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Theoretical and Practical Design Approach of Wireless Power Systems

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Abstract

The paper introduces the main issues concerned with the conceptual design process of wireless power systems. It analyses the electromagnetic design of the inductive magnetic coupler and proposes the key formulas to optimize its electrical parameters for a particular load. For this purpose, a very detailed analysis is given focusing on the mathematical concept procedure for determination of the key factors influencing proper coupling coils design. It also suggests basic topologies for conceptual design of power electronics and discusses its proper connection to the grid. The proposed design strategy is verified by experimental laboratory measurement including analyses of leakage magnetic field.

Keywords: wireless power transfer, coil design, analytical approach, electromagnetic field, efficiency, optimization, shielding

1. Introduction

There are currently many methods, links and approaches for wireless power transmission. Each of the available solutions is characterized by its advantages and disadvantages, which result in their application [1–3].

Inductive coupling is currently the most widely used method of wireless energy transfer. This method works on the principle of an air transformer with a tight magnetic coupling of the primary and secondary windings. The energy exchange between two or more coils takes place by means of an inductive current Φ , i.e. by means of an induced voltage. The main disadvantage of inductive coupling is the transmission distance, which ranges from millimeters to several centimeters [4].

Resonance compensation is a specific case of inductive coupling. Resonant compensation is used in cases where it is necessary to achieve impedance dependence on frequency. Resonant compensation is provided by adding a capacitive member to the primary as well as the secondary coil. After applying a magnetic field with a suitably selected frequency, the phenomenon of mutual interference of the impedance of the coil and the capacitor occurs, which ideally ensures zero phase shift against the current flowing through the primary coil. For resonant compensation, there are four configurations of the primary and secondary side of the wireless charging system [5, 6].

The system using resonant coupling fully compensates for the scattering fields of the coupling coils, thus significantly extending the working distance while maintaining high energy transfer efficiency. Thanks to its advantages, resonant coupling is used mainly in the field of electromobility, where it allows charging with high power and in the case of a variable load, it can be easily frequency-adjusted for optimal efficiency [7, 8].

Energy transmission through capacitive coupling is currently used relatively little due to limitations on the transmission distance, which is limited by the level of tenths of a millimeter. This method is mainly used for charging consumer electronics such as tablets, laptops and more. They also have great potential in the field of medicine for charging various implants. Capacitive coupling is a phenomenon occurring between all conductive objects, i.e. between systems between which there is a mutual difference of potentials and between them there is an environment with a positive dielectric constant (permittivity) [9–11].

This work aims to point out the main design issues related to wireless power transmission and demonstrate their operational characteristics. An important aspect in this area is undoubtedly interaction of living organisms with a strong electromagnetic field, and therefore it is necessary to pay attention also legislation and hygiene standards [12–15]. Another goal is to provide a clear mathematical description of the system using intuitive methods for circuit analysis. Mathematical models must consider, in addition to the coupling itself, the inverters (inverter and rectifier) on the primary and secondary side of the system. An equally important goal of the work is experimental verification of all achieved theoretical conclusions. For this purpose, it was necessary to develop a prototype of a WPT charging system capable to supply sufficient power needed to charge conventional electric car. The text is supplemented by accompanying graphics that illustrates efficiency characteristics and also analysis of the spatial distribution of the electromagnetic field at different states of the system.

2. Basic resonant coupling techniques

There are four basic configurations of the primary and secondary side of the WPT system to realize resonant compensation of the leakage inductance [16–18]. This chapter focuses on investigating the properties of possible compensation methods, which include serial-serial, serial-parallel, parallel-serial, parallel–parallel.

Parameter	Value	
R_1	0.45 Ω	
R_2	0.45 Ω	
L_1	145.6 μH	
L_2	145.6 μH	
C_1	3.1 nF	
C_2	3.1 nF	
k	0÷0.1	
R_Z	Series compensation: 5÷200 Ω	Parallel comp.: 5 Ω ÷2.5 k Ω
U_b	100 V	

Table 1.
Circuit parameters for the evaluation of compensation techniques.

The analysis of individual configurations is further provided, while and examples of characteristics derivations are based on the circuit parameters listed in **Table 1**.

2.1 Series-series compensation

Serial-series compensation uses two external capacitors C_1 and C_2 connected in series with the primary and secondary windings. The circuit model of the system is shown in **Figure 1**.

The system is powered by an inverter with a rectangular voltage profile with amplitude u_{m1} , and therefore the circuit must be described by a system of integrodifferential equations forming a full-fledged dynamic model.

$$\begin{aligned} -u_1 + \frac{1}{C_1} \int_0^t i_1 dt + u_{C_1(0)} + R_1 i_1 + L_1 \frac{di_1}{dt} + M \frac{di_2}{dt} &= 0 \\ -L_2 \frac{di_2}{dt} - M \frac{di_1}{dt} - R_2 i_2 - \frac{1}{C_2} \int_0^t i_2 dt - u_{C_2(0)} - R_Z i_2 &= 0 \end{aligned} \tag{1}$$

All models will be derived for the fundamental harmonic and therefore we can use Eq. (1) to describe the model, while the inverter voltage can be considered in the form of (2).

$$\dot{U}_1 = \frac{2\sqrt{2}}{\pi} u_{m1} \tag{2}$$

The solution we get loop currents, from which it is possible to further determine all operating variables of the system (3).

$$\begin{bmatrix} \dot{I}_{S1} \\ \dot{I}_{S2} \end{bmatrix} = \begin{bmatrix} R_1 + j\left(\omega L_1 - \frac{1}{\omega C_1}\right) & -j\omega M \\ -j\omega M & R_2 + R_Z + j\left(\omega L_2 - \frac{1}{\omega C_2}\right) \end{bmatrix} \begin{bmatrix} \dot{U}_1 \\ 0 \end{bmatrix} \tag{3}$$

For a better idea, we draw the efficiency and power on the load depending on the frequency and the coupling factor, respectively the load. This creates two pairs of maps in which two functions are plotted separately:

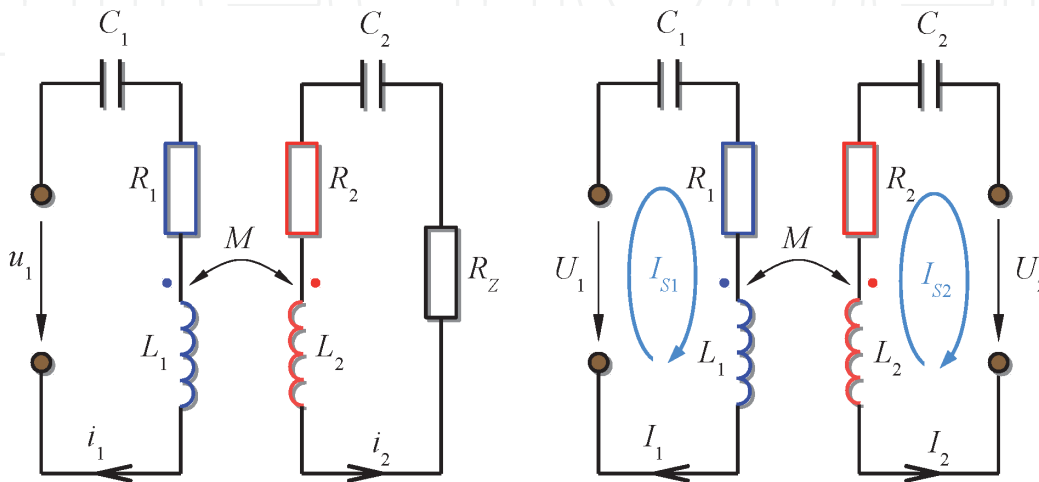


Figure 1.
Simplified equivalent circuit for series-series compensation, left – Circuit with initial variables, right – Circuit suited for loop current analysis.

$$P_2 = f(f_{sw}, k), P_2 = f(f_{sw}, R_Z), \text{ and } \eta = f(f_{sw}, k), \eta = f(f_{sw}, R_Z) \quad (4)$$

The first map will consider a constant load, which will be set as optimal for the working distance corresponding to $k = 0.1$ [19]. The optimal load for the map was determined based on the relationship:

$$R_{Zs-efficiency} = \sqrt{\frac{R_2(R_1 R_2 + M^2 \omega^2)}{R_1}} \doteq 22 \Omega. \quad (5)$$

The resulting maps are shown in **Figure 2**.

