Design of a 13.56 MHz IPT System Optimised for Dynamic Wireless Charging Environments

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Abstract—Inductive power transfer (IPT) systems are often designed to achieve their highest power efficiency at a fixed load value and at a fixed coil separation distance and misalignment. A variation in the position of the coils or the load value tends to drastically affect the efficiency, and therefore makes the designed IPT system not practical for applications that are mobile with variable loading conditions such as dynamic wireless charging for electric vehicles. This paper presents a novel design approach for loosely-coupled IPT systems that can inherently maintain efficient operation against changes in the system’s characteristics, coil geometries and loading conditions. The transmitting-end of the designed IPT system consists of a Load-Independent Class EF inverter that provides a constant amplitude current in the transmitting-end coil and achieves zero-voltage switching (ZVS) independently of the coupling factor and the load resistance. A Class D rectifier with a resistance compression network (RCN) was implemented in the receiving-end of the IPT system to ensure that the reflected resistance to the transmitting-end is at its optimum value with minimal dependence on the DC load resistance. The combination of the features of the inverter and rectifier allow the IPT system to operate efficiently in an ample set of conditions. Experimental results show a maximal DC-DC efficiency of 83% with a coil-diameter separation at 85 W, and a weighted average DC-DC efficiency of 73% considering 0-100% misalignment.

I. INTRODUCTION

The efficiency of an IPT system can be apportioned according to the different stages of power conversion seen in Fig. 1: the coil driver, the inductive link and the rectifier circuit. Improving the efficiency of each of these stages is key to achieving optimal performance in an IPT system. Nonetheless, the robustness of the system’s performance cannot be neglected as variations on the characteristics of the inductive link may drastically affect the overall efficiency of the system.

Many considerations have been taken into account in all of these power conversion stages to improve the end-to-end efficiency of an IPT system. For instance, the coil driver is often based on resonant Class D and Class E ZVS inverter topologies as they are known to be efficient when operating at megahertz frequencies [1]. In [2], a 77% DC-AC efficiency was achieved in an IPT system with a coupling factor (k) of less than 5% by using a Class E ZVS inverter as the coil driver at 6 MHz.

The link efficiency of an IPT system can be improved by optimising the frequency of operation for a given set of coils, increasing the coupling factor with magnetic materials, or by improving the design of the coils. Furthermore, the link efficiency can be improved by increasing the unloaded quality factor of the transmitting and receiving-end coils (Q_t and Q_r respectively) for a given frequency [3]. In an IPT system with a series-resonant secondary topology, the link efficiency, which accounts only for the losses in the transmitting and receiving-end coils, is given by [3],

\[ \eta_{\text{link}} = \frac{k^2 Q_t \alpha}{\left(\alpha + k^2 Q_t + \frac{1}{Q_r}\right) \left(\alpha + \frac{1}{Q_r}\right)} \]  

where

\[ \alpha = \omega_{\text{res}} C_r R_{\text{AC}}, \]  

\( \omega_{\text{res}} \) is the resonant angular frequency, \( C_r \) is the series resonant capacitor and \( R_{\text{AC}} \) is the AC load.

According to (1) and (2), the link efficiency depends on the input resistance of the rectifier (\( R_{\text{AC}} \)), and reaches a maximum at a certain optimal resistance value [3]. In order to reach the highest link efficiency, the rectifier should be designed to have an input resistance equal to the optimum resistance of the inductive link. The input resistance of the rectifier dictates how much power is delivered to the load, therefore this design approach is more suitable to applications in which the receiving-end of the IPT system can store energy.

Current-driven rectifiers such as Class D and Class E have been integrated and compared in [4] for loosely-coupled IPT systems. In [5], a DC-DC efficiency of 70% was achieved by reflecting the optimal load with the integration of a Class E low
dV/dt rectifier for a 120 W IPT system at 5.56 MHz. Further improvement in high frequency rectifiers show promising results for IPT in [6], [7].

