac_{opt}} = R_{in} = \frac{R_o}{2M_v^2}. \quad (7.3)$$

$M_v$  depends of three parameters. First, the diode's duty cycle which is strongly related to the loaded Q-factor of the rectifier,

$$Q_{rect} = \frac{R_o}{X_{rect}}. \quad (7.4)$$

$X_{rect}$  is the reactance of  $C_{hp} + C_{hs}$ :

$$X_{rect} = \frac{1}{2\pi f_{ip} (C_{hp} + C_{hs})}. \quad (7.5)$$

The diode's duty cycle can therefore be designed by sizing the inductance  $L_h$  and the capacitance  $C_{hp} + C_{hs}$  in relationship with the range of  $R_o$ .

Second, the ratio between the resonant frequency of the reactive elements of the rectifier and the frequency of operation:

$$A_r = \frac{f_{rect}}{f_{ip}}, \quad (7.6)$$

where,

$$f_{rect} = \frac{1}{2\pi \sqrt{L_h (C_{hs} + C_{hp})}}. \quad (7.7)$$

Third, the ratio between the two capacitances of the topology:

$$B = \frac{C_{\text{hp}}}{C_{\text{hs}}}. \quad (7.8)$$

The equation that describes  $M_v$  as a function of these variables can be found in Appendix A of [87].

In addition to designing the input resistance of the rectifier to be equal to the optimal load, the input reactance, along with how it changes as coupling or power demand changes, has to be considered in the design of the rectifier. This can be done by representing the circuit connected to the receive coil as a single impedance  $\mathbf{Z}_{\text{sec}}$  and can be calculated from the circuit in Fig 7.8 as:

$$R_{\text{sec}} = \frac{1}{R_{\text{in}} \left( \frac{1}{R_{\text{in}}^2} + \left( \frac{1}{X_{\text{Cs}}} - \frac{1}{X_{\text{in}}} \right)^2 \right)}, \quad (7.9)$$

$$X_{\text{sec}} = \left( \frac{1}{X_{\text{in}}} - \frac{1}{X_{\text{Cs}}} \right) \left( \frac{1}{\frac{1}{R_{\text{in}}^2} + \left( \frac{1}{X_{\text{Cs}}} - \frac{1}{X_{\text{in}}} \right)^2} \right), \quad (7.10)$$

considering that  $X_{\text{in}}$  is positive if inductive and negative if capacitive. In order to achieve the maximum link efficiency at  $k_{\text{opt}}$ ,  $X_{\text{sec}}$  has to be equal to  $-2\pi f_{\text{ip}} L_s$ . However, since  $R_{\text{in}}$  and  $X_{\text{in}}$  change with load demand and  $\mathcal{E}_s$ , the link becomes slightly detuned when operating at a different point than the one of optimisation. This feature can have a negative impact on the end-to-end efficiency as it slightly detunes the receive side when operating away from  $k_{\text{opt}}$ , but nonetheless it allows operating at maximum coupling and minimum power without exceeding the voltage ratings of the DC-DC converter, significantly broadening the range of operation of the system. The design variables of the rectifier are presented in Table 7.2, and the values of the components are in Table 7.3.

Table 7.2: Design variables of the hybrid Class E Rectifier on the wirelessly powered drone

| $k_{\text{opt}}$ | $M_v$  | $d_{\text{opt}}$ | $A_r$ | B    |
|------------------|--------|------------------|-------|------|
| 0.1              | 0.0519 | 0.4              | 0.3   | 0.95 |

Table 7.3: Components of the hybrid Class E rectifier for the wirelessly powered drone

| Component        | Value              | Description              |
|------------------|--------------------|--------------------------|
| $C_p$ [pF]       | 54                 | Vishay QUAD HIFREQ       |
| $C_{hs}$ [pF]    | 420                | Vishay QUAD HIFREQ       |
| $C_{hp}$ [pF]    | $343 + C_{oss}$    | Vishay QUAD HIFREQ       |
| $L_h$ [ $\mu$ H] | 2.2                | Coilcraft XAL40xx Series |
| $C_o$ [ $\mu$ F] | 20                 | AVX X7R                  |
| $D$              | 2x Cree C3D1P7060Q |                          |

### 7.2.9 A SPICE Simulation on Changes in Load and Coupling

In order to provide the required power to the load, the designed amplitude of  $i_p$  is determined for the range of  $\mathcal{E}_s$  and  $k$ . It is critical that the design is verified for the entire range of  $k$  and  $P_o$ , especially considering that the IPT receiver AC to DC efficiency ( $\eta_{\mathcal{E}-o}$ ) is also bounded to change with these variables.

The following were considered for the SPICE simulation. The average power consumption of the drone was measured at full throttle at 13.0 W when fed from a 4.2 V DC source, and the most probable  $k$ , selected to be  $k_{\text{opt}}$  in the design of the rectifier, is 10 %. This was obtained from the analysis in Section 6.4. The input voltage of the inverter was set at 74 V resulting in an  $i_p$  of approximately 3.8 A.

The behaviour of the system was simulated to verify its operation at the critical cases ( $k_{\text{max}}$  and  $P_{o\text{min}}$ , and  $k_{\text{min}}$  and  $P_{o\text{max}}$ ) and the dynamic response between those cases assuming step changes in load and  $k$ , from minimum to maximum and conversely. The simulated circuit consists of a voltage source representing  $\mathcal{E}_s$ , the passive components of the rectifier and the off-the-shelf SPICE model of the DC-DC converter. The load was modelled as a resistor. The results of this simulation are shown in Fig. 7.9.



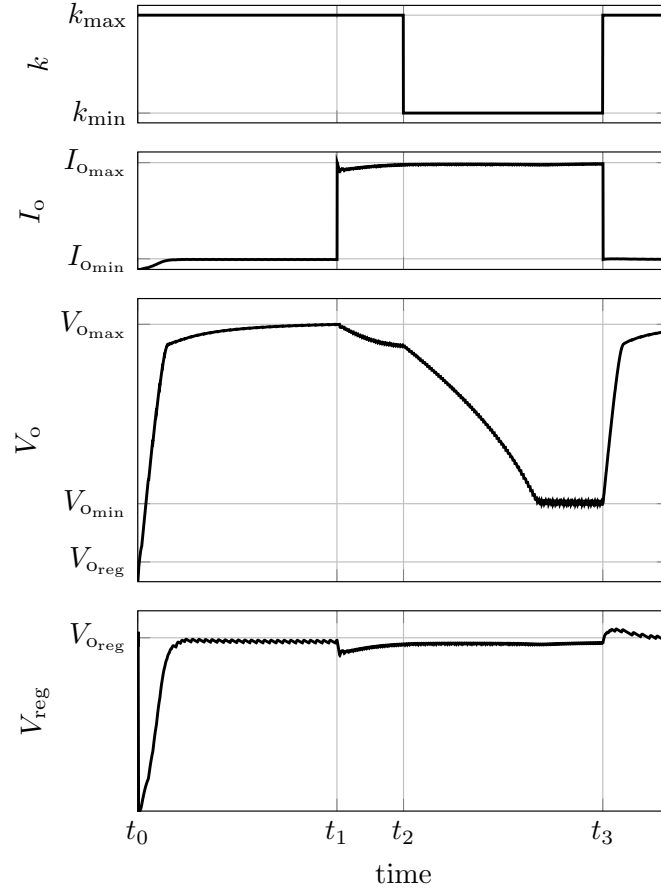


Figure 7.9: SPICE simulation results performing step-changes between critical conditions in  $k$  and load resistance.

The range in  $V_{\text{reg}}$  is well within the range of operation considering that the drone without modification is fed from a battery, the voltage of which ranges from 4.2 to 3.7 V depending on the state of charge. The most critical value in this analysis is the range in  $V_o$  which was verified to be within the boundaries allowed by the DC-DC converter: 42 and 6 V. Interestingly, if the DC-DC converter allowed a broader input voltage range, the system could operate at a broader range in  $k$ , resulting in higher spatial freedom.

### 7.2.10 Measurements on the Wirelessly Powered Drone Demonstrator

The average power consumption of the drone was measured at 13.0 W and the average input power of the inverter was measured at 21.8 W, and therefore the average DC to regulated-DC efficiency is calculated to be 60 %. The total losses on average calculated at 8.8 W are distributed, according to simulations, as shown in Fig. 7.10. The power

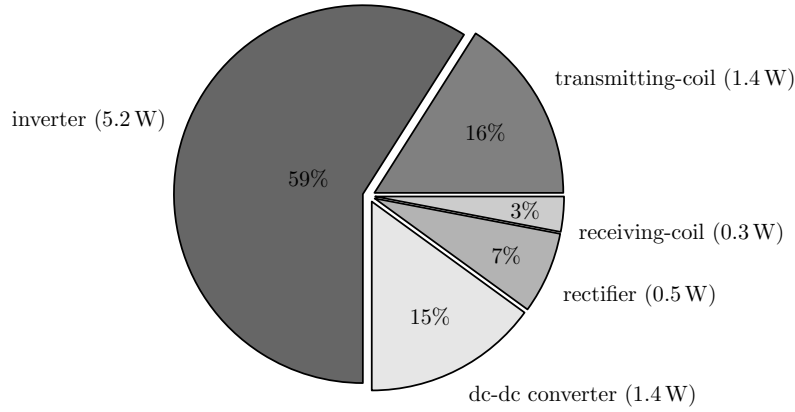


Figure 7.10: Distribution of losses in the wirelessly powered drone demonstrator from simulation.

consumption at the transmitter, when the system was driving the transmit coil without the drone, was measured at 6.6 W; therefore, it can be assumed that the losses in the transmitter account for approximately 75 % of the total losses since the losses in the load-independent Class EF inverter are fairly constant as the load varies due to its load-independent switching performance. Also, since  $i_p$  is load independent, the losses in the transmit coil do not change with load.

According to the calculations presented in Fig. 7.10, the average link efficiency is 90 %.

#### Thermal Considerations at the Transmit Side

The inverter accounts for most of the power losses in the system. Most of the losses in this topology are in  $L_2$  and the transistor [73]. The power losses in the transistor (estimated

at 2.0 W) are dissipated through the PCB and a small heatsink attached at the bottom of the PCB. The temperature of the board remained well under 70 °C at no load, and it decreased when the load was operating because of the forced airflow produced by the drone.

### **Thermal Considerations at the Receive Side**

A small heatsink was added to the rectifier board since the temperature was increasing above 70 °C at low coupling and maximum throttle (i.e. low voltage and high current). The forward voltage of silicon carbide diodes tends to increase with temperature [153], and therefore, in order to achieve good efficiencies at low couplings, a maximum temperature rating of 70 °C was determined at the rectifier board. At maximum  $k$  and maximum throttle the temperature of the diodes was much lower than 70 °C as they draw less current. This point, however, is critical to the DC-DC converter as the voltage ratio between the input and output terminals is at its maximum, resulting in lower efficiency at this stage. The DC-DC converter could operate enclosed in the compartment that typically holds the battery without a heatsink.

## **7.2.11 Additional Experiments With Wirelessly Powered Drones**

An ongoing project related to wireless power to small drones is finding alternative designs for the transmit coil. Fig. 7.11 shows a photograph of such example. The coupling distribution of a system using this coil, comprised of a single-turn larger copper trace (500 mm by 280 mm), is steadier as the alignment of the coils change. It can also feed power to multiple drones without batteries, simultaneously. It uses the same inverter with a slightly different tuning (slightly higher  $V_{dc}$  and a different  $C_3$ ), and the additional load to the inverter caused by the second drone did not require an additional heatsink or a fan. Coupling distribution and system optimisation, as proposed in Chapter 6, could be implemented for the design of the coils in order to obtain different specifications, for

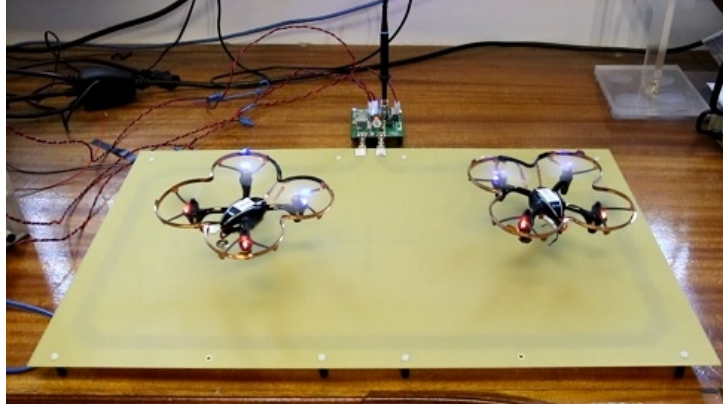


Figure 7.11: Photograph of two wirelessly powered drones fed from a single transmitter.

example a lower variation in coupling in terms of linear misalignment.