2.2 Series-parallel compensation

Serial-parallel compensation uses two resonant capacitors connected in series with a coupling coil on the primary side and in parallel on the secondary side (**Figure 3**). In the left part you can see the diagram for the dynamic model and in the right part its simplification for the supply of the harmonic course of the voltage.

Unlike the previous circuit, it is now necessary to compile three equations of three unknowns. The integrodifferential form is given in the (6).

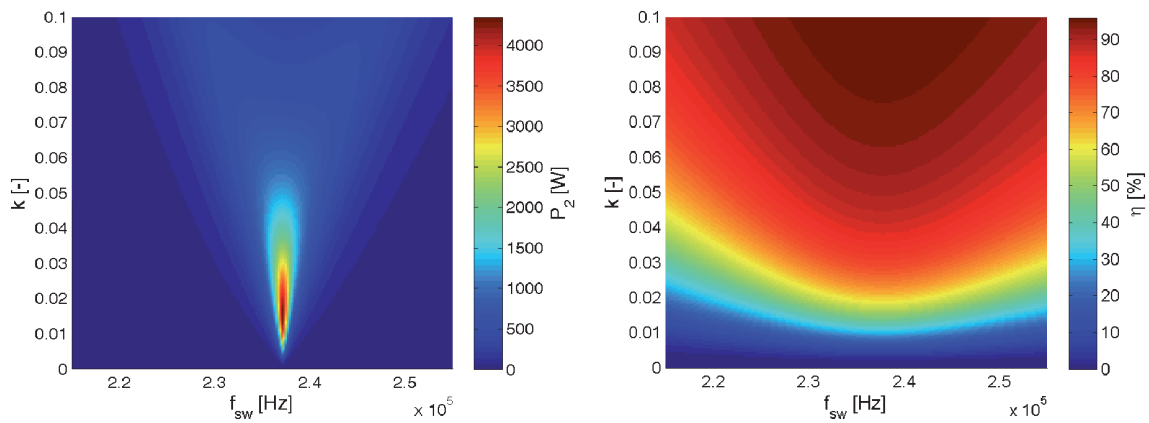


Figure 2. Dependency of output power (left) and system efficiency (right) on frequency and power transfer distance for S-S compensation.

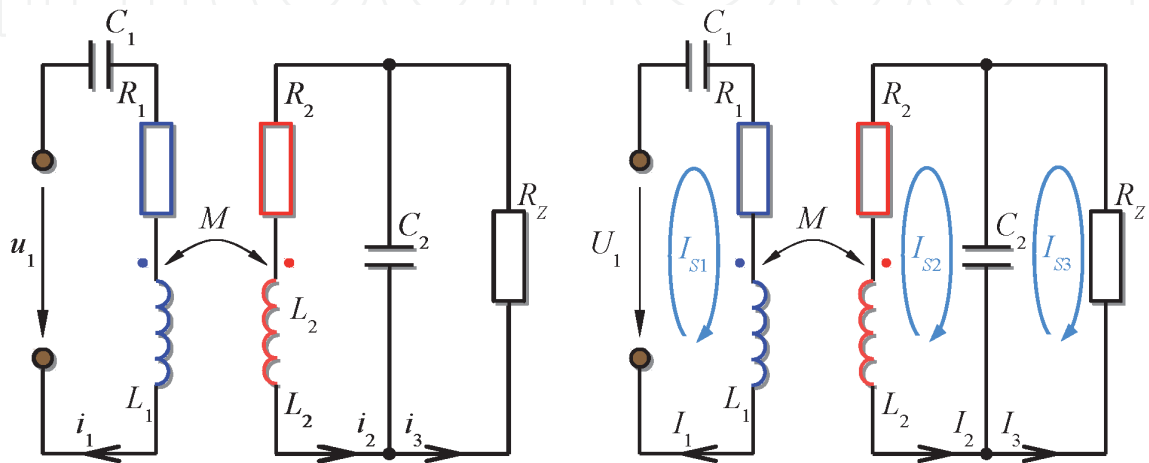


Figure 3. Simplified equivalent circuit for series-parallel compensation, left – Circuit with initial variables, right – circuit suited for loop current analysis.

$$\begin{aligned} -u_1 + \frac{1}{C_1} \int_0^t i_1 dt + u_{C_1(0)} + R_1 i_1 + L_1 \frac{di_1}{dt} + M \frac{di_2}{dt} &= 0 \\ -L_2 \frac{di_2}{dt} - M \frac{di_1}{dt} - R_2 i_2 + \frac{1}{C_2} \int_0^t (i_3 - i_2) dt + u_{C_2(0)} &= 0 \\ -\frac{1}{C_2} \int_0^t (i_3 - i_2) dt - u_{C_2(0)} - R_Z i_3 &= 0 \end{aligned} \tag{6}$$

For the calculation by the loop current method, the equations take the form (7).

$$\begin{bmatrix} \dot{I}_{S1} \\ \dot{I}_{S2} \\ \dot{I}_{S3} \end{bmatrix} = \begin{bmatrix} R_1 + j\left(\omega L_1 - \frac{1}{\omega C_1}\right) & -j\omega M & 0 \\ -j\omega M & R_2 + j\left(\omega L_2 - \frac{1}{\omega C_2}\right) & \frac{j}{\omega C_2} \\ 0 & \frac{j}{\omega C_2} & \frac{-j}{\omega C_2} + R_Z \end{bmatrix} \begin{bmatrix} \dot{U}_1 \\ 0 \\ 0 \end{bmatrix} \tag{7}$$

The optimal load is then given by the Eq. (8)

$$R_{Zp-efficiency} = \frac{L_2 \omega \sqrt{R_1 R_2^2 + \omega^2 (L_2^2 R_1 + M^2 R_2)}}{\sqrt{R_2 (R_1 R_2 + M^2 \omega^2)}} \doteq 2176 \Omega. \tag{8}$$

The resulting waveforms are graphically summarized in **Figure 4**.

2.3 Parallel-series compensation

Parallel-series compensation is practically only like the previous variant, which uses two resonant capacitors connected in parallel with the coupling coil on the primary side and in series on the secondary side. The circuit model is apparent from **Figure 5**.

As in the previous case, it is enough to compile three equations of three unknowns for the description (9).

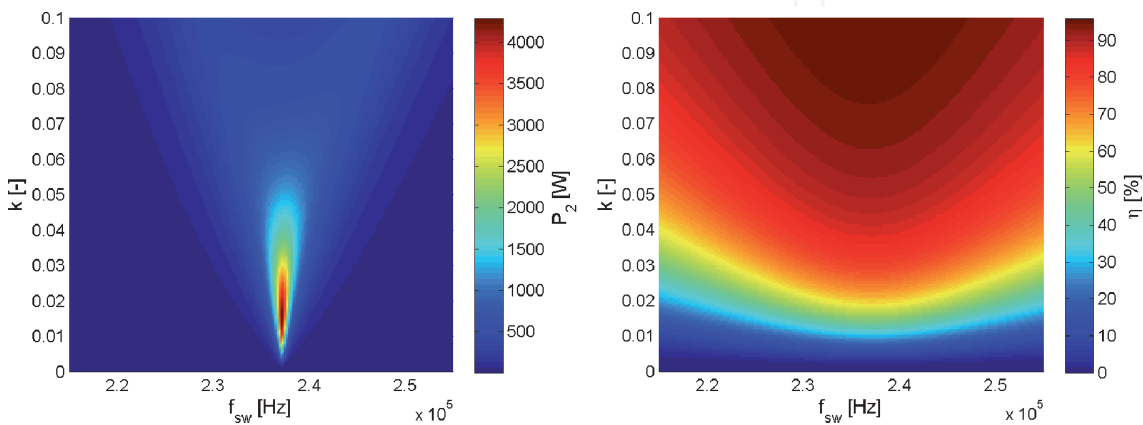


Figure 4.
 Dependency of output power (left) and system efficiency (right) on frequency and power transfer distance for S-P compensation.

