Development of a 5 kW Inductive Power Transfer System Including Control Strategy for Electric Vehicles

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Abstract
In this contribution a dimensioning and analysis of an Inductive Power Transfer (IPT) system for Electric Vehicles exemplarily for a system with a battery voltage of 370 V and an output power of 5 kW is presented. First by magnetic field analysis an oval-shaped single-sided magnetic coupler design is determined. Series-Series (SS) compensation topology is presented and shown to be suitable for the IPT system. The magnetic coupler inductances are determined on the basis of minimum power transfer capability for a resonance frequency of 60 kHz. Output power is controlled by a primary DC/DC-Converter. Simulations and experimental results show good conformity with analytical approach. An efficiency of 90.5% is realized in the laboratory for an output power of 5 kW and a coupling coefficient of 0.46.

1. Introduction
Charging of EVs can either be done via a cable connection or by usage of a wireless technique. It is mandatory that the charging process has to be safe, efficient and maintenance-free and should be finished in an acceptable period of time [1]. Apart from that a high comfort of the whole charging process is preferred by the user. Under these conditions wireless charging techniques are of high interest.

Several IPT systems were developed in the past. Some of them reach efficiency values >95% such as [2] where an IPT system for an output power of 22 kW based on SiC semiconductors is presented with a DC to DC link efficiency of 97 %. In [3] a contactless power transfer system is developed which uses MOSFETs as power semiconductors with an efficiency of 96.6 % at 4 kW output power. In contrast other systems such as the one in [4] only reach an efficiency of 88.75 % at approximately 3 kW which is considerably less but a realistic value. The objective of this contribution is to investigate further in the IPT system dimensioning with focus on the efficiency. The analysis presented here is of general validity. Nevertheless the investigations are done for a system with three phase input voltage of 400 VAC, a nominal battery voltage of 370 V and a nominal output power of 5 kW.

The paper is arranged as follows. First an optimal magnetic coupler design is found by magnet field simulations. The compensation topology that is applicable as well as the power electronics configuration is presented. Based on the magnetic coupler specifications and the compensation topology the magnetic coupler parameters are determined. A control strategy for power control in the IPT system is presented. Simulations and experimental analyses in the lab complete the investigations.

2. Magnetic Coupler Design
The magnetic coupler in an inductive power transfer system has to fulfill a variety of requirements as are low design height, functionality with large air gaps, low weight and high positioning tolerances [5]. The geometrical requirements for the IPT system considered in this paper are given in tab. 1.
maximum primary / secondary coupler size 500 x 640 mm
minimum / maximum air gap width 50 / 150 mm
maximum lateral / forward displacement 100 / 100 mm

Tab. 1: Geometrical conditions for the magnetic coupler of the regarded IPT system

The objective is to find a design with a high coupling coefficient over the whole operation range as the coupling coefficient is an important criteria for the performance and efficiency of an IPT system [5].

The general question is what kind of coil and core geometry is optimal to be used. In the literature two basic types of designs can be identified. The first one is a core with single-sided windings, which normally has a circular shape [5, 6] and the second one is a core with double-sided windings and rectangular or H-shape [7, 8]. Single-sided core designs normally have high coupling when the air gap is small compared to the dimension of the coupler, but are much more dependent on displacement than double-sided designs. However, averaged over the whole operation range, which is regarded here, single-sided designs show a better performance [9]. Furthermore they inherently reduce stray flux [10] and have no heat problems like double-sided designs [11]. For these reasons a single-sided design is chosen.

To determine the optimal coil and core position and spread extensive 2D axisymmetrical and 3D simulations are performed with the software Comsol Multiphysics. The procedure in detail is described in the literature [5, 9]. It results in the dimensions shown in tab. 2. It should be noted that the magnetic coupler has a rectangular shape, thus lateral and forward dimensions are different (fig. 1).