### 7.2.12 An Alternative Rectifier for the Wirelessly Powered Drone: A Synchronous Class E

A viable alternative to the hybrid Class E rectifier designed as shown in Section 7.2.8, is a synchronous Class E rectifier operating at constant duty cycle. The benefits of this mode of operation are: avoiding very low duty cycles (less than 1 %) when the power demand is at its minimum, which could be the case for any non-synchronous Class E rectifier. Also, this type of rectifier achieves broader possibilities in shaping the reflected impedance of the rectifier as the load changes. For the wirelessly powered drone prototype, the rectifier was designed using the solutions for load independent operation in [73] to have an output voltage that is independent of the load.

The solution implemented to trigger the operation of the transistor, which is difficult in this type of system, is detailed in [81].

The demo implementing the synchronous rectification solution draws more power from the source than the one that uses the hybrid Class E rectifier, mostly because of losses associated with driving the transistor, which in an application of such low power are not negligible. The synchronous rectifier can produce a lower  $X_{eq}$  than the hybrid Class E, hence avoiding detuning the inverter at higher power demands. This, and the possible

efficiency benefits of using a GaN transistor instead of a diode as the main switching device, suggest that this alternative might be more suitable for higher power applications.

### 7.2.13 Experiments with a HF-IPT Transmitter on a Drone

The lightweight features of HF-IPT are not limited to the receive side. Fig. 7.12 shows a drone with an on-board HF-IPT transmitter, powered by the drone's battery. Notwithstanding that Inverter II: Class EF is capable of powering loads of up to 150 W, the power capability in this application is dictated by the drone's power supply (the drone uses a battery model TB47D-4500-22.2-6S from DJI that can output a maximum of 100 W continuously, including the power needed to operate the drone). A DC-DC converter was integrated to boost up the voltage from a 22.2 V level of the battery to the voltage required by the inverter, in this case 74 V. This additional power conversion stage can be used to regulate the intensity of the magnetic field produced by the on-board transmitter as input voltage dictates the amplitude of the current in the transmitting-coil.

The load in the experiment is a light-emitting diode (LED), representing a sensor positioned in environments with difficult access by other means, coupled with an receiver that uses a hybrid Class E rectifier with the same design used in the wirelessly powered drone, Section 7.2.8.



Figure 7.12: Photograph of a DJI Matrice 100 drone with an on-board HF-IPT transmitter.

### 7.2.14 Scaled-up Version of the Wirelessly Powered Drone

The development of the wirelessly powered drone demonstrator inspired the development of a scaled-up version of the HF-IPT system to wirelessly charge the on-board battery of a Matrice 100 UAV by DJI. A photograph of the system is shown in Fig. 7.13. The receive coil is a one-turn 400 mm round copper pipe conductor supported by the structure of the drone.



Figure 7.13: Photograph of the wirelessly charged drone taking off from the developed transmitter platform. The photograph was taken in the Maritime Autonomous Systems Trials and Evaluation Centre located at Turnchapel Wharf, Plymouth, UK, on the 11th of June 2019.

The system was designed to deliver 100 W to the battery (equivalent to the plug-in charger that comes with the drone). It operates at 6.78 MHz and the transmitter is based on Inverter II: Class EF, tuned for the two-turn copper pipe transmit coil.

This is an ongoing project of the Wireless Power Lab at Imperial College London.

## 7.3 The Wirelessly Charged E-Scooter

An alternative application that benefits from HF-IPT is the micro e-mobility market, and here an electric scooter wireless charging solution is described. This section demonstrates the design and construction of a 6.78 MHz IPT system with air-core coils, integrated in an

electric scooter, whose on-board battery can be charged at a maximum power of 100 W. With respect to the magnetics, integration challenges arise due to the proximity of the receive coil with the steel frame of the scooter, which significantly changes the coil link parameters from those calculated and observed in air. The design of the power electronic circuits at both ends of the system are detailed and explained. The battery of the electric scooter was able to be charged with a mains-to-battery efficiency of 65.6 % with an air-gap of 17 cm, where most of the losses are found in the battery charger. A losses distribution diagram of the entire system is included. Simulations of human exposure to the electromagnetic fields demonstrate that the designed system is well within the general exposure limits given by ICNIRP.

### **7.3.1 Introduction to Wireless Power to an E-Scooter**

Electric scooters can greatly benefit from wireless charging, as it eliminates direct interaction between the user and the charging mechanism, making such solution potentially more convenient, more reliable and safer.

WPT systems operating with air-core coils at megahertz frequencies have been demonstrated in a variety of sources [31, 40, 48, 154]. However, in many real-life applications, the receive coil is likely to be in proximity to another object: it may be embedded in a mobile phone or mounted under an electric vehicle. Therefore, the integration of IPT system with air-core coils into physical products can lead to design parameters that are different to those of a system in free space, due to the dielectric and magnetic properties of the surrounding environment.

This system demonstrates a 100 W application in which the receive coil is situated fairly close to the steel frame of the electric scooter, and shows how the system design parameters and coil characteristics in a scooter differ from those in air. The design and integration of the coils and power electronics in the scooter are described, the compliance to safety limits on human electromagnetic field exposure are evaluated, and experimental

results of the wireless charging system operating in situ are presented.

### 7.3.2 Designed System Overview

The electric scooter used in this work is the E-Rider Model 18 [155] (Fig. 7.14), which has a 48 V, 24 A h rechargeable Li-ion battery located inside the seat compartment. It comes with a 54.6 V, 2 A off-board battery charger, which specifies a universal input voltage: 100 V to 240 V at 50 or 60 Hz. The HF-IPT system was designed to integrate this charger,



Figure 7.14: E-Rider Model 18 Electric Scooter.

which can also be fed with a DC voltage. However, to make the system work, the rectifier must provide a  $V_o$  within the range required by the charger.

### 7.3.3 HF-IPT System Architecture and Power Throughput Control for the E-Scooter

The proposed HF-IPT charging system block diagram can be seen in Fig. 7.15. The front-end converter represents the power transformation from the mains to  $V_{dc}$ . This can be a mains-rectifier and a power factor correction (PFC) stage, and it can be simply a diode bridge, in case the system architecture proposed in Chapter 5 is implemented.



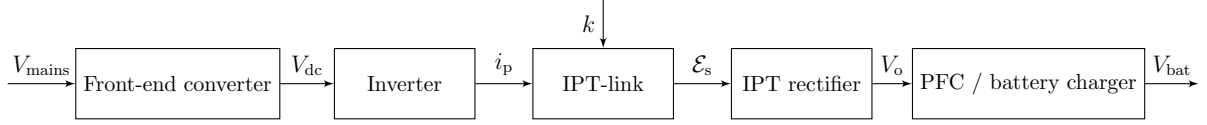


Figure 7.15: E-Rider HF-IPT system block diagram.

### 7.3.4 Design of an Inductive Link for Wireless Charging of an E-Scooter

The transmit and receive coils consist of standard PCBs with 1 oz copper on a 1.6 mm thick FR4 substrate. PCB coils are flat and planar, so they can be incorporated into the electric scooter and onto the road without reducing the ground clearance of the scooter significantly.

The receive coil is a 2-turn square PCB coil with an outer dimension of 230 mm, a trace width of 14 mm and a turn spacing of 14 mm, as shown in Fig. 7.16 (a). The size of the receiver coil was chosen to fit the flat portion of the underside of the scooter (250 mm width) and to avoid the side stand mount comprising a solid metal plate.

The transmit coil is a 2-turn square PCB coil with an outer dimension of 320 mm, a trace width of 20 mm and a turn spacing of 20 mm, as shown in Fig. 7.16 (b). It is greater in size than the receive coil to allow for better misalignment tolerance by having wider magnetic flux coverage, but not too large so as to reduce the coupling factor due to size asymmetry.

The inductance and resistance of the PCB coils were measured using the impedance analyser. The results are presented in Table 7.4. For the receive coil, impedance measurements were taken in air and underneath the scooter in order to evaluate the effect of the metal frame on the coil characteristics. The presence of the frame causes the inductance of the receive coil to decrease by 4.5 %; this is due to the eddy currents in the metal frame induced by a magnetic field that is in anti-phase to that generated by the receive coil current. The resistance of the receive coil also increased by 91 % as a result of eddy currents and magnetic losses in the frame.

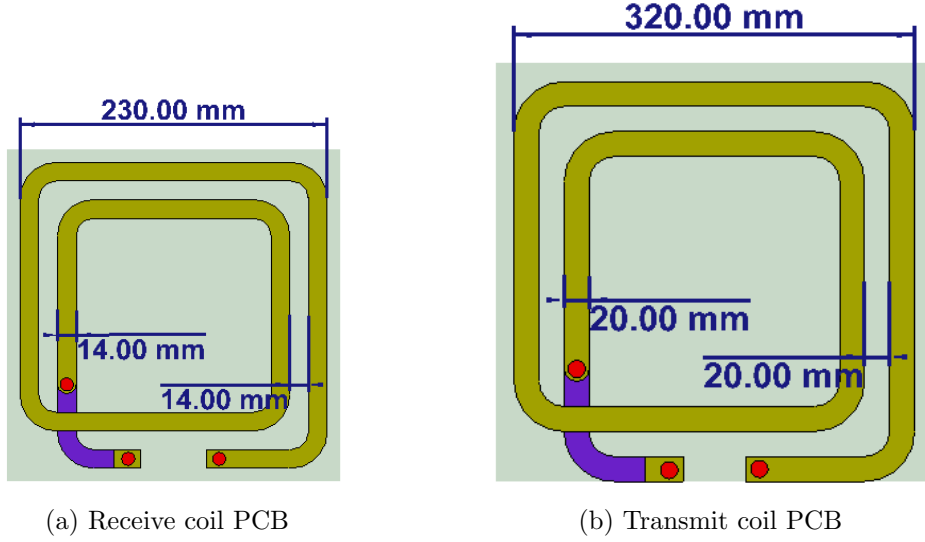


Figure 7.16: Coils diagrams for the wireless charging E-scooter.

Table 7.4: Measurements of coils designed for the wireless charger of an E-scooter

| Parameter              | Measurement          |
|------------------------|----------------------|
| $L_p$                  | 1.82 $\mu\text{H}$   |
| $R_p$                  | 189 $\text{m}\Omega$ |
| $L_{s,\text{air}}$     | 1.34 $\mu\text{H}$   |
| $R_{s,\text{air}}$     | 164 $\text{m}\Omega$ |
| $L_{s,\text{scooter}}$ | 1.28 $\mu\text{H}$   |
| $R_{s,\text{scooter}}$ | 313 $\text{m}\Omega$ |

The clearance between the bottom of the scooter and the ground is 170 mm when the electric scooter is on its centre stand. At this distance,  $k$  was measured as proposed in Section 4.2.4. In air, the mutual inductance is 137 nH and  $k$  is 8.8%. With the receiver coil mounted underneath the scooter, the mutual inductance falls to 119 nH and  $k$  drops to 7.8%, due to the presence of metal and metallic loops in the frame.

### 7.3.5 Simulations of Electromagnetic Exposure

The wireless charger for the electric scooter must consider safety limits on human exposure to electromagnetic fields, as it is possible for a person to be situated very close to the system. Since the frequency of operation is 6.78 MHz, basic restrictions from both ICNIRP 1998 [21] (for thermal effects, SAR) and ICNIRP 2010 [22] (for non-thermal

effects, internal E-field) are relevant.

Electromagnetic simulations were performed with CST Microwave Studio's frequency domain solver with the Nelly human body model. The primary coil current was set to 7.45 A, whilst the secondary coil current was 6.00 A; these current amplitudes enable charging of the battery at 100 W. Due to the use of secondary resonance, the coil currents are 90° out of phase. Two scenarios were considered for safety evaluation: Nelly standing with the feet 50 mm from the transmit pad and Nelly lying down 50 mm from the transmit pad (pad centre aligned with chest).