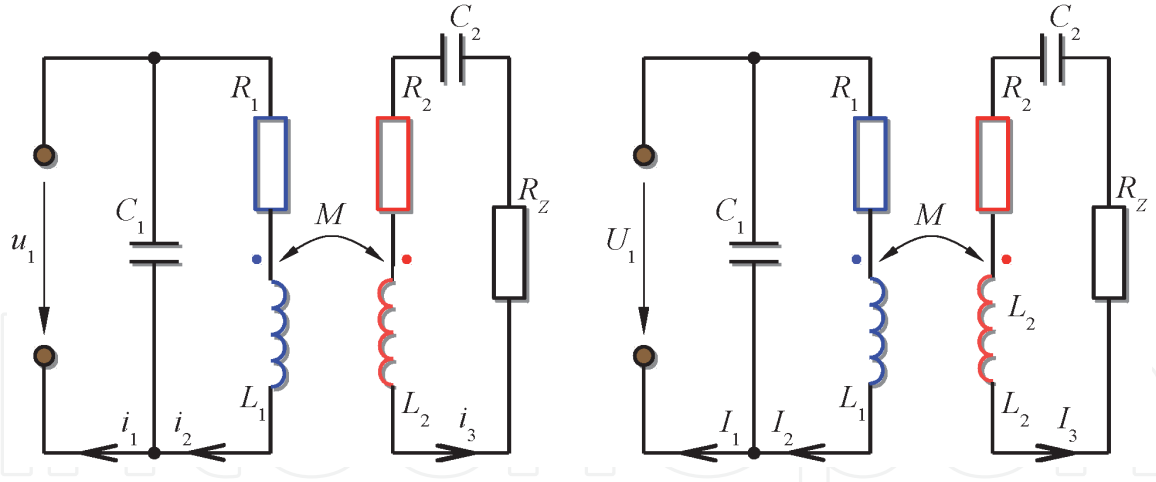


Figure 5.
Simplified equivalent circuit for parallel-series compensation, left – Circuit with initial variables, right – Circuit suited for loop current analysis.

$$\begin{aligned}
 -u_1 + \frac{1}{C_1} \int_0^t (i_1 - i_2) dt + u_{C_1(0)} &= 0 \\
 -\frac{1}{C_1} \int_0^t (i_1 - i_2) dt - u_{C_1(0)} + R_1 i_2 + L_1 \frac{di_2}{dt} + M \frac{di_3}{dt} &= 0 \\
 -L_2 \frac{di_3}{dt} - M \frac{di_2}{dt} - R_2 i_3 - R_Z i_3 &= 0
 \end{aligned} \tag{9}$$

The equations below for the calculation by the loop method can be expressed as (10).

$$\begin{bmatrix} \dot{I}_{S1} \\ \dot{I}_{S2} \\ \dot{I}_{S3} \end{bmatrix} = \begin{bmatrix} \frac{-j}{\omega C_1} & \frac{j}{\omega C_1} & 0 \\ \frac{j}{\omega C_1} & R_1 + j\left(\omega L_1 - \frac{1}{\omega C_1}\right) & -j\omega M \\ 0 & -j\omega M & R_2 + R_Z + j\left(\omega L_2 - \frac{1}{\omega C_2}\right) \end{bmatrix} \begin{bmatrix} \dot{U}_1 \\ 0 \\ 0 \end{bmatrix} \tag{10}$$

From the above equations we again obtain the courses of all-important operating variables, shown in **Figure 5**. As in the previous case, the system achieves maximum efficiency at the optimized load ($R_Z = 22 \Omega$), while the transmitted power is significantly lower compared to the achievable value (**Figure 6**).

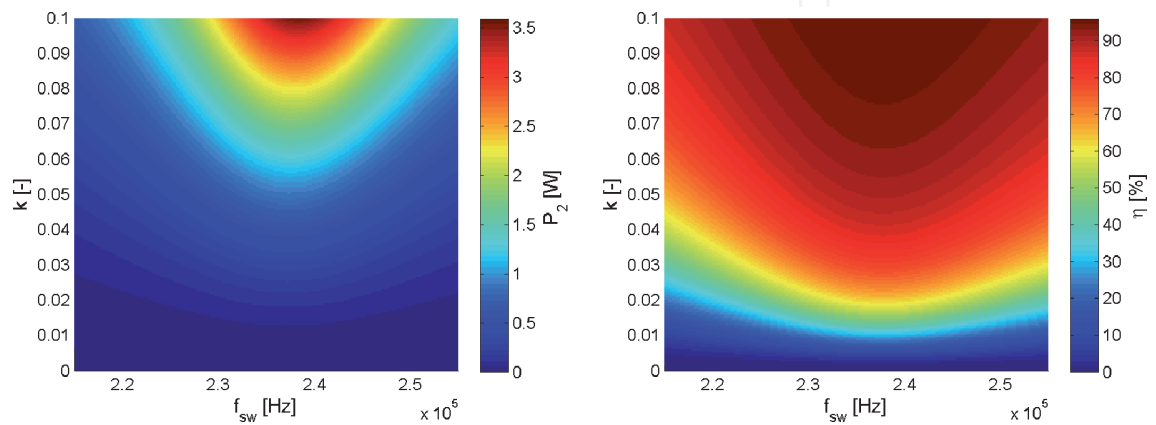


Figure 6.
Dependency of output power (left) and system efficiency (right) on frequency and power transfer distance for P-S compensation.

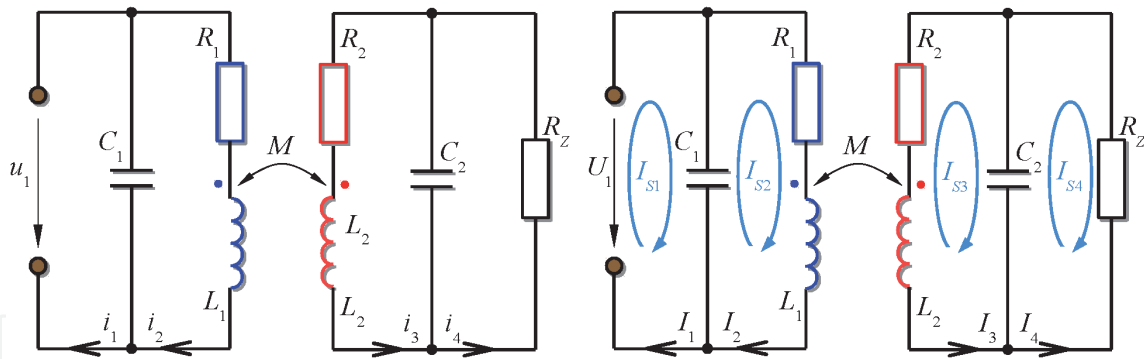


Figure 7.
Simplified equivalent circuit for parallel - parallel compensation, left – Circuit with initial variables,
right – Circuit suited for loop current analysis.

2.4 Parallel-parallel compensation

Parallel-parallel compensation uses two resonant capacitors connected in parallel to both coupling coils. The circuit model can be seen in **Figure 7**.