<table>
<thead>
<tr>
<th></th>
<th>forward direction</th>
<th>lateral direction</th>
</tr>
</thead>
<tbody>
<tr>
<td>$r_{pad}$</td>
<td>500 mm</td>
<td>640 mm</td>
</tr>
<tr>
<td>$w_{FE}$</td>
<td>160 mm</td>
<td>160 mm</td>
</tr>
<tr>
<td>$r_{LFE}$</td>
<td>75 mm</td>
<td>145 mm</td>
</tr>
<tr>
<td>$r_{o,FE}$</td>
<td>235 mm</td>
<td>305 mm</td>
</tr>
<tr>
<td>$w_{coil1}$</td>
<td>150 mm</td>
<td>150 mm</td>
</tr>
<tr>
<td>$r_{o,coil1}$</td>
<td>80 mm</td>
<td>150 mm</td>
</tr>
<tr>
<td>$r_{o,coil2}$</td>
<td>230 mm</td>
<td>300 mm</td>
</tr>
<tr>
<td>$w_{coil2}$</td>
<td>100 mm</td>
<td>100 mm</td>
</tr>
<tr>
<td>$r_{o,coil2}$</td>
<td>130 mm</td>
<td>200 mm</td>
</tr>
<tr>
<td>$r_{o,coil2}$</td>
<td>230 mm</td>
<td>300 mm</td>
</tr>
</tbody>
</table>

Tab. 2: Dimensions of the magnetic coupler as a result of magnetic field simulation in accordance to fig. 2

As can be seen from the magnetic field simulation results (tab. 2) primary and secondary ferrite dimensions should be the same with $w_{FE} = 160$ mm since a wider ferrite core does not result in significant improvement of the coupling coefficient but increases the ferrite volume. For the transformer coils it is found that the primary coil ($w_{coil1} = 150$ mm) should be wider than the secondary ($w_{coil2} = 100$ mm) for optimal coupling. The configuration in tab. 2 results in a coupling coefficient of 0.494 for 100 mm air gap without lateral and forward displacement.

Fig. 1: On-top view of the rectangular magnetic coupler with oval-shaped coil
To further reduce the ferrite volume and decrease the manufacturing complexity ferrite bars instead of a massive ferrite plate are used. For the regarded design 16 ferrite bars with 160 x 40 mm each are employed resulting in a reduction of 58 % ferrite volume. The drawback is a reduction of the coupling coefficient to 0.438; however the advantages are predominant. Furthermore in the simulation it was found that the thickness of the ferrite bars should be at least 3 mm to not exceed the maximum magnetic flux density in the material which is 470 mT for the regarded material (3C94). Here 5 mm ferrite bars are chosen because this is the minimum when regarding fragility.

The analysis of different displacement scenarios in accordance to the underlying conditions is shown in tab. 3. It can be seen that for the best case the coupling coefficient is 0.725 while for the worst case it is 0.234. However, the results are not exact because of the necessary assumptions simplifying the simulations. Therefore a minimum coupling coefficient of 0.2 and a maximum of 0.8 will be used in the following for the dimensioning of the system components.

<table>
<thead>
<tr>
<th>airgap</th>
<th>forward displacement</th>
<th>lateral displacement</th>
<th>coupling coefficient</th>
<th>scenario</th>
</tr>
</thead>
<tbody>
<tr>
<td>50 mm</td>
<td>0 mm</td>
<td>0 mm</td>
<td>0.725</td>
<td>best case</td>
</tr>
<tr>
<td>100 mm</td>
<td>0 mm</td>
<td>0 mm</td>
<td>0.438</td>
<td>nominal position</td>
</tr>
<tr>
<td>150 mm</td>
<td>100 mm</td>
<td>100 mm</td>
<td>0.234</td>
<td>worst case</td>
</tr>
</tbody>
</table>

Tab. 3: Coupling coefficient as a result of different displacement scenarios for a magnetic coupler design with 16 ferrite bars (160 x 40 x 5 mm)

3. Compensation Topologies

In IPT systems primary and secondary inductances of the transformer are compensated by capacitances which are connected in series (S) or in parallel (P). In general four different compensation topologies are possible: SS, SP, PS, PP (while the first letter stands for the primary and the last one for the secondary compensation).

SS compensated resonant topologies (fig. 3) show beneficial properties for the use in IPT systems and are analyzed in detail in the literature [5-8, 12]. There is no reactance being
reflected to the primary side at the resonance frequency. Thus the primary current is in phase with the primary voltage allowing Zero Current Switching (ZCS) of the semiconductor devices [12]. Another advantage is that the primary compensation capacitance can be tuned out independently of the load and the coupling coefficient [12]. For these reasons the SS topology is chosen here.