Fig. 7.17 (a) shows the SAR plot of Nelly in the standing position. In this case, the maximum localised SAR (averaged over 10 g of tissue) was  $0.0065 \text{ W kg}^{-1}$  at the left ankle, which is less than the general public exposure limit of  $4 \text{ W kg}^{-1}$  for the limbs. In addition, the 99<sup>th</sup> percentile internal E-field value is  $5.0 \text{ V m}^{-1}$  which is below the general public exposure limit of  $915 \text{ V m}^{-1}$ .

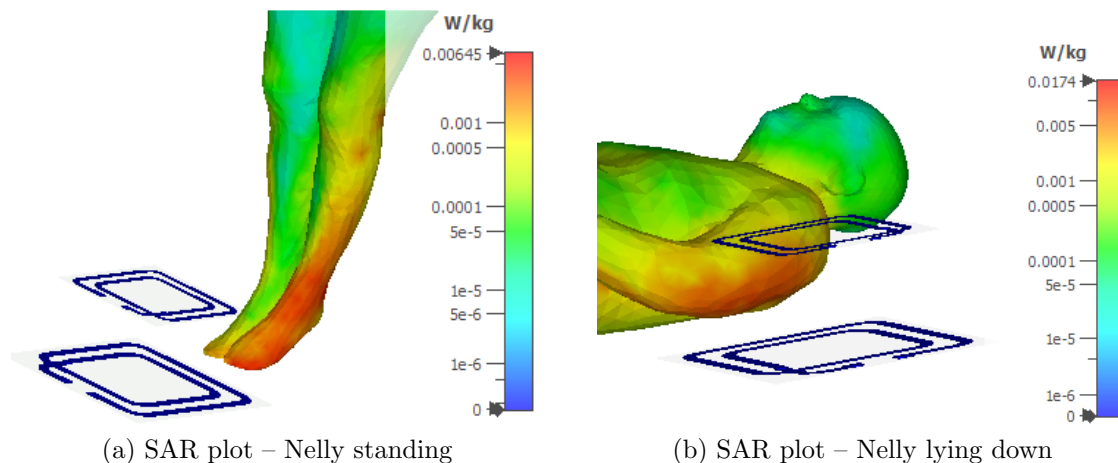


Figure 7.17: SAR simulations results for the wirelessly charged E-scooter in critical cases.

Fig. 7.17 (b) shows the SAR plot of Nelly in the lying (supine) position. In this case, the maximum localised SAR was  $0.017 \text{ W kg}^{-1}$  at the left lower arm near the elbow, which is less than the general public exposure limit of  $4 \text{ W kg}^{-1}$  for the limbs. Moreover, the 99<sup>th</sup> percentile internal E-field value is  $9.6 \text{ V m}^{-1}$ , which is below the general public exposure limit of  $915 \text{ V m}^{-1}$ .

### 7.3.6 HF-IPT Inverter and Rectifier for the E-Scooter Wireless Charger

#### Design of the 6.78 MHz Inverter

The design of the inverter is based on Inverter II: Class EF, which uses load-independent tuning for Class EF inverters [73]. The design of this single-switch resonant inverter is developed in Section 3.2.2. The circuit topology of the inverter is presented in Fig. 7.18 (a), and the components of the inverter, tuned for the proposed transmit coil, are presented in Table 7.5.

#### Design of the HF-IPT Rectifier

The rectifier was designed using a two-stage voltage-multiplier current-driven Class D rectifier, similar to Rectifier II: VM Class D as shown in Fig. 7.18(b), to increase the output voltage, and provide the required voltage to the battery charger that comes with the electric scooter, which ranges from 100 V to 240 V in root mean square (RMS) at 50 or 60 Hz. The components were selected such that  $C_s$  would resonate the receive coil ( $L_s$  in Fig. 7.18). Thus, it had to be re-tuned once the coil was positioned at the bottom of the scooter, as a result of changes in self and mutual inductance once the system was integrated in the application. The components of the rectifier are presented in Table 7.5.

### 7.3.7 Measurements on the E-Scooter Demonstrator

Experiments were performed on the prototype electric scooter wireless charger prototype shown in Fig. 7.19. Three experiments are described: A DC-DC HF-IPT system in air, a DC-DC HF-IPT system mounted on the electric scooter, and a DC-DC HF-IPT system charging the battery of the electric scooter in situ. The system was also verified using the system architecture proposed in Chapter 5. The results are summarised in Table 7.6.

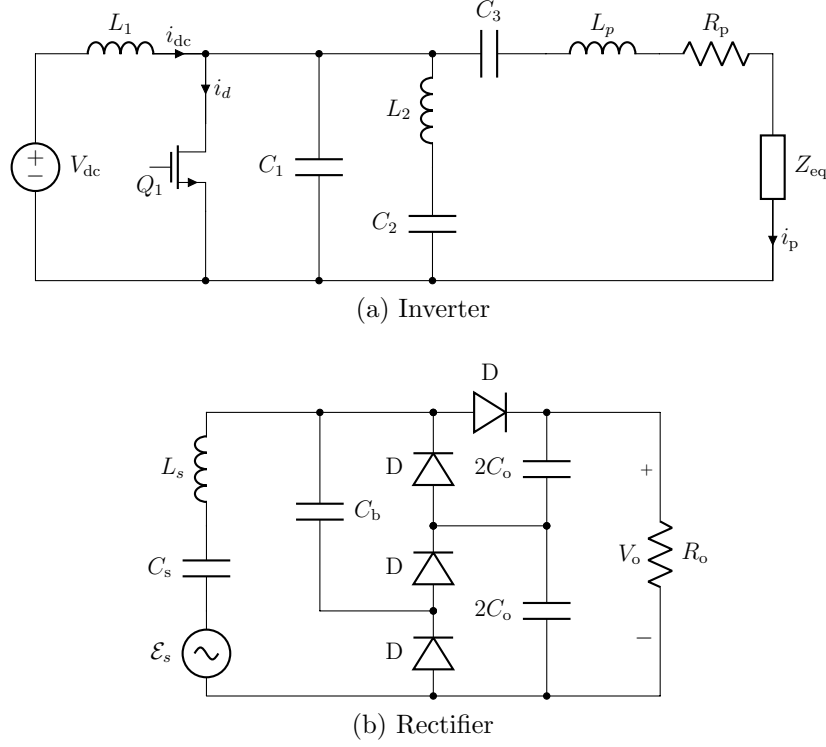


Figure 7.18: Circuit topology of the inverter and rectifier implemented for the wirelessly charged E-scooter.

Table 7.5: Components of the inverter and rectifier for the wireless charger for the E-scooter

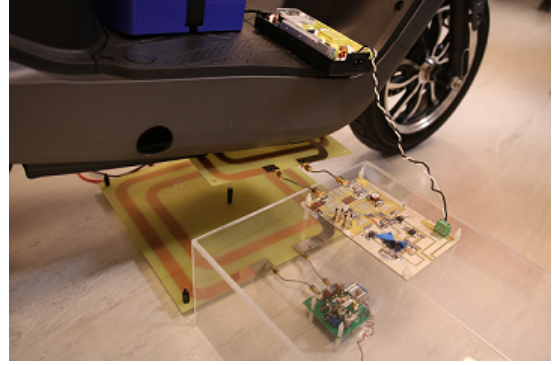
| Component          | Value                               | Description            |
|--------------------|-------------------------------------|------------------------|
| $C_1$ (pF)         | $439 + C_{oss}$                     | Vishay QUAD HIFREQ     |
| $C_2$ (pF)         | 418                                 | Vishay QUAD HIFREQ     |
| $C_3$ (pF)         | 390                                 | Vishay QUAD HIFREQ     |
| $L_1$ ( $\mu$ H)   | 88                                  | Würth Elektronik WE-PD |
| $L_2$ (nH)         | 404                                 | Coilcraft 2014VS       |
| $L_p$ & $L_s$ (nH) | PCB coils (Table 7.4)               |                        |
| $Q_1$              | GS66504B (650 V, 15 A) GaN FET      |                        |
| $C_s$ (pF)         | 430                                 | Vishay QUAD HIFREQ     |
| $C_b$ (nF)         | 100                                 | Vishay QUAD HIFREQ     |
| $C_o$ (nF)         | 500                                 | Vishay QUAD HIFREQ     |
| $D_1 - D_4$        | C3D02060E-TR (600 V, 2 A) SiC diode |                        |

## IPT System Operating in Air with a Dummy Load

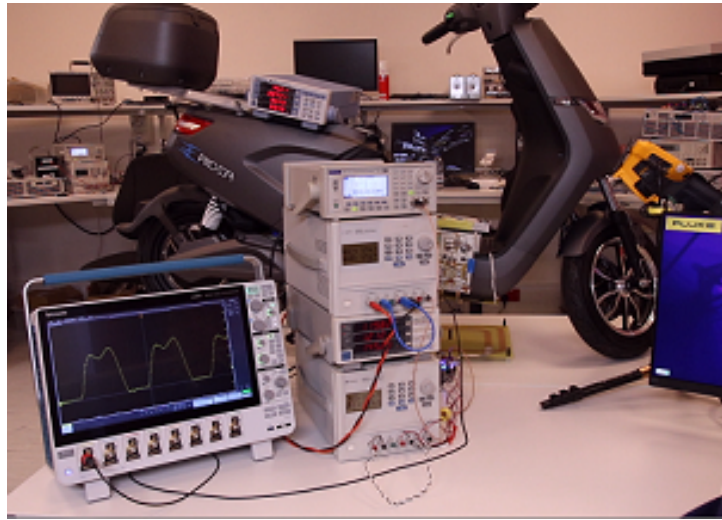
The wireless charging system was initially tested in air with a dummy load at the output of the rectifier. The receiver coil was tuned with a series capacitance to resonate out its inductance as measured in air. In summary, a IPT system DC to DC efficiency ( $\eta_{dc-o}$ ) of 80.0 % was achieved; detailed results are shown in Table 7.6 (‘IPT in air’ column).



(a) E-scooter's chassis



(b) IPT circuits



(c) Experimental setup

Figure 7.19: Photographs of the E-scooter wireless charger prototype.

Table 7.6: Experimental results of the E-scooter wireless charging system

| Measurement          | IPT in air | IPT in scooter | Charging battery |
|----------------------|------------|----------------|------------------|
| $R_o$ [ $\Omega$ ]   | 243        | 194            | -                |
| $V_{dc}$ [V]         | 121        | 121            | 120              |
| $V_{mains}$ [V, RMS] | -          | -              | 120              |
| $I_{dc}$ [A]         | 1.03       | 1.11           | 1.21             |
| $P_{dc}$ [W]         | 125        | 133            | 145              |
| $V_o$ [V]            | 156        | 139            | -                |
| $V_{bat}$ [V]        | -          | -              | 45.7             |
| $I_o$ [A]            | 0.643      | 0.717          | -                |
| $I_{bat}$ [A]        | -          | -              | 2.08             |
| $P_o$ [W]            | 100        | 99.7           | -                |
| $P_{bat}$ [W]        | -          | -              | 95.1             |
| $\eta_{dc-o}$        | 80.0 %     | 75.0 %         | -                |
| $\eta_{mains-bat}$   | -          | -              | 65.6 %           |

## IPT System Operating in Scooter with a Dummy Load

The receive coil was mounted underneath the scooter for the second experiment. Due to the change in coil inductance when placed under the scooter, the series capacitance at the receiver needed to be re-tuned so that it resonated with the coil inductance when positioned under the scooter. As the transmit coil current amplitude is constant with the proposed inverter, the change in  $k$  and  $R_{eq}$  required a decrease in the dummy-load resistance to regain 100 W at the output. Table 7.6 (‘IPT in scooter’ column) shows the results of the experiment; the  $\eta_{dc-o}$  in this scenario was measured at 75.0 %.