Unlike previous models, in this case it is necessary to compile four equations with four unknowns, the integrodifferential form is represented by the Eqs. (11).

$$\begin{aligned}
 & -u_1 + \frac{1}{C_1} \int_0^t (i_1 - i_2) dt + u_{C_1(0)} = 0 \\
 & -\frac{1}{C_1} \int_0^t (i_2 - i_1) dt - u_{C_1(0)} + R_1 i_2 + L_1 \frac{di_2}{dt} + M \frac{di_3}{dt} = 0 \\
 & -L_2 \frac{di_3}{dt} - M \frac{di_2}{dt} - R_2 i_3 + \frac{1}{C_2} \int_0^t (i_4 - i_3) dt + u_{C_2(0)} = 0 \\
 & -\frac{1}{C_2} \int_0^t (i_4 - i_3) dt - u_{C_2(0)} - R_Z i_4 = 0
 \end{aligned} \tag{11}$$

After stabilization we can rewrite the equations into the form (12).

$$\begin{bmatrix} \dot{I}_{S1} \\ \dot{I}_{S2} \\ \dot{I}_{S3} \\ \dot{I}_{S4} \end{bmatrix} = \begin{bmatrix} \frac{-j}{\omega C_1} & j \frac{1}{\omega C_1} & 0 & 0 \\ \frac{j}{\omega C_1} & R_1 + j \left(\omega L_1 - \frac{1}{\omega C_1} \right) & -j\omega M & 0 \\ 0 & -j\omega M & R_2 + j \left(\omega L_2 - \frac{1}{\omega C_2} \right) & \frac{j}{\omega C_2} \\ 0 & 0 & \frac{j}{\omega C_2} & R_Z - \frac{j}{\omega C_2} \end{bmatrix} \begin{bmatrix} \dot{U}_1 \\ 0 \\ 0 \\ 0 \end{bmatrix} \tag{12}$$

Graphical interpretations of the characteristics are shown in **Figure 8**.

2.5 Overall comparison

Based on the above results, we can compile a table that compares the key properties of individual compensation methods (**Table 2**). In the evaluation, we consider a system operating to the optimal load at a distance corresponding to the coupling factor $k < 0.1$. The supply voltage is the same for all compensation topologies and is not regulated.

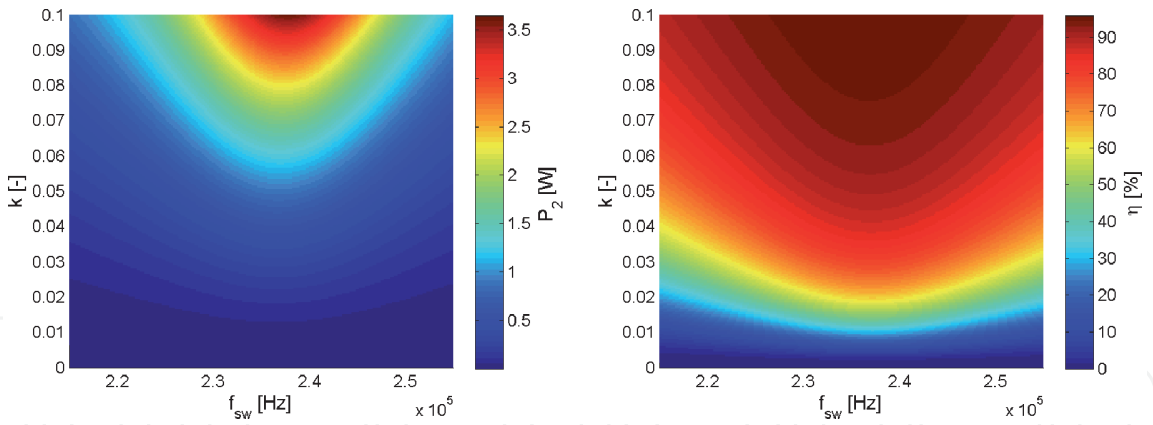


Figure 8. Dependency of output power (left) and system efficiency (right) on frequency and power transfer distance for P–P compensation.

Criteria	Compensating topology			
	S-S	S-P	P-S	P-P
Source output type	current	voltage	voltage	Current
Power transfer ability	higher	higher	lower	lower
Max power and efficiency overlap	No	No	Partial	Partial
Optimum load value	lower	higher	lower	higher
Frequency sensitivity	higher	higher	lower	lower

Table 2. Comparisons of key attributes of individual compensation techniques.

Serial-to-series compensation acts as a current source over the monitored operating distance range, allowing higher power to be delivered to the load compared to other topologies. The disadvantage is the minimum overlap of work areas with maximum system performance and efficiency.

Serial-parallel compensation in this case does not bring any significant operational benefits, the only difference lies in the higher values of the optimal load. Unlike the previous solution, the circuit acts as a voltage source.

Parallel-series compensation offers partial overlap of work areas with maximum performance and efficiency. However, the theoretically achievable transmitted power values are significantly lower compared to the two previous configurations. The circuit has a voltage output and is much less sensitive to frequency detuning.

Parallel-parallel compensation offers current output with better coverage of areas of maximum power and efficiency along with better frequency stability. However, the transmitted powers are very low, as with parallel-series compensation.

3. Identification of key system parameters and analytical approach for design of coupling elements

In the previous chapter, the principal characteristics regarding basic modifications of the main circuits for wireless power systems were derived and described. All models, in some form, use concentrated parameters of spare electrical circuits. These parameters can be determined basically in two ways, i.e. by calculation and by measurement, while the measurement can be used only if the analyzed system

already physically exists. On the other side the calculations offer the possibility of optimization within design procedure.

Within next text, the systematic analytical procedure for calculation of key parameters of coupling elements for wireless power transfer is described, whereby the application area for any compensation technique can be considered here.

3.1 Mutual inductance of planar coils

In practice, circular or spiral coils are most often used for high-frequency purposes. The reason is the high gradients of the electric field, which arise on all structural edges of the coil in the case of parallel resonance. These gradients significantly worsen the quality factor and thus the operating characteristics of the resulting system [20, 21].

The mutual inductance of different clusters of air coils of spiral shape can be based on the application of the analytical rule for the magnetic vector potential in cylindrical coordinates (13).

$$A_{\varphi}[r, \varphi, z] = \frac{\mu_0 I}{4\pi} r_1 \int_0^{\pi} \frac{\cos(\varphi')}{\sqrt{r^2 + r_1^2 - 2rr_1 \cos(\varphi') + (z - z')^2}} d\varphi', \quad (13)$$

The resulting relationship is based on the direct application of Biot-Savart's law. In technical practice, these integrals are abundant, and therefore considerable attention has been paid to their enumeration in the past. The literature defines three basic types of these (elliptic) integrals, which can be combined with each other and easily converted to any special case.

The Eq. (14) was determined based on Eq. (13).

$$A_{\varphi}[r, \varphi, z] = \frac{\mu_0 I}{4\pi} r_1 \frac{2}{\sqrt{r_1 r}} \left[\left(\frac{2}{k_I} - k_I \right) K(k_I) - \frac{2}{k_I} E(k_I) \right] \quad (14)$$

Where $K(k_I)$ and $E(k_I)$ are elliptical integrals of first and second type and have following forms:

$$K(k_I) = \int_0^{\frac{\pi}{2}} \frac{d\varphi}{\sqrt{1 - k_I^2 \sin^2(\varphi)}} \quad E(k_I) = \int_0^{\frac{\pi}{2}} \left[\sqrt{1 - k_I^2 \sin^2(\varphi)} \right] d\varphi. \quad (15)$$

Module of these integrals is determined using (16).

$$k_I = \sqrt{\frac{4rr_1}{(r + r_1)^2 + (z - z')^2}} \quad (16)$$

The derived relations correspond to a simplified geometry, where only a coaxial arrangement is considered [22]. In order to be able to calculate the mutual inductance of the coils of general geometry and arrangement, we must introduce the possibility of deflection, see **Figure 9** left.