This paper proposes a robust novel configuration of resonant circuits in which the coil driver’s performance is independent of the receiving-end’s load value or the coupling factor. This allows the inverter to operate at ZVS for any reflected resistance. Additionally, the rectifier at the receiving-end is designed to reflect the AC load value that achieves the highest link efficiency at maximal coupling factor, with a compressed dependence on the DC load. The result of this configuration is that the efficiency of a loosely-coupled IPT system becomes remarkably less susceptible to misalignments between the transmitting and receiving-end coils and to large variations in the DC load. Both proposed circuits do not require control feedback or communication to operate thoroughly.

II. System Design

A dynamic charging environment with a variable coupling factor is represented by the diagram in Fig. 2. It shows a mobile receiving-end that changes position relatively to the transmitting-end and consequently the coupling factor between the coils changes. Since the proposed circuit topologies do not require a control loop or a communication link between the transmitting-end and the receiving-end of the IPT system, the design of the link is done such that the rectifier reflects the optimal resistance for maximal coupling factor for the entire range of operation. This design approach of transferring the optimal amount of power, is intended for systems that can store energy; for example, electric vehicles, mobile portable phones, computers and tablets, and wireless charged drones.

A. Design of the IPT link

As can be seen in Fig. 3 and Fig. 4, both transmitting-end and receiving-end circuits were designed with a series-tuned resonant tank in order to avoid changes in the resonant frequency of the system when changing the load’s resistance. Additionally, series-tuned resonant circuits have a low-voltage high-current output which can directly interface with the load without the need for a DC-DC conversion stage.

According to (1), in order to achieve a high link efficiency at low k values, the unloaded Q factor of both the transmitting-end and receiving-end coils, which depends on the equivalent series resistance (ESR) of the coil,

\[ Q_L = \frac{\omega L}{ESR_L(\omega)}, \]

should be as high as possible. For this reason, loosely-coupled IPT systems tend to operate in relatively high frequencies. The ISM bands of 6.78 MHz, 13.56 MHz and 27.12 MHz have been broadly used in loosely-coupled IPT systems showing promising values in end-to-end efficiency [2], [5], [6], [8].

The optimal resistance of the IPT system and the reflected resistance to the transmitting-end should be designed accordingly, taking into account the maximal coupling factor of the IPT system for a given set of coils. The optimal AC load, which corresponds to the input resistance of the rectifier, can be calculated by solving

\[ \frac{\partial \eta_{\text{link}}}{\partial R_{AC}} = 0, \]

for \( R_{AC} \). The reflected resistance in the transmitting-end, for an IPT system with a series-tuned secondary is given by [3]

\[ R_{eq} = \frac{k^2 L_t}{C_r(R_{AC} + ESR_{Lr})}, \]

where \( L_t \) is the inductance of the transmitting-end coil, \( C_r \) is the capacitance of the resonant capacitor of the receiving-end circuit, and \( ESR_{Lr} \) is the equivalent series resistance of the coil in the receiving-end circuit, which should be asserted at the frequency of operation chosen.

B. Design of the Receiving-End Circuit

Resistance compression networks were introduced in [9]. The principle of operation of a RCN is introducing an additional LC circuit designed to reflect a real impedance at the frequency of operation, and therefore having a reflected resistance based not only on the value of the load, but also on the value of the inductance and capacitance of the additional resonant LC circuit. Hence, a rectifier with RCN has a compressed dependence on the DC load. This allows the input resistance to be designed to meet a required value.

In [10] the principle of RCNs was proposed to reflect the optimal load in an IPT system at 6.78 MHz. In that work, the DC load range was between 10-80 Ω and the RCN was designed to compress this range to an input resistance of 7.5-10.5 Ω. Fig. 3 shows a Class D rectifier with a RCN in an IPT system, for this circuit the input resistance is

\[ R_i = \frac{L_{RCN}}{2RC_{RCN}} \left[ 1 + \frac{C_{RCN}R_i^2}{L_{RCN}} \right], \]

where \( R \) is the reflected resistance of the Class D rectifier.
C. Design of the Transmitting-End Circuit

In most IPT applications that operate at tens of megahertz, the topology of the DC/AC inverter in the transmitter is usually based on a Class E or Class EF\textsubscript{2} configuration [2], [11], [12]. While these configurations are power efficient and simple to construct, they can only maintain their optimum switching operation for a fixed load, and therefore are highly dependent on the load value. Consequently, this limits an IPT system with a Class E or Class EF\textsubscript{2} based transmitter to function efficiently only at a fixed coil separation distance and for a narrow load range.