In this paper unity winding ratio of the transformer is assumed for simplification reasons of the analysis. Furthermore, the magnetic coupler inductances \( L_1 \) and \( L_2 \) with \( L_1 = L_h + L_{1\sigma} \) and \( L_2 = L_h + L_{2\sigma} \) are supposed to be equal. The mutual coupling \( M \) of the primary and secondary coils can be expressed in dependency of the coupling coefficient \( k \):

\[
M = k \sqrt{L_1 L_2}
\]

In SS compensated IPT systems the capacitances \( C_1 \) and \( C_2 \) are determined such that the self-inductances of the transformer are compensated:

\[
C_1 = C_2 = \frac{1}{\omega_r^2 L_1} = \frac{1}{\omega_r^2 L_2}
\]

with \( \omega_r \) as the angular resonance frequency. To determine the transfer characteristics an analysis of the resonant transmission shown in fig. 3 is performed. The input impedance of the circuit is given by:

\[
Z_{1} = R_1 + j\omega L_{1\sigma} + \frac{1}{j\omega C_1} + \frac{1}{j\omega M^2 + R_2 + \frac{1}{L_2} + \frac{1}{L_2} + \frac{1}{R_L}}
\]

The load resistance \( R_L \) can be calculated from the battery characteristics for a particular operation point in steady state [13]. With the compensation capacities tuned at the resonance frequency \( \omega_r \) and an operation at the resonance frequency the expression for the input impedance can be simplified \( R_1 = R_2 = 0 \) to:

\[
Z_{1} = R_1 + \frac{M^2 \omega_r^2}{R_2 + R_L} = \frac{M^2 \omega_r^2}{R_L} \mid_{R_1 = R_2 = 0}
\]

Applying a circuit analysis in frequency domain the output power and the efficiency can be calculated:

\[
P_{\text{out}} = \Re \{U_2 I_2^*\} = \frac{M^2 \omega_r^2 R_L U_1^2}{(M^2 \omega_r^2 + R_1 R_2 + R_L)^2} = \frac{R_L U_1^2}{M^2 \omega_r^2} \mid_{R_1 = R_2 = 0}
\]

\[
\eta = \frac{P_{\text{in}}}{P_{\text{out}}} = \frac{U_1^2}{P_{\text{out}}} = \frac{R_L M^2 \omega_r^2}{(R_2 + R_L)(M^2 \omega_r^2 + R_1 R_2 + R_L)}
\]

4. Power Electronics Topology and Control Scheme

A standard power electronics topology for the IPT system is investigated. It consists of a three phase diode bridge with PFC boost converter that feeds the primary DC link. However, for simplification reasons the AC source and the front end rectifier are not regarded in the following analysis but modeled by a constant voltage source as shown in fig. 4. The primary DC/DC converter allows a reduction of the DC link voltage \( U_{\text{DCin}} \) to \( U_{\text{DC1}} \). The high frequency (HF) inverter is realized in full bridge topology and converts the DC link voltage into a rectangular AC voltage with high frequency. On the secondary side the voltage is rectified by a diode bridge and filtered by a DC capacitor. The battery is connected to the secondary DC link.
The frequency applied to the high frequency inverter which is equal to the resonance frequency is selected to 60 kHz for this application. The IGBT type IKW40N120 from Infineon with anti-parallel diode is used for the HF inverter and the secondary diode bridge as it is designed for applications up to 70 kHz and optimized for low switching and conduction losses. It is rated with a blocking voltage of 1200 V and a maximum collector current of 40 A at 100 °C.

For simplification only the fundamental part of $U_1$ of the rectangular waveform that is generated by the primary HF inverter in fig. 4 is regarded in the analytical investigation. The RMS value of the fundamental oscillation of the rectangular waveform can be calculated with the following equation [6]:

$$U_1 = \frac{4}{\sqrt{2\pi}} U_{DC1}$$

(7)

On the secondary side the diode bridge in combination with the DC capacitance act as a filter transforming the equivalent battery resistance $R_L$ into the transformed value $R_L^*$ [14]:

$$R_L^* = \frac{8}{\pi^2} R_L$$

(8)

As shown in fig. 4 the battery voltage and charging current are measured. The charging power is calculated in an FPGA that has the function of the IPT system controller. To control the charging power or the battery voltage the primary input voltage $U_{DC1}$ of the primary high frequency inverter is adjusted. This is done by using the primary buck converter depicted in fig. 4 with the duty cycle $D$ as the control variable:

$$U_{DC1} = D \cdot U_{DCin}$$

(9)

In normal operation the high frequency converter is driven with maximum duty cycle, considering a dead time for the IGBTs, so that the switching takes place when the current is minimum thus reducing switching losses. In addition to the battery voltage and the charging current the primary current $I_1$ is measured for safety reasons and to determine the optimal operation frequency where the phase shift between primary current and voltage is zero minimizing the reactive power.