For this special case, the procedure for modifying the previous equations was indicated in [22, 23]. For the mutual inductance of the two turns from **Figure 9** (left) we can write according to **Figure 9** (right) next Eq. (17).

$$M_{ij} = \frac{2\mu_0}{\pi} \sqrt{r_{1i} r_{2j}} \frac{\int_0^{\pi} \left[\cos(\varphi) - \frac{d}{r_{2j}} \cos(\varphi) \right] \left[\left(1 - \frac{k_{II}^2}{2} \right) K(k_{II}) - E(k_{II}) \right] d\varphi}{k_{II} \sqrt{3}}, \quad (17)$$

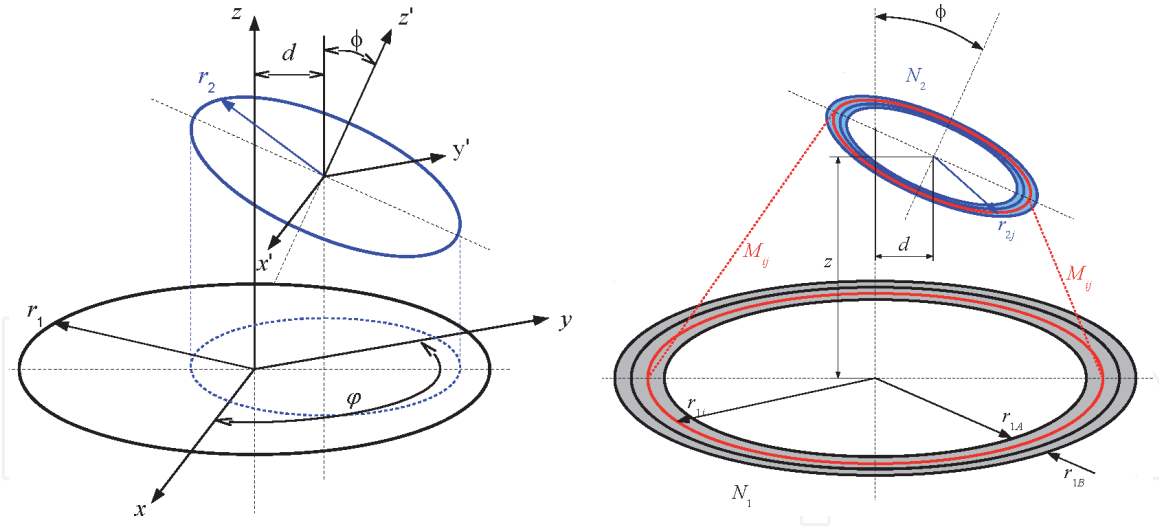


Figure 9. Displacement of coil's turns and presentation of the situation for determination of mutual inductance for circular coils.

where

$$\sqrt{1 - \sin^2(\phi) \cos^2(\varphi) + \left(\frac{d}{r_{2j}}\right)^2 - \frac{2d}{r_{2j}} \cos(\phi) \cos(\varphi)}, \quad (18)$$

and then

$$k_{II} = \sqrt{\frac{4 \frac{r_{2j}}{r_{1i}}}{\left(1 + \frac{r_{2j}}{r_{1i}}\right)^2 + \left(\frac{z}{r_{1i}} - \frac{r_{2j}}{r_{1i}} \cos(\phi) \sin(\varphi)\right)^2}}. \quad (19)$$

Both coils have one or more turns, and since the equations derived above apply only to the arrangement of simple loops, it is not possible to apply them directly. The calculation is divided into $N_1 N_2$ sub-steps, where the mutual inductance of all turn's combinations of the first and second coil is determined. Substituting (17) into (20) we get the total mutual inductance.

$$M = \sum_{i=1}^{N_1} \sum_{j=1}^{N_2} M_{ij} \quad (20)$$

As mentioned above, spiral-shaped coils are rather used for high-frequency applications, while rectangular or square-shaped coils are suitable for applications operating at lower frequencies. On the one hand, there are no electric field gradients, the coupling is rather inductive, and on the other hand we try to make maximum use of the built-up areas to maximize the coupling factor between the coils. As in the previous case, Biot-Savart's law can be applied here as well.

However, since it is not a circular coil, the advantages of the cylindrical coordinate system cannot be used, and the calculation is considerably complicated. To avoid confusing relationships, we will only consider the coaxial arrangement of two coils. These can have different geometries and different numbers of turns.

Figure 10 shows the real and simplified geometry of the coil, on which the derivation of the calculations will be performed. As can be seen in the figure on the left (**Figure 10**), the actual turns have different lengths at the same position, making the whole arrangement asymmetrical. The analytical solution of the field would then be very complicated and quite confusing.

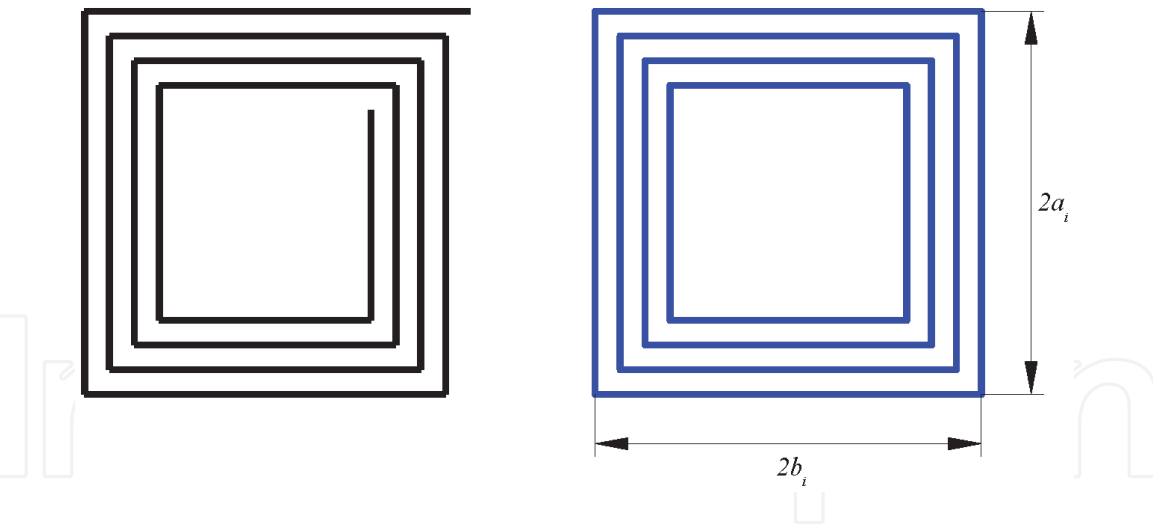


Figure 10.
 Real (left) and simplified (right) geometry of the coil with rectangular shape.

Thanks to the equivalent replacement of individual turns with concentric rectangles/squares, we are able to solve the magnetic field around the coil relatively easily and analytically. **Figure 11** indicates the relative position of two coils of different dimensions and number of turns spaced by a length z .

Let us now focus on the i -th turn of the lower coil and the j -th turn of the upper coil. The magnetic field passing through the upper coil (excited by the lower coil) can be calculated from (21), where B_{iz} is the induction of the magnetic field in the z -axis.

$$\phi_{ij} = \int_{S_j} \vec{B}_i d\vec{S}_j = \int_{S_j} B_{iz} dS_j = \int_{S_j} B \cos(\theta) dS_j \tag{21}$$