In [11], a Load-Independent Class EF inverter was introduced that is able to maintain efficient operation by achieving ZVS regardless of the load resistance value. In addition, unlike the typical Class E and Class EF\textsubscript{2}, the Load-Independent Class EF inverter can also deliver a constant amplitude output AC current that doesn’t change with load. Therefore, the Load-Independent Class EF inverter can be a more suitable inverter for an IPT system that operates with variable distances and loads. The topology of the Load-Independent Class EF inverter circuit is shown in Fig. 4. The circuit is similar to that of a Class EF\textsubscript{2} or φ\textsubscript{2}, however the resonant frequency of \( L_2 \) and \( C_2 \) is tuned to a frequency between 1.50-1.65 times the switching frequency rather than being tuned at second harmonic. Inductor \( L_1 \) is assumed to be infinite such that its current is DC, and inductor \( L_p \) represents the inductance of the transmitter coil. It should be noted that this inverter cannot maintain load-independent operation when the reactance of the load network is changed.

III. CASE STUDY AND EXPERIMENTAL RESULTS

With the aim of testing a dynamic charging environment, an IPT system was mounted on a linear slide to set the transmitting and receiving-end coils at different distances and misalignments. The experiments were conducted with a separation of approximately one coil-diameter and several different misalignments. Fig. 5 shows the set-up at three of the eight different misalignment configurations that were tested.

The transmitting and receiving-end coils were designed using two-turn 1 cm copper-pipe coils with an approximate diameter of 17.5 cm each. The inductance and the ESR of the coils were measured using a Keysight E4990A impedance analyser and the values are shown in Table I and Table II. An 80 mΩ resistance was also considered for the connectors to the circuit board in the link design. The link efficiency was optimised for these parameters considering a maximum coupling factor of 4.4%, which was calculated through simulation. The range of coupling factor accounted in the experiment varies from 0% to 4.4% by changing the misalignment (dx) from 20-0 cm.

A. Implementation of a Current-Driven Class D Rectifier with a RCN in a 13.56 MHz IPT System

The rectifier was designed to reflect the optimal load to the transmitter throughout a wide range of DC load values. According to the experimental results in [10], the rectifier’s maximum efficiency is achieved when the input resistance
reaches its minimal value. This guideline was followed in the design of the proposed rectifier.

According to (6), the minimum input resistance of the proposed rectifier is equal to the reactance of \( L_{RCN} \). Taking into account that the optimal load for this IPT system at a coupling factor of 4.4 \( \% \) is 3.5 \( \Omega \), a 42 nH inductor was chosen for the RCN. The RCN capacitor was designed to resonate with this inductor at the frequency of operation.

Silicon Schottky diodes were chosen due to their low forward-voltage, and the diode’s junction capacitance is mostly absorbed into the RCN resonant circuit. During the experimental evaluation, it was noticed that at the lower end of the DC load resistance range a low reactance was present in the reflected impedance to the transmitting-end, which resulted in slightly detuning the inverter. This reactance could be attributed to the variation of the diodes’ junction capacitance with DC load. As a result, the system’s performance with a 5 \( \Omega \) DC load could not be verified experimentally for \( V_{DC} > 120 \) V in the higher coupling factor configurations.

The final layout of the implemented rectifier, shown in Fig. 6, includes a common mode choke filter which was incorporated to obtain stable DC measurements in the output. The components chosen for this design are listed in Table I.