## IPT System in Scooter with Battery Charger and Li-ion Battery

For the final experiment, the dummy load at the output of the rectifier was replaced by the battery charger which in turn was connected to the Li-ion battery. Table 7.6 (‘Charging battery’ column) shows the results of the experiment; the IPT system mains AC to regulated DC efficiency ( $\eta_{mains-reg}$ ) (which was congruent to the dc-to-battery efficiency) achieved was 65.6 %. The distribution of the losses in the end-to-end system, obtained from simulation, is shown in Fig. 7.20. The battery charger turned out to be 90 % efficient when separately measured without the HF-IPT link.

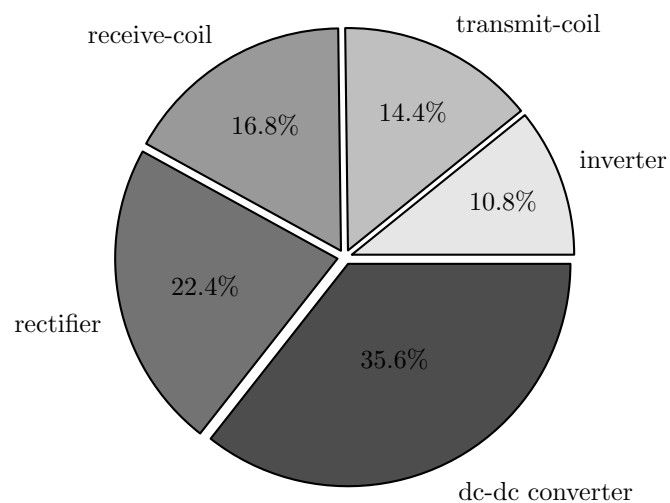


Figure 7.20: Distribution of losses in the wireless charged E-scooter from simulation.

### 7.3.8 IPT System in Scooter with Battery Charger and Li-ion Battery Powered from Mains AC

An additional experiment was performed using the same system but fed from a 120 V 60 Hz voltage source, integrating a diode bridge as the front-end conversion stage in Fig. 7.15, as opposed to a bench power supply that gives a regulated DC voltage. HF-IPT systems can not only be fed from a regulated DC source, but also from a voltage source with the shape of the absolute value of a low frequency alternating current (AC) source, as proposed in Chapter 5. Active PFC is performed at the receive side in this type of system configuration. Single switch resonant inverters, such as Class E and Class EF, conveniently have a large inductance at the input, similar to that of a boost-converter. This inductance intrinsically filters the input current, thereby making such converters convenient for placement at the front end of an active PFC system.

Since the scooter's battery charger which was used in the experiments as the output voltage regulator allows an AC input voltage and is also designed to perform active PFC, the system did not require any modifications other than the integration of a diode bridge at the front end of the inverter.

The efficiency recorded for this solution was congruent with that having a DC-input voltage at 120 V. However, the AC-mains fed solution allows for one less power conversion stage since the end-to-end efficiency is calculated from  $V_{\text{mains}}$  in Fig. 7.15 and not from  $V_{\text{dc}}$ . The power factor at the source was measured to be higher than 0.99 in a set of experiments performed.

### 7.3.9 Scaled-up Version of the Wirelessly Charged E-scooter

A scaled-up version of the E-rider scooter was conducted using Inverter III: PP Class EF and Rectifier II: VM Class D. Bigger heatsinks were used for higher power throughput. A photograph of an experiment can be seen in Fig. 7.21. The Govecs S2.4 Electric Scooter



is being charged with 712.9 W at the input with a calculated efficiency of 85 %.



Figure 7.21: Photograph of the 600 W wireless charging experiment of the on-board battery of a Govecs S2.4 Electric Scooter.

This is an ongoing project of the Wireless Power Lab at Imperial College London.

## 7.4 Summary

With newly-developed resonant power converters and tuning methods, in addition to the advantageous features of wide-bandgap devices, lightweight and efficient solutions for WPT using HF-IPT are made possible. This chapter covers two design examples of end-to-end HF-IPT systems which consider coil design, circuit design and compliance with safety standards with respect to exposure to electromagnetic fields.



# Chapter 8

## Conclusion

This chapter discusses the general conclusions of the work presented in this thesis and summarises the main outcomes of the research.

### 8.1 Overview and Main Findings

This thesis showcases examples of high frequency inductive power transfer (HF-IPT) systems which successfully demonstrate wireless power to real applications. These demonstrators help validate the technology as an alternative to integrate wireless charging to low and mid-power commercial products. An overview and the main findings of this thesis are summarised below.

Some of the main challenges of designing an end-to-end HF-IPT system come from the interaction between the two ends of the system. There are challenges in extracting power from a voltage source to drive a transmit coil and there are challenges in extracting power from an induced electromotive force ( $\mathcal{E}$ ) on a receive coil to feed power to an electronic device. Formulating the interaction between the two ends of the system, and specifically between the two coils, is fundamental to design the system and make it work. This thesis formulates the basic theory of inductive power transfer (IPT) in a form which aims to

detach the design of the different stages of the system while considering the effects of actions that might take place if the system was integrated in an application, for example relative movement of the coils, load variations, or the interaction of the system with other objects in proximity to the coils. In the proposed model, the induced electromotive force on the receive coil ( $\mathcal{E}_s$ ) is represented as a voltage source dependent of the self-inductance of the coils, the transmit-coil current ( $i_p$ ), the transmit-coil current frequency ( $f_{i_p}$ ) and the coupling factor ( $k$ ); and similarly the reflected impedance on the transmit side ( $Z_{eq}$ ) is modelled with respect to the self-inductance of the coils, the input impedance of the IPT-rectifier ( $Z_{sec}$ ),  $f_{i_p}$  and  $k$ . If the self-inductance of the coils and  $f_{i_p}$  are considered constant, changes in load or in  $k$  can be represented as changes in  $Z_{eq}$  at the transmit side, and changes in  $i_p$  and  $k$ , can be represented as changes in  $\mathcal{E}_s$ , at the receive side.

The work involving the power electronic circuits at both ends of the system unveils inherent properties of a group of resonant power converters which can tackle challenges of performing HF-IPT using loosely coupled inductors. For example, the load-tolerant Class E and the load-independent Class EF inverters can feed a fairly constant current amplitude to the transmit coil throughout the reflected resistance range, and the resistance compression network (RCN) Class D rectifier can be designed to emulate the optimal load of the link with a compressed dependence of the load resistance.

An advantageous feature of utilising resonant converters is that the shape of the voltage and current waveforms of these circuits contain information about the tuning and loading conditions. Experiments are shown where the waveforms of an inverter under test allowed identifying  $Z_{eq}$ , and thus obtaining information of the receiver and the link, without any sort of communication with the receiver. These differences on the waveforms allowed implementing a machine learning algorithm to identify the patterns on the drain voltage waveform of the inverter which were used to estimate  $Z_{eq}$  with a high degree of accuracy.

A system architecture is introduced where a Class E or a Class EF inverter, if designed for a broad input voltage, can be fed from a mains source directly, not requiring a power factor correction (PFC) stage at the transmit side between the mains and the inverter. By

employing this system architecture, the design can take advantage of the inverter’s large input inductance to more easily shape the current drawn from the mains with a combined output voltage regulation and PFC stage at the receive side after the IPT rectifier.

A system-level analysis using probability theory describes a method of designing HF-IPT systems for variable  $k$ , thus taking into account movement between the coils. The formulation of motion as a probability distribution allows tuning the system taking into account which values of  $k$  are more likely to occur. This probability-based formulation also allows the coils to be designed based on motion and the probability distribution of  $k$ .

Additional experiments allowed the effects of introducing alternative media between the coils to be assessed and a study on the effects of corrosion on an air-core coil for HF-IPT was undertaken. The presence of three materials under test, representing layers of wireless power transfer (WPT) to or from a device buried underground: soil, water and a brick, were found to have a minimal effect on the high frequency inductive power transmission. With respect to corrosion of air-core coils, where the hypothesis in [31] describes that a thin layer of copper oxide on the coils was thought to be causing a large drop in the Q factors, was challenged. The work presented here demonstrates that in printed circuit board (PCB) coils at 13.56 MHz, corrosion does not significantly affect the Q factor unless the conduction path of the coil is heavily compromised: it is not a skin-type phenomenon.

The main findings of this thesis fed into the development of two HF-IPT systems that exemplify WPT to real-world applications.

## 8.2 Author’s Contribution

This thesis showcases two examples of end-to-end HF-IPT systems to demonstrate WPT into applications, that of a drone and an electric scooter. These demonstrators were constructed as collaborations among several members of the Wireless Power lab and

showcase some of the author's contributions of this thesis. The author's contributions are now described.

Some of the most important challenges in the field of HF-IPT are the development of the power electronics circuits at both ends of the system. In this thesis, the first main contribution is the implementation of a RCN rectifier topology to achieve an IPT system AC load ( $R_{ac}$ ) that converges to the optimal load formulated in (2.17), with a compressed dependence of the load resistance. The combination of this rectifier, a load independent Class EF inverter and two air-core copper-pipe coils, allowed a peak efficiency of 88 % at 6.78 MHz and 83 % at 13.56 MHz to be achieved at a coil separation higher than a coil radius. To the author's knowledge this is the highest DC-DC efficiency ever achieved with the corresponding coil separations at such frequencies. The experiments using these circuits also show, for the first time, a HF-IPT system with a power capability of 150 W operating at variable coupling lower than 12 % at 6.78 MHz and lower than 5 % at 13.56 MHz.

The second main contribution of this work was the introduction of a system architecture for HF-IPT systems where the current-driven Class E and Class EF inverters, commonly employed for megahertz switching power conversion, are fed directly from a mains voltage supply, allowing the elimination of a power conversion stage at the front end of the system. This type of system architecture performs PFC at the receive side, therefore a PFC stage at the receiver after the HF-IPT rectifier can regulate the output voltage whilst causing the system input to be of high power factor and low distortion.

Considering that motion is one of the most attractive features of WPT, addressing the challenge of designing HF-IPT systems for variable  $k$  is paramount. The third main contribution of this thesis consists of developing a probability-based model where the systems are not only designed for variable  $k$ , but the motion pattern between the coils is taken into account, allowing tuning the system for maximum energy efficiency with respect to  $k$  and the distribution of  $k$  throughout the charging time. Furthermore, the model can help evaluate the system with respect to the relative position of the coils and

how motion affects the distribution of  $k$ . This allows high-level control to dictate the range of motion at which the system transfers power, and also the shape of the coils can be redesigned to optimise the distribution of  $k$  for a given motion profile.

Various other contributions were also made, including an impedance estimation model using machine learning, a study of the effects produced by alternative media between two coils performing HF-IPT and the evaluation of corrosion on a PCB coil designed for HF-IPT.

The HF-IPT programme is a collaborative effort involving the members of the Wireless Power Lab at Imperial College London. The work in this thesis was completed from October 2015 until September 2019, and is the work of the author and the following collaborations: The work presented in Section 4.7 was a collaboration with Dr. Kyra Sedransk-Campbell from the Department of Chemical Engineering and Mr. Epameinondas Skountzos of the Materials Department of Imperial College London. Prof. Paul D. Mitcheson and Dr. David C. Yates supervised this PhD and lead the programme. Dr. Samer Aldhaher introduced the load independent Class EF inverter and was in charge of designing the inverter boards. Dr. George Kkelis introduced the hybrid Class E rectifier and collaborated in the development of some of the rectifiers used in this thesis. Dr. Christopher H. Kwan generated the electromagnetic exposure simulations, and he led the magnetic design of the wireless charging solution for the E-scooter. Mr. Lingxin Lan programmed the machine learning algorithms to estimate  $\mathbf{Z}_{eq}$  using the transmitter waveforms.

## 8.3 Patents, Publications, Awards and Outreach

The patent applications, journal and conference papers, awards and outreach activities of this research are listed in this section.

### 8.3.1 Patent Applications

- Juan M. Arteaga, David C. Yates and Paul D. Mitcheson, “Transmitter and Receiver Circuitry for Power Converter Systems”. Filing Date: 4 July 2018, International Application Number: PCT/GB2019/051892. This work is listed as reference [122].