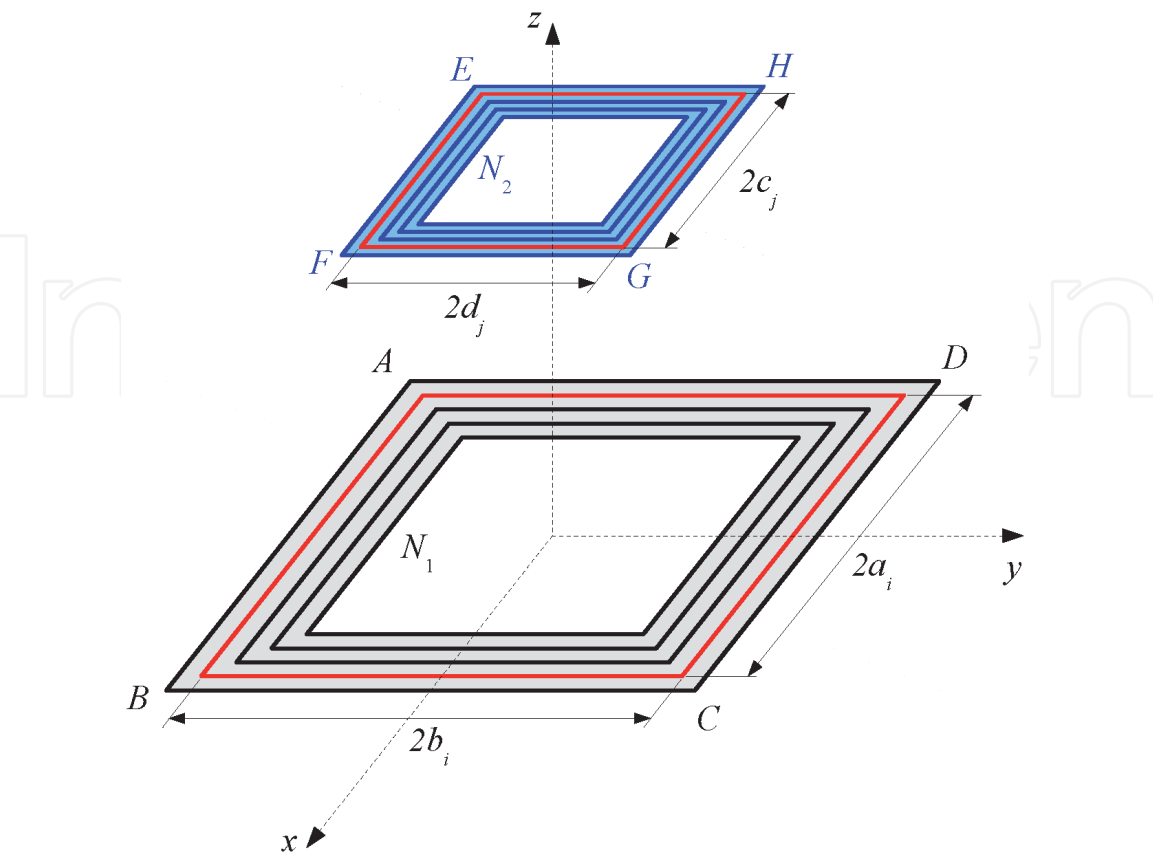


Figure 11.
 Simplified situation for determination of mutual inductance between rectangular coils.

Furthermore, we can use Biot-Savart's law to determine the increment of the B_{CD} magnetic field from the current-carrying segment of the i -th turn as:

$$\vec{B}_{CD} = \frac{\mu_0}{4\pi} \frac{1}{\sqrt{(b_i - y)^2 + z^2}} \left[\frac{a_i + x}{\sqrt{(b_i - y)^2 + z^2 + (a_i + x)^2}} + \frac{a_i - x}{\sqrt{(b_i - y)^2 + z^2 + (a_i - x)^2}} \right], \quad (22)$$

An if applicable next equation:

$$\cos(\theta) = \frac{b_i - y}{\sqrt{(b_i - y)^2 + z^2}}, \quad (23)$$

For z -component of B_{CD} induction we can derive (24)

$$B_{CD-z} = \frac{\mu_0}{4\pi} \frac{b_i - y}{(b_i - y)^2 + z^2} \left[\frac{a_i + x}{\sqrt{(b_i - y)^2 + z^2 + (a_i + x)^2}} + \frac{a_i - x}{\sqrt{(b_i - y)^2 + z^2 + (a_i - x)^2}} \right]. \quad (24)$$

For the total coupled magnetic flux with CD segment, we can write integral in form of (25)

$$\Psi_{CD-z} = \int_{-d_j}^{d_j} dy \int_{-c_j}^{c_j} B_{CD-z} dx. \quad (25)$$

As shown in [24], although the solution of the integral (25) is more complicated, we obtain a purely analytical relation (26).

$$\begin{aligned} \Psi_{CD-z} = \frac{\mu_0}{4\pi} & \left[\Gamma_1 - (a_i + c_j) \tanh^{-1} \left(\frac{a_i + c_j}{\Gamma_1} \right) - \Gamma_2 \right. \\ & + (a_i - c_j) \tanh^{-1} \left(\frac{a_i - c_j}{\Gamma_2} \right) - \Gamma_3 \\ & + (a_i + c_j) \tanh^{-1} \left(\frac{a_i + c_j}{\Gamma_3} \right) - \Gamma_4 \\ & \left. - (a_i - c_j) \tanh^{-1} \left(\frac{a_i - c_j}{\Gamma_4} \right) - \Gamma_4 \right] \end{aligned} \quad (26)$$

In relation (26) it is still necessary to substitute substitution (27).

$$\begin{aligned} \Gamma_1 &= \sqrt{(b_i + d_j)^2 + z^2 + (a_i + c_j)^2} \\ \Gamma_2 &= \sqrt{(b_i + d_j)^2 + z^2 + (a_i - c_j)^2} \\ \Gamma_3 &= \sqrt{(b_i - d_j)^2 + z^2 + (a_i + c_j)^2} \\ \Gamma_4 &= \sqrt{(b_i - d_j)^2 + z^2 + (a_i - c_j)^2} \end{aligned} \quad (27)$$

The magnetic flux from the other segments (AB, BC and DA) can be easily determined using the same relations. For example, to calculate the segment BC, it is

enough to swap a_i with b_i and c_j with d_j in (26). Due to the symmetry, the Eq. (28) will apply.

$$\Psi_{AB-z} = \Psi_{CD-z} a \Psi_{DA-z} = \Psi_{BC-z} \tag{28}$$

And because in the case of a unity current considering mutual inductance between the i -th and j -th turns next equation is valid (29)

$$M_{ij} \Psi_{ij-z} = \Psi_{AB-z} + \Psi_{BC-z} + \Psi_{CD-z} + \Psi_{DA-z}, \tag{29}$$

the total mutual inductance of both coils is based on (30).

$$M = \sum_{i=1}^{N_1} \sum_{j=1}^{N_2} M_{ij} \tag{30}$$

3.2 Self-inductance of planar coils

To calculate the intrinsic inductance of a planar coil, it is possible to find simple approximation relations, which are suitable for consequent mathematical derivations. However, their big disadvantage is only an approximate calculation with an often-indeterminate error. In addition, the relationships apply only to coils with an equilateral plan

$$L_a = \frac{\mu_0}{2} N^2 D_{AVG} K_1 \left[\ln \left(\frac{K_2}{p} \right) + K_3 p + K_4 p^2 \right] \tag{31}$$

Here, for p , the turn's filling factor on the coil surface and D_{AVG} is represented as the mean winding diameter.

$$p = \frac{D_2 - D_1}{D_2 + D_1}, D_{AVG} = \frac{D_2 + D_1}{2} \tag{32}$$

In **Table 3**, the coefficients depending on the approximated coil geometry are calculated (**Figure 12**). The coefficients $K_1 - K_4$ must always be selected according to the current geometry.

Coil shape	K_1	K_2	K_3	K_4
Circular	1	2.46	0	0.2
Squared	1.27	2.07	0.18	0.13
Hexagonal	1.09	2.23	0	0.17
Octagonal	1.07	2.29	0	0.19

Table 3.
Estimated coefficients for identification of the shape of planar coil.

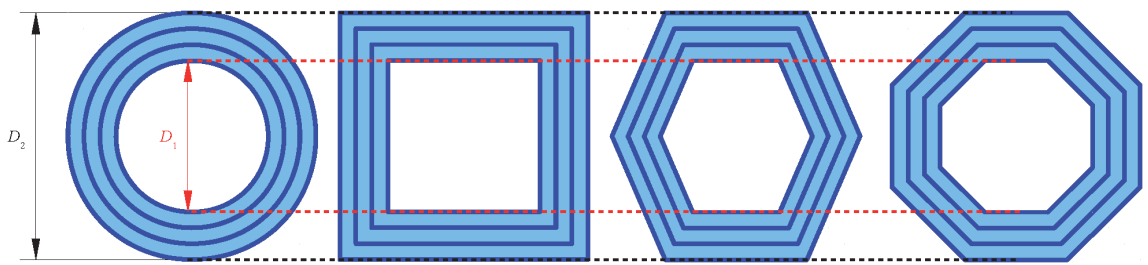


Figure 12.
Allowed degenerations of the coil's geometry for calculation of inductance using Eq. (31).