The inherent power regulation characteristic of the RCN can be verified in Fig. 7 as the power tends to reach a maximum for a large range of DC loads. When the DC load is set at 10-40 \( \Omega \), the output power variation is less than 5 \( \% \) for 3.9\( \% < k < 4.4\% \) and a 20 \( \% \) variation in the output power was measured for a range of DC loads of 20-60 \( \Omega \) for \( k < 3.9\% \). A nonlinear function between DC load and power output can be noticed from the waveform in Fig. 7 in which maximum output power is reached for the lowest input resistance of the rectifier.

The DC-DC efficiency of the system remained consistently high when varying the load as can be seen in Fig. 7.

### B. Implementation of a Load-Independent Class EF Inverter in a 13.56 MHz IPT System

The design and the mode of operation of the inverter was based on the initial design presented in [11]. The components were adjusted to allow it to deliver the same maximum current of 10 A but at 13.56 MHz. The reflected resistance seen by the inverter is expected to vary from 0 \( \Omega \) when the transmitting and receiving-end coils are completely separated from each to other, to 5 \( \Omega \) when the coils are aligned.

Table II lists the components values for the implemented load-independent Class EF inverter. The transistor used was the GS66504B mosfet (600V) from GaN Systems which was driven using the ISL55110 from Intersil. The gate drive voltage was set at 4.75 V throughout the experiment and gate drive power consumption was constant at 0.14 W. Large copper pads were soldered on the bottom layer of the PCB for improve the thermal performance and to dissipate the heat. A function

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**TABLE I**

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<thead>
<tr>
<th>Component</th>
<th>Value</th>
<th>Description</th>
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<tr>
<td>( C_D ) (pF)</td>
<td>147</td>
<td>Vishay QUAD HIFREQ Series</td>
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<tr>
<td>( C_{RCN} ) (nF)</td>
<td>3.11</td>
<td>Johanson Technology E Series</td>
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<tr>
<td>( C_J ) (pF)</td>
<td>70 to 200</td>
<td>Diode’s capacitance at ( V_o )</td>
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<td>( L_n ) (nH)</td>
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<td></td>
</tr>
<tr>
<td>( ESR_n ) (( \Omega ))</td>
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<td>( L_{RCN} ) (nH)</td>
<td>42</td>
<td>Coilcraft 2014VS</td>
</tr>
<tr>
<td>( R_{DC} ) (( \Omega ))</td>
<td>5 to 120</td>
<td>DC load resistance</td>
</tr>
<tr>
<td>( C_o ) (( \mu F ))</td>
<td>1</td>
<td>Murata GRM Series</td>
</tr>
<tr>
<td>( D )</td>
<td>10 A, 120 V Schottky diodes (FSV10120V)</td>
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**Fig. 6.** Class D rectifier with a RCN for 13.56 MHz.

**Fig. 7.** Measured output power and DC-DC efficiency against DC load for different coil misalignments and an input voltage of 120 V.
TABLE II
COMPONENT VALUES OF THE COIL DRIVER

<table>
<thead>
<tr>
<th>Component</th>
<th>Value</th>
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<td>MOSFET output capacitance at $V_{in}$</td>
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<td>$C_2$ (pF)</td>
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<td>$C_3$ (pF)</td>
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<td>Wurth Elektronik WE-PD</td>
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<td>$L_F$ (nH)</td>
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<td>$R_L$ ($\Omega$)</td>
<td>up to 5.00</td>
<td>reflected resistance</td>
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<tr>
<td>$Q_1$</td>
<td>GS66504B (650 V, 15 A) GaN FET</td>
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Fig. 8. Experimental drain-source voltage waveforms of the switching device of the Load-Independent Class EF inverter for different coil misalignments using a constant DC load of 40Ω and a DC input voltage of 180 V.

Fig. 9. Load-Independent Class EF inverter for 13.56 MHz.

generator was used to provide the gate drive signal and a high-voltage probe with a division ratio of 1000:1 was used to observe the mosfet's drain waveforms which are shown in Fig. 8. Fig. 9 shows a photograph of the implemented load-independent Class EF inverter.