### 8.3.2 Journal Publications as Lead Author

- IEEE Transactions on Power Electronics: A paper titled ”Dynamic Capabilities of Multi-MHz Inductive Power Transfer Systems Demonstrated with Batteryless Drones” and authored by: Juan M. Arteaga, Samer Aldhafer, George Kkelis, Christopher H. Kwan, David C. Yates and Paul D. Mitcheson, was accepted in September 2018 and published in June 2019. This paper includes active content: a 105s video file which can be found in IEEEExplore. This publication is listed as reference [84].
- IEEE Transactions on Power Electronics: A paper titled “Multi-MHz IPT systems for variable coupling” and authored by: Juan M. Arteaga, Samer Aldhafer, George Kkelis, David C. Yates and Paul D. Mitcheson, was accepted in October 2017 and published in September 2018. This publication is listed as reference [53].

### 8.3.3 Conference Presentations and Publications as Lead Author

- IEEE Applied Power Electronics Conference 2019, Anaheim, California, United States of America: A paper titled “A Multi-MHz Wireless Power Transfer System With Mains Power Factor Correction Circuitry on the Receiver”, and authored by: Juan M. Arteaga, Samer Aldhafer, David C. Yates and Paul D. Mitcheson, was



presented at the technical lecture session titled “Optimisation of Wireless Power Transfer Systems”. A travel grant, sponsored by the Power Sources Manufacturers Association, was awarded to support the presentation of this work. This publication is listed as reference [121].

- IEEE PELS Workshop on Emerging Technologies: Wireless Power (WoW) 2018, Montréal, Canada: A paper titled “Probability-based optimisation for a multi-MHz IPT system with variable coupling”, and authored by: Juan M. Arteaga, George Kkelis, Samer Aldhaher, David C. Yates and Paul D. Mitcheson, was presented at the technical lecture session titled “Performance Optimisation”. In addition, this work was presented in the form of a poster for the Student Paper Competition, where it was awarded the first prize. This publication is listed as reference [133].
- IEEE MTT-S Wireless Power Transfer Conference 2018, Montréal, Canada: A paper titled “A multi-MHz IPT-link developed for load characterisation at highly variable coupling factor”, and authored by: Juan M. Arteaga, Lingxin Lan, Samer Aldhaher, George Kkelis, David C. Yates and Paul D. Mitcheson, was presented at the technical lecture session titled “HF innovative circuits”. This publication is listed as reference [95].
- IEEE Southern Power Electronics Conference 2016, Auckland, New Zealand: A paper titled “Design of a 13.56 MHz IPT system optimised for dynamic wireless charging environments”, and authored by: Juan M. Arteaga, Samer Aldhaher, George Kkelis, David C. Yates and Paul D. Mitcheson, was presented at the technical lecture session titled “New Developments in WPT Systems”. This publication is listed as reference [55].
- IEEE MTT-S Wireless Power Transfer Conference 2016, Aveiro, Portugal: A paper titled “A current driven Class D rectifier with a resistance compression network for 6.78 MHz IPT systems”, and authored by: Juan M. Arteaga, George Kkelis,

David C. Yates and Paul D. Mitcheson, was presented in the poster session. This publication is listed as reference [54].

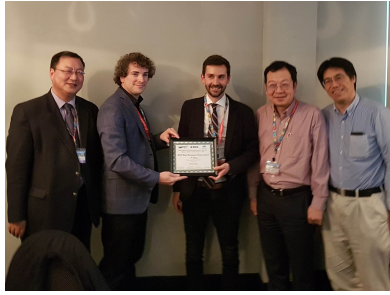
### 8.3.4 Conference Publications as Co-Author

- E. Skountzos, J. M. Arteaga, E. Hadjittofis, D. C. Yates, K. L. Sedransk-Campbell and P. D. Mitcheson, “A 13.56 MHz Inductive Power Transfer System Operating with Corroded Coils”, IEEE PELS Workshop on Emerging Technologies: Wireless Power Transfer (WoW), 2019, London, U.K., pp. 1-5. This publication is listed as reference [97].
- L. Lan, J. M. Arteaga, D. C. Yates and P. D. Mitcheson, “A Reflected Impedance Estimation Technique for Inductive Power Transfer”, IEEE PELS Workshop on Emerging Technologies: Wireless Power Transfer (WoW), 2019, London, U.K., pp. 1-4. This publication is listed as reference [96].
- C. H. Kwan, J. M. Arteaga, D. C. Yates and P. D. Mitcheson, “Design and Construction of a 100 W Wireless Charger for an E-Scooter at 6.78 MHz”, IEEE PELS Workshop on Emerging Technologies: Wireless Power Transfer (WoW), 2019, London, U.K., pp. 1-5. This publication is listed as reference [144].
- P. D. Mitcheson, G. Kkelis, S. Aldhafer, J. M. Arteaga, D. C. Yates, D. Boyle, E. M. Yeatman, “Power Electronics for Wireless Power Delivery in Synthetic Sensor Networks”, Journal of Physics: Conference Series, July, 2018, pp. 1-2. This publication is listed as reference [147].
- L. Lan, N. M. Ting, S. Aldhafer, G. Kkelis, C. H. Kwan, J. M. Arteaga, D. C. Yates, P. D. Mitcheson, “Foreign Object Detection for Wireless Power Transfer,” 2018 2nd URSI Atlantic Radio Science Meeting (AT-RASC), Meloneras, 2018, pp. 1-2. This publication is listed as reference [98].

- P. D. Mitcheson, D. Boyle, G. Kkelis, D. Yates, J. M. Arteaga, S. Aldhafer, E. Yeatman “Energy-autonomous sensing systems using drones,” 2017 IEEE SENSORS, Glasgow, 2017, pp. 1-3. This publication is listed as reference [146].
- G. Kkelis, S. Aldhafer, J. M. Arteaga, D. C. Yates and P. D. Mitcheson, “Hybrid Class-E synchronous rectifier for wireless powering of quadcopters,” 2017 IEEE Wireless Power Transfer Conference (WPTC), Taipei, 2017, pp. 1-4. This publication is listed as reference [83].
- S. Aldhafer, P. D. Mitcheson, J. M. Arteaga, G. Kkelis and D. C. Yates, “Light-weight wireless power transfer for mid-air charging of drones,” 2017 11th European Conference on Antennas and Propagation (EUCAP), Paris, 2017, pp. 336-340. This publication is listed as reference [145].

### 8.3.5 Awards

- First prize in the Student Hardware Demonstration Competition at the IEEE Wireless Power Transfer Conference 2018, Montréal, QC, Canada, “The Wireless Power Drone”.
- First prize in the Student Paper Competition: J. M. Arteaga, G. Kkelis, S. Aldhafer, D. C. Yates and P. D. Mitcheson, “Probability-Based Optimisation for a Multi-MHz IPT System with Variable Coupling,” 2018 IEEE PELS Workshop on Emerging Technologies: Wireless Power Transfer (Wow), Montréal, QC, 2018, pp. 1-5.
- First prize in the Centre for Power Electronics UK Future Power Challenge 2016, Nottingham, UK. The project title was: “A 13.56 MHz Load Independent Inductive Power Transfer System with 73% DC-DC Efficiency” and was authored by Juan M. Arteaga, Samer Aldhafer, George Kkelis, David C. Yates and Paul D. Mitcheson.



(a) WPTC 2018



(b) WoW 2018



(c) CPE 2016

Figure 8.1: Photographs of the awards presentations.

### 8.3.6 Outreach

#### Industrial Involvement and Commercial Impact

- Founding Member of Imperial College London - Wireless Power Lab spin-out company Bumblebee Power Ltd.
- Partnership with the French multinational company Thales Group in displaying the wireless charging capabilities for autonomous drones in maritime surveillance systems.
- Partnership with the Canadian gallium nitride transistor manufacturer company GaN Systems in the exhibition of the Wirelessly Powered Drone at the Applied Power Electronics Conference (APEC) 2017 and Power Conversion and Intelligent Motion (PCIM) Europe 2017.

#### Conference Organisation

- Co-chair the Local Organising Committee of the IEEE Wireless Power Week 2019, London.
- Member of the Local Organising Committee of the Centre for Power Electronics UK in the New Scientist Live 2016.

- Member of the Local Organising Committee of the Centre for Power Electronics UK in the Cheltenham Science Festival 2016.

## Peer Reviewing

- IEEE Transactions on Power Electronics: nine manuscript reviews have been submitted.
- IEEE Transactions on Industrial Electronics: six manuscript reviews have been submitted.
- IEEE Journal of Emerging and Selected Topics in Power Electronics: five manuscript reviews have been submitted.
- IEEE Wireless Power Week 2019: Member of the Technical Program Committee, reviewing over twenty manuscripts, and co-chair of a speaker session on high frequency inductive power transfer.

## 8.4 Suggestions for Future Work

This thesis has revealed several potential areas of research for further investigation which are summarised below.

There are properties, inherent of resonant megahertz-switching power electronic circuit topologies, that have desirable features for HF-IPT, some of which were employed in the development of the WPT systems in this thesis. The state-space-equations which describe these circuits allow design for specific solutions. Thus, investigating the possibilities of including additional features that are crucial for the commercial integration of this technology, as for example minimising the total harmonic distortion (THD) on the transmit-coil current, should be further investigated. Alternative topologies to the ones presented in this thesis can also be reevaluated as the transistor technologies move

forward. Most notably, the Class D inverters with gallium nitride (GaN) transistors have been recently demonstrated at high efficiencies ( $>85\%$ ) and high power throughput (500 W) [43]. There is also significant room for improvement in the auxiliary power converters at both ends of the system, before the inverter and after the high frequency rectifier as the requirements of these can differ from standard solutions designed to operate from a traditional voltage source. Finally, there are also practical challenges which need to be addressed in the power electronic circuits design, for these to be feasible to be integrated in commercial applications, as for example, electrical and mechanical robustness, packaging, thermal management of the boards and cost.

Most of the experimental work presented in this thesis was conducted using PCB coils, which were designed and built, not necessarily to achieve the highest efficiency, but mainly to allow for a precise reproduction of the experiments. A commercial HF-IPT solution would require significant improvement in the design and construction of the coils so that high IPT link efficiency ( $\eta_{\text{link}}$ ) is achieved with low-cost coils embedded in an application. The latter also poses challenges with respect to shielding the magnetic flux of IPT systems with air-core coils. Different shapes and materials in the conductor that conforms the coils can also be further explored to obtain different coupling distributions, especially considering the possibility of relative motion of the coils in dynamic applications.

The techniques proposed to tune a receiver using an impedance analyser and therefore estimate  $\mathbf{Z}_{\text{eq}}$  using machine learning algorithms, solely by observing and comparing the waveforms of the inverter unveils many characterisation possibilities that can be further explored. For example, the experimental characterisation of the  $\mathbf{Z}_{\text{eq}}$  produced by different tunings and different topologies of rectifiers, also the effects of integrating other type of materials between the coils to represent different IPT environments. The technique can also be used for foreign object detection if a more cost-effective mechanism of obtaining the data is achieved.

The proposed energy-efficiency formulation using the probability distribution of  $k$  to characterise motion points out key parameters that can be optimised for energy-efficiency.

First, the range of  $k$  at which the system operates can be selected not only based on the capabilities of the circuits but depending on the performance at the different values of  $k$  and accounting for how probable these values of  $k$  are. Second, the coils can be designed to achieve desirable features in the distribution of  $k$  for a particular motion profile. For example, the shape of the conductors of the coils can be optimised for maximum  $k$  at the most probable position, or a shape can be found where the change of  $k$  is minimised for the given movement profile. And third, the model can help determine at which  $k$  it is best to optimise the inverter, the rectifier and the link.