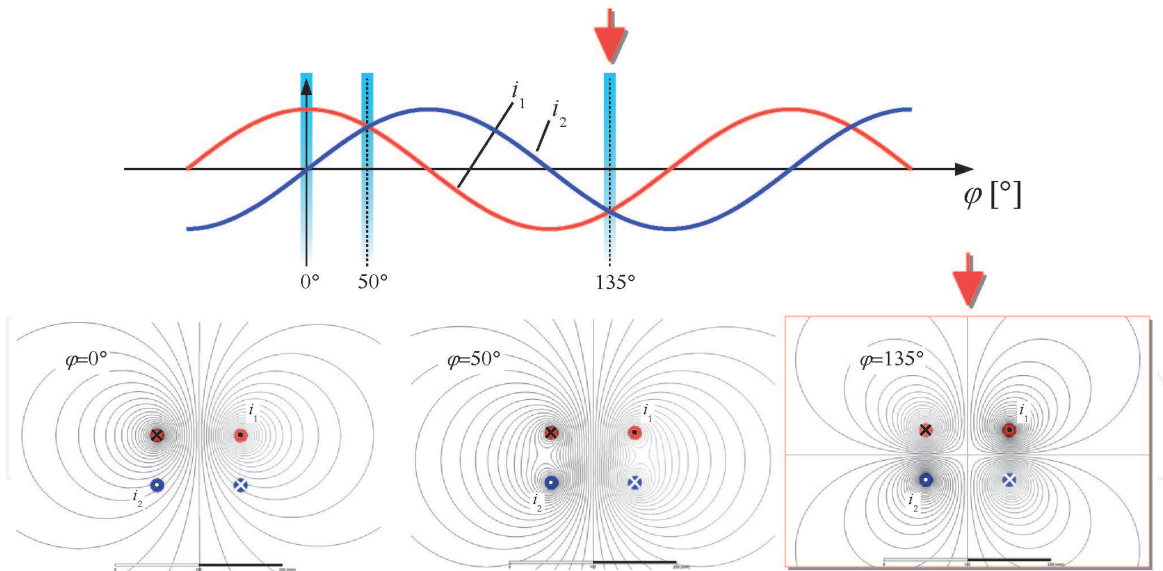


Figure 13.
Time-varying electromagnetic field around the system of coupling coils.

3.3 Coupling coefficient

The magnetic coupling between two coils is formed by a magnetic field, which is generated by a transmitting coil. For many reasons, this array can never be coupled to the receiving coil in its full size, and the larger the array, the better the coupling is achieved. This phenomenon is described by the so-called coupling factor k , which takes values from 0 to 1. The coupling factor depends on the geometry of the coils and their mutual position. Many authors mistakenly qualify the degree of coupling of two coils based on the shape of the electromagnetic field in their surroundings. This approach leads to misinterpretations mainly because the field itself is variable in time and looks different at different times. The situation is clearly shown in **Figure 13**.

For example, in S-S compensation, the currents flowing through the primary and secondary windings are time-shifted by 90 electrical degrees. While at the instant $j = 0^\circ$ and $j = 50^\circ$ the coils appear to be coupled, while at the instant $j = 135^\circ$ they are without mutual coupling according to the shape of the field (**Figure 13**). The only reliable way to determine the coupling factor is to apply relation (33).

$$k = \frac{M}{\sqrt{L_1 L_2}} \quad (33)$$

We will explain some relationships on a simple example, in which we determine the coupling factor of two coaxially placed coils of circular shape with planar design. We will perform the calculation on three similar geometries (**Figure 12** left), where the first pair of coils will have an inner diameter $D_1 = 100$ mm and an outer diameter $D_2 = 200$ mm, whereby the other two pairs of coils will have the same dimensions multiplied by two and three. The number of turns is the same $N = 5$ for all cases. The coupling factor will be plotted for a distance $z = 5 \div 300$ mm.

As shown in **Figure 14** (left), changing the distance of the coils, the coupling factor k decreases rapidly, while the rate of this decrease is highly dependent on the respective geometry of the coils. It is therefore better to choose larger coil dimensions to improve the coupling at higher distances. In **Figure 14** (right) we see the effect of the misalignment of the coils in the x -axis at their constant distance in the z -axis. The coupling factor is somewhat less sensitive to this method of deflection.

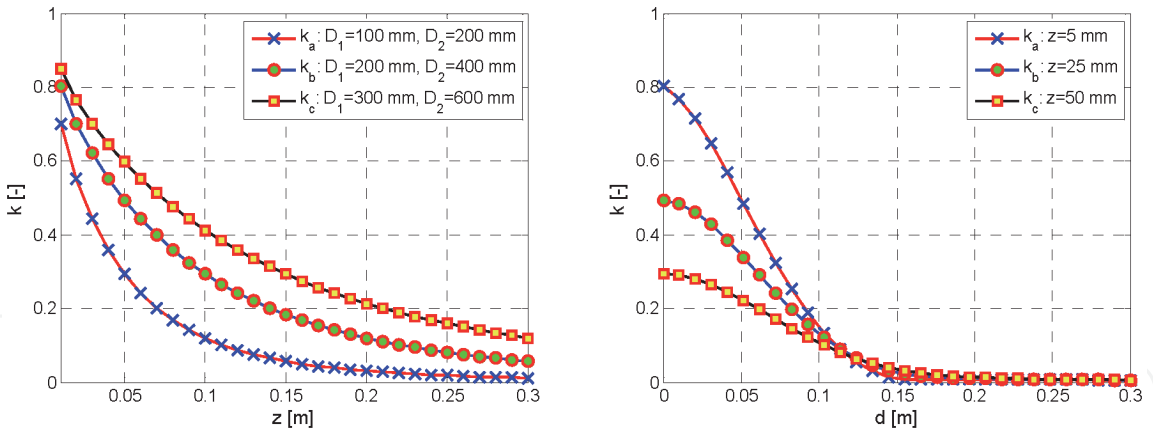


Figure 14.
Dependency of coupling factor on the geometrical properties of the coil and mutual distance (left) and mutual radial displacement (right).

3.4 Parasitic capacitance of the coil

For the capacitance between two turns with mean radius r_i and the distance between individual turns e we can write the relation (34).

$$C_{ip} = \frac{3}{2} \frac{\pi \epsilon_0 r_i}{\ln \left(\frac{e}{r_v} \right)} \tag{34}$$

If we have a coil with N turns, the total parasitic capacitance must be determined as

$$C_{Cp} = \left(\sum_{i=1}^{N-1} \left[\left(\frac{3}{2} \frac{\pi \epsilon_0 r_i}{\ln \left(\frac{e}{r_v} \right)} \right)^{-1} \right] \right)^{-1} . \tag{35}$$

Further to the pattern of **Figure 12** to the left we denote the outer diameter $D_2 = 2r_2$, we get the modification (35) in the form (36).

$$C_{Cp} = \left(\sum_{i=1}^{N-1} \left[\frac{2}{3} \frac{\ln \left(\frac{e}{r_v} \right)}{\pi \epsilon_0 [r_2 - (r_v + e/2)i]} \right] \right)^{-1} \tag{36}$$

The geometric arrangement, according to which (36) can be easily applied, can be seen in **Figure 15**.

Unlike high-frequency systems, at lower operating frequencies, ropes with insulated conductors are used almost exclusively. The reason is the lower influence of parasitic capacitances and especially the better current utilization of the coil.

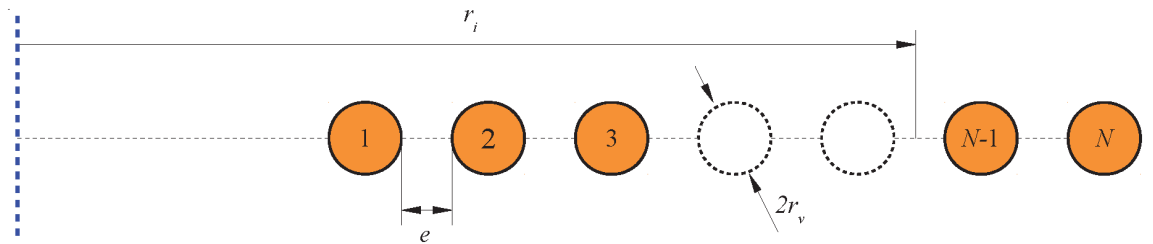


Figure 15.
Situation of the coil's turn placement for the calculation of parasitic capacitance.

3.5 Series parasitic resistance of the coil

The series resistance of the coil is one of the most critical parameters of the system with the greatest influence on its operational efficiency and it is therefore very important to know this value as accurately as possible. We can start with the general relation for resistance according to (37).

$$R = \rho \frac{l}{S} \quad (37)$$

So far, we will not consider temperature or frequency dependences. While the effective area of the conductor S depends only on the current load, the length l already depends on the geometric shape of the coil. As mentioned earlier, spiral planar coils of solid conductor are more suitable for high frequency applications.