C. Power Measurements and Efficient Energy Transfer

Experimental results were obtained for various configurations consisting of eight different receiving-end positions to change the coupling factor, and three different input voltage values ranging from 120 V to 180 V. The recorded measurements were input DC voltage, input DC current, output DC voltage and output DC current for all cases. In order to assess the system’s performance in a dynamic charging environment, the weighted average DC-DC efficiency ($\eta_{avg}$) was calculated as follows:

$$\eta_{avg} = \frac{\int_{x=0}^{x_{Max}} V_o I_o dx}{\int_{x=0}^{x_{Max}} V_{DC} I_{DC} dx}.$$  (7)

This parameter is useful to evaluate the performance of the proposed configuration because if the receiving-end of an IPT system moves at a constant speed, from zero coupling to maximum coupling, the total amount of energy that is transferred from the DC source to the DC load is the product of $\eta_{avg}$ and the total amount of energy extracted from the source. This calculation was done for three input voltage cases, for which $\eta_{avg} = 73.25\%$ with a maximum variation of less than a 0.25 % for all three cases. This result alludes that a load moving at a constant speed, would extract a 73.25 % of the total energy given by the source if the inverter is powered-on from the exact position that the coupling factor is no longer negligible until it reaches maximum coupling. Fig. 10 shows the graphical representation of the results of this experiment.

An additional feature of this topology, verified by the experiments, is that the total power losses from end-to-end were relatively independent of the load. Taking into account that the Load-Independent Class EF inverter provides a constant amplitude current to the transmitting-end coil and that the ZVS characteristic is not affected by the load, the losses in the transmitting-end coil and the switching device are nearly load-independent. In the experiment, the total losses from end-to-end increased in 7.15 W from no-load to full-load in the worst case, and considering that the losses measured also include the losses in the receiving-end of the IPT system, results strongly suggest that the losses in the transmitting-end did not have a significant change throughout the entire load range. This feature could be beneficial when taking into account thermal cycling effects and control implementation.

The efficiency of the system was consistent for the entire range of input voltage as can be seen in Fig. 10. This allows further improvement of the proposed topology by introducing a power control stage that changes the input voltage of the inverter, and within a certain range, should not affect the efficiency of the IPT system.

IV. CONCLUSIONS

A novel combination of resonant topologies was presented with the aim of achieving high efficiency in dynamic wireless charging environments.

Experimental results show an IPT system capable of transferring up to 120 W at 13.56 MHz with an air gap of 17.5 cm, approximately a coil-diameter distance. A DC-DC efficiency of up to 83 % was measured for a coupling factor of approximately 4.4 % and a 73.25 % weighted average DC-DC efficiency was measured for a dynamic coupling factor of 0-4.4 %.
The rectifier’s reflected resistance was fairly constant for a wide range of DC loads, which proves a compressed transformation between the DC load and the reflected load to the inverter. This helped reflecting an optimal load to the inverter, and maintaining high efficiency for a wide range of DC load.

The Load-Independent Class EF inverter’s ZVS characteristic was not altered by changing the DC load and the coupling factor, and the system did not require additional tuning when changing either the coupling factor or the DC load at the receiving-end. Additionally, the power losses of the system depend mostly on the DC input voltage and not on the coupling factor or the DC load, which strongly suggests that the losses of the transmitting-end have a minimal variation when varying the reflected load from 0 to 100%.

ACKNOWLEDGEMENT

The authors would like to acknowledge the following funding sources: the Department of Electrical and Electronic Engineering of Imperial College London, PINN Programme by the Ministry of Science and ICT of Costa Rica MICITT, University of Costa Rica, EPSRC/EDF Case award number: 1401488 and EPSRC UK-China: Interface and Network Infrastructure to Support EV Participation in Smart Grids; grant ref: EP/L00089X/1, EPSRC Underpinning Power Electronics 2012: Components Theme; grant ref: EP/K034804/1.

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