# References

- [1] M. Faraday, “Experimental researches in electricity,” *Philosophical transactions of the Royal Society of London*, vol. 122, pp. 125–162, 1832.
- [2] B. Dibner, “History of electrical engineering Andre Marie Ampere,” *IEEE Power Engineering Review*, vol. PER-4, no. 2, pp. 15–16, Feb 1984.
- [3] T. K. Sarkar, R. Mailloux, A. A. Oliner, M. Salazar-Palma, and D. L. Sengupta, *History of wireless*. John Wiley & Sons, 2006, vol. 177.
- [4] M. Faraday, “Lviii. on the physical character of the lines of magnetic force,” *The London, Edinburgh, and Dublin Philosophical Magazine and Journal of Science*, vol. 3, no. 20, pp. 401–428, 1852.
- [5] J. C. Maxwell, “On Faraday’s lines of force,” *Transactions of the Cambridge Philosophical Society*, vol. 10, p. 27, 1864.
- [6] —, “Xxv. On physical lines of force: Part I. The theory of molecular vortices applied to magnetic phenomena,” *The London, Edinburgh, and Dublin Philosophical Magazine and Journal of Science*, vol. 21, no. 139, pp. 161–175, 1861.
- [7] —, “A dynamical theory of the electromagnetic field,” *Philosophical transactions of the Royal Society of London*, vol. 155, pp. 459–512, 1865.
- [8] Y. Hu, X. Zhang, J. Yang, and Q. Jiang, “Transmitting electric energy through a metal wall by acoustic waves using piezoelectric transducers,” *IEEE Trans. Ultrason., Ferroelectr., Freq. Control*, vol. 50, no. 7, pp. 773–781, July 2003.

- [9] A. Denisov and E. Yeatman, “Ultrasonic vs. inductive power delivery for miniature biomedical implants,” in *2010 International Conference on Body Sensor Networks*, June 2010, pp. 84–89.
- [10] L. Summerer and O. Purcell, “Concepts for wireless energy transmission via laser,” *Europeans Space Agency (ESA)-Advanced Concepts Team*, 2009.
- [11] L. Xie, Y. Shi, Y. T. Hou, and A. Lou, “Wireless power transfer and applications to sensor networks,” *IEEE Wireless Communications*, vol. 20, no. 4, pp. 140–145, August 2013.
- [12] Z. Ding, C. Zhong, D. W. K. Ng, M. Peng, H. A. Suraweera, R. Schober, and H. V. Poor, “Application of smart antenna technologies in simultaneous wireless information and power transfer,” *IEEE Communications Magazine*, vol. 53, no. 4, pp. 86–93, April 2015.
- [13] W. C. Brown, “Experiments involving a microwave beam to power and position a helicopter,” *IEEE Trans. on Aerospace and Electron. Systems*, vol. AES-5, no. 5, pp. 692–702, Sept 1969.
- [14] J. Dai and D. Ludois, “A survey of wireless power transfer and a critical comparison of inductive and capacitive coupling for small gap applications,” *IEEE Trans. on Power Electron.*, vol. 30, no. 11, pp. 6017–6029, Nov 2015.
- [15] H. Zhang, F. Lu, H. Hofmann, W. Liu, and C. Mi, “An lc compensated electric field repeater for long distance capacitive power transfer,” *IEEE Trans. Ind Appl.*, vol. PP, no. 99, pp. 1–1, 2017.
- [16] F. Lu, H. Zhang, and C. Mi, “A review on the recent development of capacitive wireless power transfer technology,” *Energies*, vol. 10, no. 11, p. 1752, 2017.
- [17] B. Regensburger, A. Kumar, S. Sinha, and K. Afridi, “High-performance 13.56-mhz large air-gap capacitive wireless power transfer system for electric vehicle charging,” in *2018 IEEE 19th Workshop on Control and Modeling for Power Electronics (COMPEL)*. IEEE, 2018, pp. 1–4.

- [18] B. Regensburger, A. Kumar, S. Sinha, J. Xu, and K. K. Afridi, “High-efficiency high-power-transfer-density capacitive wireless power transfer system for electric vehicle charging utilizing semi-toroidal interleaved-foil coupled inductors,” in *2019 IEEE Applied Power Electronics Conference and Exposition (APEC)*. IEEE, 2019, pp. 1533–1538.
- [19] C. H. Kwan, J. Lawson, D. C. Yates, and P. D. Mitcheson, “Position-insensitive long range inductive power transfer,” *Journal of Physics: Conference Series*, vol. 557, p. 012053, nov 2014.
- [20] R. Bosshard, J. W. Kolar, J. Mühlethaler, I. Stevanović, B. Wunsch, and F. Canales, “Modeling and  $\eta$  -  $\alpha$  -pareto optimization of inductive power transfer coils for electric vehicles,” *IEEE Trans. Emerg. Sel. Topics Power Electron.*, vol. 3, no. 1, pp. 50–64, March 2015.
- [21] International Commission on Non-Ionizing Radiation Protection and others, “Guidelines for limiting exposure to time-varying guidelines for limiting exposure to time-varying electric, magnetic and electromagnetic fields,” *Health Physics*, vol. 74, pp. 494–522, 1998.
- [22] —, “Guidelines for limiting exposure to time-varying electric and magnetic fields (1 Hz to 100 kHz),” *Health physics*, vol. 99, no. 6, pp. 818–836, 2010.
- [23] “IEEE standard for safety levels with respect to human exposure to radio frequency electromagnetic fields, 3 kHz to 300 GHz,” *IEEE Std C95.1-2005 (Revision of IEEE Std C95.1-1991)*, pp. 1–238, April 2006.
- [24] A. Christ, M. G. Douglas, J. M. Roman, E. B. Cooper, A. P. Sample, B. H. Waters, J. R. Smith, and N. Kuster, “Evaluation of wireless resonant power transfer systems with human electromagnetic exposure limits,” *IEEE Trans. Electromagn. Compat.*, vol. 55, no. 2, pp. 265–274, April 2013.

- [25] J. Nadakuduti, M. Douglas, L. Lu, A. Christ, P. Guckian, and N. Kuster, “Compliance testing methodology for wireless power transfer systems,” *IEEE Trans. on Power Electron.*, vol. 30, no. 11, pp. 6264–6273, Nov 2015.
- [26] C. H. Kwan, “Design of wireless power transfer systems in the presence of living objects,” Ph.D. dissertation, Imperial College London, 2018.
- [27] C. A. Walton, “Portable radio frequency emitting identifier,” May 17 1983, US Patent 4,384,288.
- [28] J. E. Petrovic and J. Trenary, “Rechargeable electric portable appliance,” Feb. 15 1983, US Patent 4,374,354.
- [29] J. T. Boys and G. A. Covic, “The inductive power transfer story at the university of auckland,” *IEEE Circuits and Systems Magazine*, vol. 15, no. 2, pp. 6–27, Secondquarter 2015.
- [30] G. A. Covic and J. T. Boys, “Inductive power transfer,” *Proceedings of the IEEE*, vol. 101, no. 6, pp. 1276–1289, June 2013.
- [31] A. Kurs, A. Karalis, R. Moffatt, J. D. Joannopoulos, P. Fisher, and M. Soljačić, “Wireless power transfer via strongly coupled magnetic resonances,” *science*, vol. 317, no. 5834, pp. 83–86, 2007.
- [32] Witricity. (2019) Witricity drive solutions. [Online]. Available: <http://witricity.com/products/automotive/>
- [33] G. Covic, J. T. Boys *et al.*, “Modern trends in inductive power transfer for transportation applications,” *IEEE Trans. Emerg. Sel. Topics Power Electron.*, vol. 1, no. 1, pp. 28–41, 2013.
- [34] S. Li and C. C. Mi, “Wireless power transfer for electric vehicle applications,” *IEEE Trans. Emerg. Sel. Topics Power Electron.*, vol. 3, no. 1, pp. 4–17, March 2015.

- [35] “Wireless Power Transfer for Light-Duty Plug-in/Electric Vehicles and Alignment Methodology,” Society of Automotive Engineers (SAE) International, United States of America, Standard, Apr. 2019.
- [36] Wireless Power Consortium. (2019) Wireless power consortium products. [Online]. Available: <https://www.wirelesspowerconsortium.com/products>
- [37] AirFuel Alliance. (2019) Airfuel about us. [Online]. Available: <https://www.airfuel.org/about/>
- [38] A. Alliance. (2019) Airfuel certified products. [Online]. Available: <https://www.airfuel.org/build-with-airfuel/products/>
- [39] H. A. Mantooth, M. D. Glover, and P. Shepherd, “Wide bandgap technologies and their implications on miniaturizing power electronic systems,” *IEEE Trans. Emerg. Sel. Topics Power Electron.*, vol. 2, no. 3, pp. 374–385, Sept 2014.
- [40] M. Pinuela, D. C. Yates, S. Lucyszyn, and P. D. Mitcheson, “Maximizing DC-to-load efficiency for inductive power transfer,” *IEEE Trans. on Power Electron.*, vol. 28, no. 5, pp. 2437–2447, May 2013.
- [41] K. Surakitbovorn and J. Rivas-Davilla, “Design of a GaN-based wireless power transfer system at 13.56 MHz to replace conventional wired connection in a vehicle,” in *International Power Electronics Conference (IPEC-Niigata 2018-ECCE Asia)*. IEEE, 2018, pp. 3848–3854.
- [42] H. Li, K. Wang, J. Fang, and Y. Tang, “Pulse density modulated ZVS full-bridge converters for wireless power transfer systems,” *IEEE Trans. on Power Electron.*, vol. 34, no. 1, pp. 369–377, 2018.
- [43] G. Zulauf and J. M. Rivas Davila, “Single-turn air-core coils for high-frequency inductive wireless power transfer,” *IEEE Trans. on Power Electron.*, pp. 1–1, 2019.

- [44] M. Huang, Y. Lu, and R. P. Martins, “A reconfigurable bidirectional wireless power transceiver for battery-to-battery wireless charging,” *IEEE Transactions on Power Electronics*, pp. 1–1, 2018.
- [45] S. Y. R. Hui, W. Zhong, and C. K. Lee, “A critical review of recent progress in mid-range wireless power transfer,” *IEEE Trans. on Power Electron.*, vol. 29, no. 9, pp. 4500–4511, Sept 2014.
- [46] C. H. Kwan, D. C. Yates, and P. D. Mitcheson, “Design objectives and power limitations of human implantable wireless power transfer systems,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2016, pp. 1–4.
- [47] K. Van Schuylenbergh and R. Puers, *Inductive powering: Basic theory and application to biomedical systems*. Springer Science & Business Media, 2009.
- [48] C. H. Kwan, G. Kkelis, S. Aldhafer, J. Lawson, D. C. Yates, P. C.-K. Luk, and P. D. Mitcheson, “Link efficiency-led design of mid-range inductive power transfer systems,” in *Emerging Technologies: Wireless Power (WoW), IEEE PELS Workshop*. IEEE, 2015, pp. 1–7.
- [49] G. Kkelis, J. Lawson, D. C. Yates, M. Pinuela, and P. D. Mitcheson, “Integration of a Class-E low  $dv/dt$  rectifier in a wireless power transfer system,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2014, pp. 72–75.
- [50] *C0G (NP0) Dielectric*, AVX, A Kyocera Group Company, 2019, 051818.
- [51] N. Tesla, *Nikola Tesla on his work with alternating currents and their application to wireless telegraphy, telephony, and transmission of power: an extended interview*. 21st Century Books, 2002.
- [52] A. Massarini and M. K. Kazimierczuk, “Self-capacitance of inductors,” *IEEE Trans. on Power Electron.*, vol. 12, no. 4, pp. 671–676, July 1997.

- [53] J. M. Arteaga, S. Aldhafer, G. Kkelis, D. C. Yates, and P. D. Mitcheson, "Multi-MHz IPT systems for variable coupling," *IEEE Trans. on Power Electron.*, vol. PP, no. 99, pp. 1–1, 2017.
- [54] J. M. Arteaga, G. Kkelis, D. C. Yates, and P. D. Mitcheson, "A current driven Class D rectifier with a resistance compression network for 6.78MHz IPT systems," in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2016, pp. 1–4.
- [55] J. M. Arteaga, S. Aldhafer, G. Kkelis, D. C. Yates, and P. D. Mitcheson, "Design of a 13.56 MHz IPT system optimised for dynamic wireless charging environments," in *2016 IEEE 2nd Annual Southern Power Electronics Conference (SPEC)*, Dec 2016, pp. 1–6.
- [56] A. Zaheer, M. Neath, H. Z. Z. Beh, and G. A. Covic, "A dynamic ev charging system for slow moving traffic applications," *IEEE Trans. Transport. Electrification*, vol. PP, no. 99, pp. 1–1, 2016.
- [57] R. Bosshard, J. W. Kolar, J. Mühlethaler, I. Stevanović, B. Wunsch, and F. Canales, "Modeling and  $\eta - \alpha$  -pareto optimization of inductive power transfer coils for electric vehicles," *IEEE Trans. Emerg. Sel. Topics Power Electron.*, vol. 3, no. 1, pp. 50–64, March 2015.
- [58] S. Weerasinghe, U. K. Madawala, and D. J. Thrimawithana, "A matrix converter-based bidirectional contactless grid interface," *IEEE Trans. on Power Electron.*, vol. 32, no. 3, pp. 1755–1766, March 2017.
- [59] Z. Huang, S. Wong, and C. K. Tse, "Control design for optimizing efficiency in inductive power transfer systems," *IEEE Trans. on Power Electron.*, vol. 33, no. 5, pp. 4523–4534, May 2018.
- [60] N. K. Trung and K. Akatsu, "Design high power and high efficiency inverter operating at 13.56 MHz for wireless power transfer systems," in *IEEE Energy Conversion Congress and Exposition (ECCE)*, Sep. 2016, pp. 1–8.