5. Magnetic Coupler Dimensioning

From (5) it can be seen that the system designer has the possibility to affect the output power by $M$, $\omega_r$, and $U_1$ while $R_L$ is given by the battery charging characteristic as was explained before. If fixed frequency control is applied, $U_1$ is normally the control variable to control the output voltage and transferred power [3, 15, 16]. With (1) $M$ can be substituted by the coupling coefficient and the magnetic coupler inductance. However, the coupling coefficient is dependent on the magnetic coupler design and the displacement scenarios as was explained above so that $L_1$ is the only parameter to be determined.
From the simplified form of (5) \((R_1 = R_2 = 0)\) it can be seen that \(M\) and consequently \(k\) is in inverse proportion to \(P_{out}\). This means that the output power is minimum when the coupling coefficient is maximum. Therefore the maximum coupling coefficient value \(k_{\text{max}}\) has to be used to determine \(L_1\). The idea behind is that if the nominal power can be transferred in the operation point with lowest power transfer capability it can be transferred in any point. For the same reason the minimum value of the load resistance \(R_{L_{\text{min}}}^*\) must be considered which can be easily derived from the battery characteristics [13]. Here \(R_{L_{\text{min}}}^*\) is assumed to be 20 \(\Omega\). This leads to the magnetic coupler inductances:

\[
L_1 = L_2 = \frac{U_{L_{\text{max}}}}{\omega \cdot k_{\text{max}} \sqrt{P_{out,\text{nom}}}} = 121.8 \, \mu\text{H} \tag{10}
\]

The compensation capacitances according to \(L_1\) and \(L_2\) can be calculated with (2):

\[
C_1 = C_2 = \frac{1}{\omega^2 L_1} = 57.8 \, \text{nF} \tag{11}
\]

6. Simulation

The objective of the simulation is to analyze the effects of harmonics in the circuit which were neglected in the considerations above for a particular operation point with a coupling coefficient of 0.46. The underlying circuit shown in fig. 4 is used for the following analysis on the parameters shown in tab. 4.

<table>
<thead>
<tr>
<th>(a) system parameters</th>
<th>(b) simulation results</th>
<th>analytic results</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R_1, R_2)</td>
<td>0.5 (\Omega)</td>
<td></td>
</tr>
<tr>
<td>(L_1, L_2)</td>
<td>121.8 (\mu\text{H})</td>
<td></td>
</tr>
<tr>
<td>(C_1, C_2)</td>
<td>57.8 (\text{nF})</td>
<td></td>
</tr>
<tr>
<td>(k)</td>
<td>0.46</td>
<td></td>
</tr>
<tr>
<td>(R^*_L)</td>
<td>20 (\Omega)</td>
<td></td>
</tr>
<tr>
<td>(U_{DC1})</td>
<td>379.5 V</td>
<td></td>
</tr>
<tr>
<td>(U_1)</td>
<td>341.6 V</td>
<td></td>
</tr>
</tbody>
</table>

Tab. 4: IPT System parameters (a) and simulation and analytic results (b)

As can be seen from tab. 4 the simulation results show only small difference to the analytic results from the circuit analysis above. From fig. 5 it can be seen that the current waveforms are nearly sinusoidal. This can also be seen from the 3rd and 5th harmonic component of \(I_1\) in tab. 4 which have small amplitudes compared to the fundamental component. From tab. 4 it can also be seen that the apparent power is about 10 % larger than the real power. This fact has to be considered when designing the power electronics components.
7. Experimental Results
To verify the analytic and simulation results a laboratory setup shown in fig. 6 is designed. It consists of the primary high frequency inverter, the resonant transmission comprising the compensation capacitances and the magnetic coupler as well as the secondary rectifier. The battery is modeled by an equivalent load resistance.

The system parameters of the IPT system were adjusted to the parameters derived above (tab. 4), nevertheless small deviations are existing due to the non-ideal components. The laboratory parameters and results are shown in tab. 5. It should be noted that the load resistance is set to \( R_L = 24.67 \, \Omega \) in accordance to (8) so that the resistance seen by the resonant transmission is 20 \( \Omega \). The efficiency is measured from \( U_{DC1} \) to \( U_{Bat} \) according to fig. 4.

<table>
<thead>
<tr>
<th>(a)</th>
<th>simulation parameters</th>
<th>laboratory parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>( R_1, R_2 )</td>
<td>0.5 ( \Omega )</td>
<td>0.72 / 0.57 ( \Omega )</td>
</tr>
<tr>
<td>( L_1, L_2 )</td>
<td>121.8 ( \mu H )</td>
<td>119.5 / 121.0 ( \mu H )</td>
</tr>
<tr>
<td>( C_1, C_2 )</td>
<td>57.8 nF</td>
<td>55.85 / 56.37 nF</td>
</tr>
<tr>
<td>( k )</td>
<td>0.46</td>
<td>0.464</td>
</tr>
<tr>
<td>( R_L )</td>
<td>20 ( \Omega )</td>
<td></td>
</tr>
<tr>
<td>( R_L )</td>
<td>24.67 ( \Omega )</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>(b)</th>
<th>simulation results</th>
<th>laboratory results</th>
</tr>
</thead>
<tbody>
<tr>
<td>( U_{DC1} )</td>
<td>379.5 V</td>
<td>402.6 V</td>
</tr>
<tr>
<td>( U_1 )</td>
<td>341.6 V</td>
<td>362.5 V</td>
</tr>
<tr>
<td>( I_1 )</td>
<td>15.43 A</td>
<td>15.0 A</td>
</tr>
<tr>
<td>( U_{c1} )</td>
<td>706.1 V</td>
<td>746 V</td>
</tr>
<tr>
<td>( I_2 )</td>
<td>15.9 A</td>
<td>15.8 A</td>
</tr>
<tr>
<td>( U_{c2} )</td>
<td>726.4 V</td>
<td>762 V</td>
</tr>
<tr>
<td>( P_{out} )</td>
<td>5019 W</td>
<td>5007 W</td>
</tr>
<tr>
<td>( P_{in} )</td>
<td>5266 W</td>
<td>5531 W</td>
</tr>
<tr>
<td>( \eta )</td>
<td>95.31 %</td>
<td>90.5 %</td>
</tr>
</tbody>
</table>

From the results (tab. 5 and fig. 8) it can be seen that there is a deviation in current and voltage values between simulation and experimental results. This is mainly caused by the non-ideal characteristics of the transformer coils that show different resistance values (0.72 vs. 0.57 \( \Omega \)) which are moreover significantly higher than in simulation. This is one reason why the DC input voltage has to be approximately 23 V higher to obtain the same output power compared to the simulation. Another reason is the additional losses (semiconductor switching and conduction losses, skin and proximity losses) which were neglected in the simulation. The result is a lower efficiency (90.5 %) compared to the simulation (95.31 %). From this fact it can be concluded that the losses in the transformer coils account for about half of the total losses in the system. Nevertheless voltage and current waveforms in fig. 8 are in good accordance with the simulated waveforms in fig. 5. Zero Current Switching can be stated for the high frequency inverter IGBTs.
8. Conclusion
In this contribution the dimensioning and analysis of an Inductive Power Transfer (IPT) system for Electric Vehicles is presented. This is exemplarily done for a system with an output power of 5 kW and a battery voltage of 370 V. A magnetic coupler design is found with a minimum coupling coefficient of 0.234 and a maximum coupling coefficient of 0.725. SS compensation is shown to be appropriate for the IPT system regarded here. The magnetic coupler inductances are determined such that the nominal power can be transferred in any operation point. Simulation results show good conformity with the analytical approach. Experimental results show some deviation to simulation results mainly because of the parameter deviations of the transformer coils and losses that were not regarded in simulation. Nevertheless an efficiency of 90.5 % is achieved in the lab for 5 kW output power and a coupling coefficient of 0.46.

References