- [61] G. Kkelis, D. C. Yates, and P. D. Mitcheson, “Multi-frequency Class D inverter for rectifier characterisation in high frequency inductive power transfer systems,” in *Power Electronics, Machines and Drives (PEMD)*, April 2016, pp. 1–6.
- [62] N. O. Sokal and A. D. Sokal, “Class E - a new class of high-efficiency tuned single-ended switching power amplifiers,” *IEEE J. Solid-State Circuits*, vol. 10, no. 3, pp. 168–176, June 1975.
- [63] M. K. Kazimierczuk and D. Czarkowski, *Resonant power converters*. John Wiley & Sons, 2012.
- [64] R. E. Zulinski and K. J. Grady, “Load-independent class E power inverters. I. theoretical development,” *IEEE Trans. Circuits Syst.*, vol. 37, no. 8, pp. 1010–1018, Aug 1990.
- [65] S. Gandhi, R. E. Zulinski, and J. C. Mandojana, “On the feasibility of load-independent output current in class E amplifiers,” in *Proceedings of the 34th Midwest Symposium on Circuits and Systems*, May 1991, pp. 870–873 vol.2.
- [66] R. Zulinski and J. Steadman, “Class E power amplifiers and frequency multipliers with finite DC-feed inductance,” *IEEE Trans. Circuits Syst.*, vol. 34, no. 9, pp. 1074–1087, Sep. 1987.
- [67] F. Corti, F. Grasso, A. Reatti, A. Ayachit, D. K. Saini, and M. K. Kazimierczuk, “Design of class-E ZVS inverter with loosely-coupled transformer at fixed coupling coefficient,” in *42nd Annual Conference of the IEEE Industrial Electronics Society (IECON)*, Oct 2016, pp. 5627–5632.
- [68] S. Aldhafer, P. C. K. Luk, and J. F. Whidborne, “Tuning Class E inverters applied in inductive links using saturable reactors,” *IEEE Trans. on Power Electron.*, vol. 29, no. 6, pp. 2969–2978, June 2014.
- [69] Z. Kaczmarczyk, “High-efficiency Class E,  $EF_2$ , and  $E/F_3$  inverters,” *IEEE Trans. on Industrial Electron.*, vol. 53, no. 5, pp. 1584–1593, Oct 2006.



- [70] S. Aldhafer, D. C. Yates, and P. D. Mitcheson, “Modeling and analysis of Class EF and Class E/F inverters with series-tuned resonant networks,” *IEEE Trans. on Power Electron.*, vol. 31, no. 5, pp. 3415–3430, May 2016.
- [71] —, “Design and development of a Class EF<sub>2</sub> inverter and rectifier for multimegahertz wireless power transfer systems,” *IEEE Trans. on Power Electron.*, vol. 31, no. 12, pp. 8138–8150, Dec 2016.
- [72] R. C. N. Pilawa-Podgurski, A. D. Sagneri, J. M. Rivas, D. I. Anderson, and D. J. Perreault, “Very-high-frequency resonant boost converters,” *IEEE Trans. on Power Electron.*, vol. 24, no. 6, pp. 1654–1665, June 2009.
- [73] S. Aldhafer, D. C. Yates, and P. D. Mitcheson, “Load-independent class E/EF inverters and rectifiers for MHz-switching applications,” *IEEE Trans. on Power Electron.*, vol. 33, no. 10, pp. 8270–8287, 2018.
- [74] Siu-Chung Wong and C. K. Tse, “Design of symmetrical Class E power amplifiers for very low harmonic-content applications,” *IEEE Trans. Circuits Syst. I, Reg. Papers*, vol. 52, no. 8, pp. 1684–1690, Aug 2005.
- [75] Z. Kaczmarczyk and W. Jurczak, “A push–pull Class-E inverter with improved efficiency,” *IEEE Trans. on Ind. Electron.*, vol. 55, no. 4, pp. 1871–1874, April 2008.
- [76] M. Albulet and R. E. Zulinski, “Effect of switch duty ratio on the performance of class E amplifiers and frequency multipliers,” *s IEEE Trans. Circuits Syst. I. Fundam. Theory Appl.*, vol. 45, no. 4, pp. 325–335, April 1998.
- [77] M. Fu, H. Yin, M. Liu, and C. Ma, “Loading and power control for a high-efficiency Class E PA-driven megahertz WPT system,” *IEEE Trans. on Ind. Electron.*, vol. 63, no. 11, pp. 6867–6876, Nov 2016.
- [78] S. Aldhafer, “Design and optimization of switched-mode circuits for inductive links,” Ph.D. dissertation, Cranfield University, 2014.

- [79] T. Suetsugu and M. K. Kazimierczuk, “Analysis of dynamic frequency response of class E amplifier,” in *IEEE International Symposium on Circuits and Systems*, May 2009, pp. 2866–2869.
- [80] C. H. K. Jensen, F. M. Spliid, J. C. Hertel, Y. Nour, T. Zsurzsan, and A. Knott, “Resonant full-bridge synchronous rectifier utilizing 15 V GaN transistors for wireless power transfer applications following airfuel standard operating at 6.78 MHz,” in *IEEE Applied Power Electronics Conference and Exposition (APEC)*, March 2018, pp. 3131–3137.
- [81] S. Aldhaher, D. C. Yates, and P. D. Mitcheson, “13.56MHz 50W load-independent synchronous class E rectifier using GaN devices for space-constrained applications,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, June 2018, pp. 1–4.
- [82] G. Kkelis, D. C. Yates, and P. D. Mitcheson, “Comparison of current driven Class-D and Class-E half-wave rectifiers for 6.78 MHz high power IPT applications,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2015, pp. 1–4.
- [83] G. Kkelis, S. Aldhaher, J. M. Arteaga, D. C. Yates, and P. D. Mitcheson, “Hybrid Class-E synchronous rectifier for wireless powering of quadcopters,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2017, pp. 1–4.
- [84] J. M. Arteaga, S. Aldhaher, G. Kkelis, C. Kwan, D. C. Yates, and P. D. Mitcheson, “Dynamic capabilities of multi-MHz inductive power transfer systems demonstrated with batteryless drones,” *IEEE Trans. on Power Electron.*, vol. 34, no. 6, pp. 5093–5104, June 2019.
- [85] S. Park and J. M. Rivas, “Design of a Class-DE rectifier with shunt inductance and nonlinear capacitance for high-voltage conversion,” *IEEE Trans. on Power Electron.*, vol. 33, no. 3, pp. 2282–2294, March 2018.
- [86] Z. Tong, G. Zulauf, J. Xu, J. D. Plummer, and J. Rivas-Davila, “Output capacitance loss characterization of silicon carbide schottky diodes,” *IEEE Trans. Emerg. Sel. Topics Power Electron.*, vol. 7, no. 2, pp. 865–878, June 2019.

- [87] G. Kkelis, D. C. Yates, and P. D. Mitcheson, “Class-E half-wave zero  $dv/dt$  rectifiers for inductive power transfer,” *IEEE Trans. on Power Electron.*, vol. PP, no. 99, pp. 1–1, 2017.
- [88] G. Kkelis, “Rectifier design for optimisation of multi-MHz inductive power transfer systems,” Ph.D. dissertation, Imperial College London, 2017.
- [89] Y. Han, O. Leitermann, D. A. Jackson, J. M. Rivas, and D. J. Perreault, “Resistance compression networks for resonant power conversion,” in *Power Electron. Specialists Conf. (PESC)*, June 2005, pp. 1282–1292.
- [90] —, “Resistance compression networks for radio-frequency power conversion,” *IEEE Trans. on Power Electron.*, vol. 22, no. 1, pp. 41–53, Jan 2007.
- [91] T. W. Barton, J. M. Gordonson, and D. J. Perreault, “Transmission line resistance compression networks and applications to wireless power transfer,” *IEEE J. Emerg. Sel. Topics Power Electron.*, vol. 3, no. 1, pp. 252–260, March 2015.
- [92] M. M. Weiner, “Analysis of cockcroft-walton voltage multipliers with an arbitrary number of stages,” *Review of Scientific Instruments*, vol. 40, no. 2, pp. 330–333, 1969.
- [93] M. J. Schutten, R. L. Steigerwald, and M. H. Kheraluwala, “Characteristics of load resonant converters operated in a high power factor mode,” in *Applied Power Electron. Conf. and Exposition (APEC)*, Mar 1991, pp. 5–16.
- [94] X. Xie, J. Zhang, C. Zhao, Z. Zhao, and Z. Qian, “Analysis and optimization of llc resonant converter with a novel over-current protection circuit,” *IEEE Trans. on Power Electron.*, vol. 22, no. 2, pp. 435–443, March 2007.
- [95] J. M. Arteaga, L. Lan, S. Aldhafer, G. Kkelis, D. C. Yates, and P. D. Mitcheson, “A multi-MHz IPT-link developed for load characterisation at highly variable coupling factor,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, June 2018, pp. 1–4.

- [96] L. Lan, J. M. Arteaga, D. C. Yates, and P. D. Mitcheson, "A reflected impedance estimation technique for inductive power transfer," in *IEEE PELS Workshop on Emerging Technologies: Wireless Power (WoW)*, June 2019, pp. 1–4.
- [97] E. Skountzos, J. M. Arteaga, E. Hadjittofis, D. C. Yates, K. L. S. Campbell, and P. D. Mitcheson, "A 13.56 mhz inductive power transfer system operating with corroded coils," in *IEEE PELS Workshop on Emerging Technologies: Wireless Power (WoW)*, June 2019, pp. 1–6.
- [98] L. Lan, N. M. Ting, S. Aldhafer, G. Kkelis, C. H. Kwan, J. M. Arteaga, D. C. Yates, and P. D. Mitcheson, "Foreign object detection for wireless power transfer," in *2nd URSI Atlantic Radio Science Meeting (AT-RASC)*. IEEE, 2018, pp. 1–2.
- [99] K. Kawabe, H. Koyama, and K. Shirae, "Planar inductor," *IEEE Trans. Magn.*, vol. 20, no. 5, pp. 1804–1806, Sep. 1984.
- [100] P. Meyer and Y. Perriard, "Skin and proximity effects for coreless transformers," in *International Conf. on Electrical Machines and Systems*, Aug 2011, pp. 1–5.
- [101] S. I. Babic and C. Akyel, "Calculating mutual inductance between circular coils with inclined axes in air," *IEEE Trans. Magn.*, vol. 44, no. 7, pp. 1743–1750, July 2008.
- [102] W. B. Kuhn and A. P. Boutz, "Measuring and reporting high quality factors of inductors using vector network analyzers," *IEEE Trans. Microw. Theory Techn.*, vol. 58, no. 4, pp. 1046–1055, 2010.
- [103] J. Lawson, D. C. Yates, and P. D. Mitcheson, "High Q coil measurement for inductive power transfer," *IEEE Trans. Microw. Theory Techn.*, pp. 1–12, 2019.
- [104] A. A. Bastami, A. Jurkov, P. Gould, M. Hsing, M. Schmidt, J. I. Ha, and D. J. Perreault, "Dynamic matching system for radio-frequency plasma generation," *IEEE Trans. on Power Electron.*, vol. 33, no. 3, pp. 1940–1951, March 2018.

- [105] K. M. Shams and M. Ali, “Wireless power transmission to a buried sensor in concrete,” *IEEE Sensors Journal*, vol. 7, no. 12, pp. 1573–1577, 2007.
- [106] J. Kuipers, H. Bruning, S. Bakker, and H. Rijnaarts, “Near field resonant inductive coupling to power electronic devices dispersed in water,” *Sensors and Actuators A: Physical*, vol. 178, pp. 217–222, 2012.
- [107] A. Askari, R. Stark, J. Curran, D. Rule, and K. Lin, “Underwater wireless power transfer,” in *IEEE Wireless Power Transfer Conf. (WPTC)*. IEEE, 2015, pp. 1–4.
- [108] V. N. Vapnik, *The Nature of Statistical Learning Theory*. Berlin, Heidelberg: Springer-Verlag, 1995.
- [109] H. Drucker, C. J. C. Burges, L. Kaufman, A. J. Smola, and V. Vapnik, “Support vector regression machines,” in *Advances in Neural Information Processing Systems 9*, M. C. Mozer, M. I. Jordan, and T. Petsche, Eds. MIT Press, 1997, pp. 155–161.
- [110] S. Mukherjee, E. Osuna, and F. Girosi, “Nonlinear prediction of chaotic time series using support vector machines,” in *Neural Networks for Signal Processing VII. Proceedings of the IEEE Signal Processing Society Workshop*, Sep. 1997, pp. 511–520.
- [111] K. Fukushima, “Neocognitron: A self-organizing neural network model for a mechanism of pattern recognition unaffected by shift in position,” *Biological Cybernetics*, vol. 36, no. 4, pp. 193–202, Apr 1980.
- [112] D. Graupe, R. W. Liu, and G. S. Moschytz, “Applications of neural networks to medical signal processing,” in *Proceedings of the 27th IEEE Conference on Decision and Control*, Dec 1988, pp. 343–347 vol.1.
- [113] S. Lawrence, C. L. Giles, A. C. Tsoi, and A. D. Back, “Face recognition: a convolutional neural-network approach,” *IEEE Trans. Neural Netw.*, vol. 8, no. 1, pp. 98–113, Jan 1997.

- [114] M. S. Jellesen, V. Verdingovas, H. Conseil, K. Piotrowska, and R. Ambat, “Corrosion in electronics: Overview of failures and countermeasures,” in *European Corrosion Congress*, 2014.
- [115] A. I. Sunny, G. Y. Tian, J. Zhang, and M. Pal, “Low frequency (LF) RFID sensors and selective transient feature extraction for corrosion characterisation,” *Sensors and Actuators A: Physical*, vol. 241, pp. 34–43, 2016.
- [116] H. H. Uhlig, *Uhlig’s corrosion handbook*. John Wiley & Sons, 2011, vol. 51.
- [117] W. Wang, A. Choubey, M. H. Azarian, and M. Pecht, “An assessment of immersion silver surface finish for lead-free electronics,” *Journal of electronic materials*, vol. 38, no. 6, pp. 815–827, 2009.
- [118] X. Zhang, I. O. Wallinder, and C. Leygraf, “Mechanistic studies of corrosion product flaking on copper and copper-based alloys in marine environments,” *Corrosion Science*, vol. 85, pp. 15–25, 2014.
- [119] D. Brandon and W. D. Kaplan, *Microstructural Characterization of Materials*. Chichester, UK: John Wiley & Sons, Ltd, 2008.
- [120] C. Pan, W. Lv, Z. Wang, W. Su, C. Wang, and S. Liu, “Atmospheric corrosion of copper exposed in a simulated coastal-industrial atmosphere,” *Journal of materials science & technology*, vol. 33, no. 6, pp. 587–595, 2017.
- [121] J. M. Arteaga, S. Aldhafer, D. C. Yates, and P. D. Mitcheson, “A multi-MHz wireless power transfer system with mains power factor correction circuitry on the receiver,” in *IEEE Applied Power Electron. Conf.*, March 2019, pp. 1–6.
- [122] J. M. Arteaga, D. C. Yates, and P. D. Mitcheson, “Transmitter and receiver circuitry for power converter systems.”
- [123] “Electromagnetic compatibility (EMC) - Part 3-2: Limits - Limits for harmonic current emissions (equipment input current  $\leq 16$  A per phase),” International Electrotechnical Commission, Standard, January 2018.

- [124] C. Qiao and K. M. Smedley, “A topology survey of single-stage power factor corrector with a boost type input-current-shaper,” *IEEE Transactions on Power Electronics*, vol. 16, no. 3, pp. 360–368, May 2001.
- [125] M. K. Kazimierczuk, *RF power amplifiers*. John Wiley & Sons, 2014.
- [126] J. T. Boys, C. Y. Huang, and G. A. Covic, “Single-phase unity power-factor inductive power transfer system,” in *2008 IEEE Power Electronics Specialists Conference*, June 2008, pp. 3701–3706.
- [127] H. Wang and F. Blaabjerg, “Reliability of capacitors for DC-link applications in power electronic converters—an overview,” *IEEE Trans. Ind Appl.*, vol. 50, no. 5, pp. 3569–3578, 2014.
- [128] G. R. Kalra, D. J. Thirmawithana, M. Neuburger, B. Riar, U. K. Madawala, and R. Zane, “A novel boost active bridge based wireless power interface for V2G/G2V applications,” in *IEEE Southern Power Electronics Conference (SPEC)*, Dec 2017, pp. 1–6.
- [129] A. Avila, A. Garcia-Bediaga, U. Iruretagoyena, I. Villar, and A. Rujas, “Comparative evaluation of front and back end PFC IPT systems for a contactless battery charger,” *IEEE Trans. Ind. Appl.*, vol. 54, no. 5, pp. 4842–4850, Sept 2018.
- [130] J. Choi, D. Tsukiyama, Y. Tsuruda, and J. Rivas, “13.56 MHz 1.3 kW resonant converter with GaN FET for wireless power transfer,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2015, pp. 1–4.
- [131] D. C. Yates, S. Aldhaher, and P. D. Mitcheson, “Design of 3 MHz DC/AC inverter with resonant gate drive for a 3.3 kW EV WPT system,” in *IEEE Southern Power Electronics Conference (SPEC)*, Dec 2016, pp. 1–4.
- [132] L. Jiang and D. Costinett, “A high-efficiency GaN-based single-stage 6.78 MHz transmitter for wireless power transfer applications,” *IEEE Trans. on Power Electron.*, pp. 1–1, 2018.

- [133] J. M. Arteaga, G. Kkelis, S. Aldhafer, D. C. Yates, and P. D. Mitcheson, "Probability-based optimisation for a multi-MHz IPT system with variable coupling," in *IEEE PELS Workshop on Emerging Technologies: Wireless Power (WoW)*, June 2018, pp. 1–5.
- [134] A. P. Chandrakasan, N. Verma, and D. C. Daly, "Ultralow-power electronics for biomedical applications," *Annu. Rev. Biomed. Eng.*, vol. 10, pp. 247–274, 2008.
- [135] G. Monti, P. Arcuti, and L. Tarricone, "Resonant inductive link for remote powering of pacemakers," *IEEE Trans. Microw. Theory Techn.*, vol. 63, no. 11, pp. 3814–3822, 2015.
- [136] O. Knecht, R. Bosshard, J. W. Kolar, and C. T. Starck, "Optimization of transcutaneous energy transfer coils for high power medical applications," in *Workshop on Control and Modeling for Power Electronics (COMPEL)*. IEEE, 2014, pp. 1–10.
- [137] J. H. Kim, B.-S. Lee, J.-H. Lee, S.-H. Lee, C.-B. Park, S.-M. Jung, S.-G. Lee, K.-P. Yi, and J. Baek, "Development of 1 MW inductive power transfer system for a high-speed train," *IEEE Trans. Ind. Electron.*, vol. 62, no. 10, pp. 6242–6250, 2015.
- [138] F. Y. Lin, S. Kim, G. A. Covic, and J. T. Boys, "Effective coupling factors for series and parallel tuned secondaries in IPT systems using bipolar primary pads," *IEEE Trans. Transport. Electrification*, vol. 3, no. 2, pp. 434–444, 2017.
- [139] A. Kamineni, M. J. Neath, G. A. Covic, and J. T. Boys, "A mistuning-tolerant and controllable power supply for roadway wireless power systems," *IEEE Trans. on Power Electron.*, vol. 32, no. 9, pp. 6689–6699, 2016.
- [140] L. Zhao, D. J. Thrimawithana, U. K. Madawala, A. P. Hu, and C. C. Mi, "A misalignment-tolerant series-hybrid wireless EV charging system with integrated magnetics," *IEEE Trans. on Power Electron.*, vol. 34, no. 2, pp. 1276–1285, 2018.
- [141] M. Fu, H. Yin, X. Zhu, and C. Ma, "Analysis and tracking of optimal load in wireless power transfer systems," *IEEE Trans. on Power Electron.*, vol. 30, no. 7, pp. 3952–3963, 2014.



- [142] Y. P. Su, X. Liu, and S. Y. R. Hui, “Mutual inductance calculation of movable planar coils on parallel surfaces,” *IEEE Trans. on Power Electron.*, vol. 24, no. 4, pp. 1115–1123, April 2009.
- [143] *Bottom-side cooled 650 V E-mode GaN transistor, GS66504B datasheet*, GaN Systems, 2009, rev. 170321.
- [144] C. K. Kwan, J. M. Arteaga, D. C. Yates, and P. D. Mitcheson, “Design and construction of a 100 W wireless charger for an e-scooter at 6.78 MHz,” in *IEEE PELS Workshop on Emerging Technologies: Wireless Power (WoW)*, June 2019, pp. 1–5.
- [145] S. Aldhafer, P. D. Mitcheson, J. M. Arteaga, G. Kkelis, and D. C. Yates, “Light-weight wireless power transfer for mid-air charging of drones,” in *11th European Conf. on Antennas and Propagation (EUCAP)*, March 2017, pp. 1–5.
- [146] P. D. Mitcheson, D. Boyle, G. Kkelis, D. Yates, J. A. Saenz, S. Aldhafer, and E. Yeatman, “Energy-autonomous sensing systems using drones,” in *IEEE SENSORS*. IEEE, 2017, pp. 1–3.
- [147] P. Mitcheson, G. Kkelis, S. Aldhafer, J. Arteaga, D. Yates, D. Boyle, and E. Yeatman, “Power electronics for wireless power delivery in synthetic sensor networks,” in *Journal of Physics: Conference Series*, vol. 1052, no. 1. IOP Publishing, 2018, p. 012145.
- [148] J. M. Arteaga, S. Aldhafer, G. Kkelis, C. Kwan, D. C. Yates, and P. D. Mitcheson. Wirelessly Powered Drone by Imperial College London - Wireless Power Lab. Youtube. [Online]. Available: <https://www.youtube.com/watch?v=GyOoOeNrbfM>
- [149] M. Hassanalian and A. Abdelkefi, “Classifications, applications, and design challenges of drones: a review,” *Progress in Aerospace Sciences*, vol. 91, pp. 99–131, 2017.
- [150] A. Raciti, S. A. Rizzo, and G. Susinni, “Drone charging stations over the buildings based on a wireless power transfer system,” in *Industrial and Commercial Power Systems Technical Conference (I CPS)*, May 2018, pp. 1–6.

- [151] Wibotic. (2019). [Online]. Available: <https://www.wibotic.com/>
- [152] M. McNabb. Precision Landing for DJI Drones: New Automation Solution from FlytBase. (2019, August 2). [Online]. Available: <https://dronelife.com/2019/08/02/precision-landing-for-dji-drones-new-automation-solution-from-flytbase/>
- [153] S. Jahdi, O. Alatise, P. Alexakis, L. Ran, and P. Mawby, “The impact of temperature and switching rate on the dynamic characteristics of silicon carbide schottky barrier diodes and mosfets,” *IEEE Trans. Ind. Electron.*, vol. 62, no. 1, pp. 163–171, Jan 2015.
- [154] V. Bana, M. Kerber, G. Anderson, J. D. Rockway, and A. Phipps, “Underwater wireless power transfer for maritime applications,” in *IEEE Wireless Power Transfer Conf. (WPTC)*, May 2015, pp. 1–4.
- [155] “E rider model 18 electric bike,” ”<https://eriderbikes.com/electric-bikes/model-18/>”, July 2018.