3.5.1 Series parasitic resistance of spiral planar coil

In this approximation, we will only talk about coils wound with a copper conductor of circular cross-section. These with their shape most closely resemble parts of the Archimedean spiral, where regarding **Figure 9** on the right we can denote the inner radius as r_{1A} and the outer radius as r_{1B} . The distance between the individual turns e and the number of turns N will be constant. As mentioned earlier, the key length here is played primarily by the length of the conductor of the wound coil. We can write an equation for the Archimedean spiral in polar coordinates

$$r = r_{1A} + \zeta \cdot \varphi, 2\pi\zeta = e. \quad (38)$$

The length of the spiral thus described can be determined by integration (39)

$$l_c = \int_0^{2\pi N} \sqrt{\left(r_{1A} + \frac{e}{2\pi}\varphi\right)^2 + \left(\frac{dr}{d\varphi}\right)^2} d\varphi, \quad (39)$$

However, the disadvantage remains the fact that the integral (39) cannot be solved analytically. It is therefore necessary to integrate numerically for the calculation. It is also possible to use an approximation relation for an approximate calculation

$$l_c = \frac{1}{2} \left[\frac{e}{2\pi} (2\pi N)^2 + 2\pi N \left(\sqrt{r_{1A}^2 + \left(\frac{e}{2\pi}\right)^2} + r_{1A} \right) \right], \quad (40)$$

or simplification by means of an average radius, see (41). The calculation is then very fast and convenient.

$$l_c = 2\pi N \frac{r_{1A} + r_{1B}}{2} \quad (41)$$

In addition, high-frequency applications require winding of a solid conductor to reduce the parasitic capacitance of the coil. Therefore, if we consider the effect of the skin effect, we can adjust (37) to the shape (44) for a conductor with radius r_v . Here for l_c we use one of the equations (39)–(41).

$$R_{c-AC}(\omega) = \begin{cases} \rho_{Cu} \frac{l_c}{\pi \delta (2r_v - \delta)}, \wedge \delta \leq r_v \\ \rho_{Cu} \frac{l_c}{\pi r_v^2} \left(1 + 0.0021 \left(\frac{r_v}{\delta} \right)^4 \right), \wedge \delta > r_v \end{cases}, \delta = \sqrt{\frac{2\rho_{Cu}}{\omega\mu_0}} \quad (42)$$

3.5.2 Series parasitic resistance of rectangular planar coil

The coil has rectangular turns to achieve maximum inductance (**Figure 16**). If we denote the external dimensions of the coil by the letters a and b and consider the spacing between the individual turns e constant, we can write Eq. (43) for the resulting resistance.

$$R_{co-DC}(\vartheta) = \frac{\rho(\vartheta)}{S_{Cu}} \left(2 \sum_{i=1}^N [a + b - [1 + (i - 1)4]e] \right) \quad (43)$$

Furthermore, if we choose an insulated RF cable with wires whose diameter is much smaller than the penetration depth d , we can certainly rule out the effect of the skin effect. We are then talking about a conductor with an effective cross section S_{Cu} , whose frequency dependence is caused only by the phenomenon of proximity. With a few modifications, it can be further simplified (44) by removing the summation into the shape

$$R_{co-DC}(\vartheta) = \frac{\rho(\vartheta)}{S_{Cu}} 2N[a + b + e(1 - 2N)] \quad (44)$$

In all the cases described, the turns are evenly distributed in one layer with a constant e . The current flowing through the coil thus has the same direction in all turns and generates a magnetic field with lower intensity on the external turns and higher intensity on the internal turns. This magnetic field induces eddy currents into all coil turns and thus increases its overall resistance. This process is commonly referred to as the proximity phenomenon and can be conveniently calculated from the relationship for eddy current losses in individual parallel conductors. For one fiber of diameter d_s of an insulated cable of length l , exposed to an external magnetic field B of a harmonic waveform of angular frequency ω , we can write

$$R_{prox} = \frac{\pi l d_s^4}{64\rho} \omega^2 B^2 \quad (45)$$

If we use a cable made of n_f insulated conductors for winding of the coil, it is possible to adjust e_g (44) to the shape (46) by counting (42).

$$R_{cn-AC}(\vartheta, \omega) = \left(\frac{\rho(\vartheta)}{n_s \frac{\pi d_s^2}{4}} + n_s \frac{\pi d_s^4}{64\rho(\vartheta)} \omega^2 B_{str}^2 \right) nN \left[a + \tan\left(\frac{\pi}{n}\right)(1 - N)e \right] \quad (46)$$

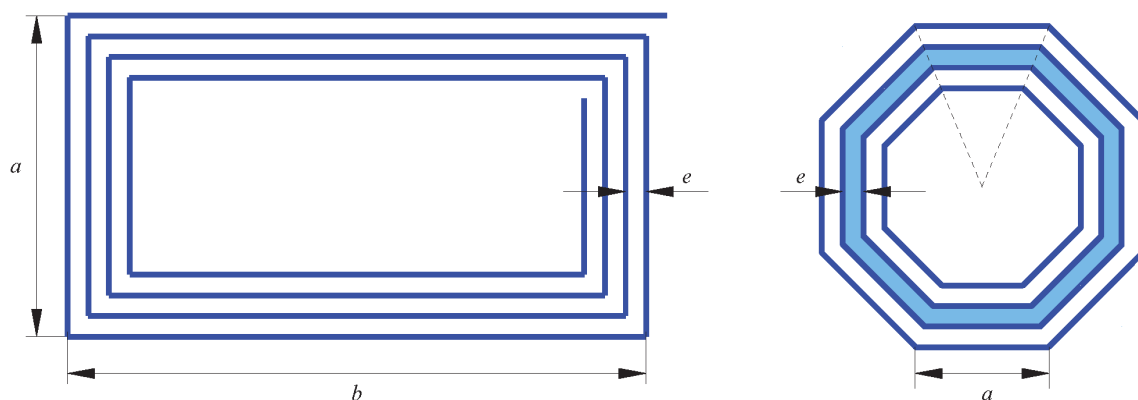


Figure 16.
Rectangular coil identification for the calculation of parasitic resistance.

To illustrate, we will analyze the following geometry. We consider a square coil with an outer edge of length $a = 500$ mm and $N = 20$ turns. We choose the spacings between the turn's axes $e = 8$ mm. The maximum operating frequency is $f = 300$ kHz, while for a nominal current $I = 5$ A we choose a current value $J = 7$ MA/m². The calculation is valid for a temperature of 20° C. **Figure 17** shows the dependence of the DC resistance from (44), the AC resistance from (46) and the total resistance. The minimum value of the number of RF wires is determined from (47) and corresponds to the rated current load ($n_{s-\min} = 61$).

$$n_{s-\min} \doteq \frac{4I}{\pi d_s^2 J} \quad (47)$$

By further increase of the number of wires, we therefore only increase the current possibilities of the coil.

As the number of wires increases, the DC resistance R_{DC} decreases sharply, but the AC resistance R_{prox} also increases. The optimum can be found by solving Eq. (48).

$$\frac{d}{dn_s} \left[\left(\frac{\rho(\vartheta)}{n_s \frac{\pi d_s^2}{4}} + n_s \frac{\pi d_s^4}{64 \rho(\vartheta)} \omega^2 B_{str}^2 \right) nN \left[a + \tan \left(\frac{\pi}{n} \right) (1 - N)e \right] \right] = 0 \quad (48)$$

The condition is met just when it applies

$$n_s \doteq \frac{16 \rho(\vartheta) \sqrt{a - e(1 + N) \tan \left(\frac{\pi}{n} \right)}}{\pi \sqrt{\omega^2 B_{str}^2 d_s^6 [a - e(1 + N) \tan \left(\frac{\pi}{n} \right)]}}. \quad (49)$$

After substituting, we get the value $n_s = 328$ wires, which corresponds to the value in **Figure 17**.

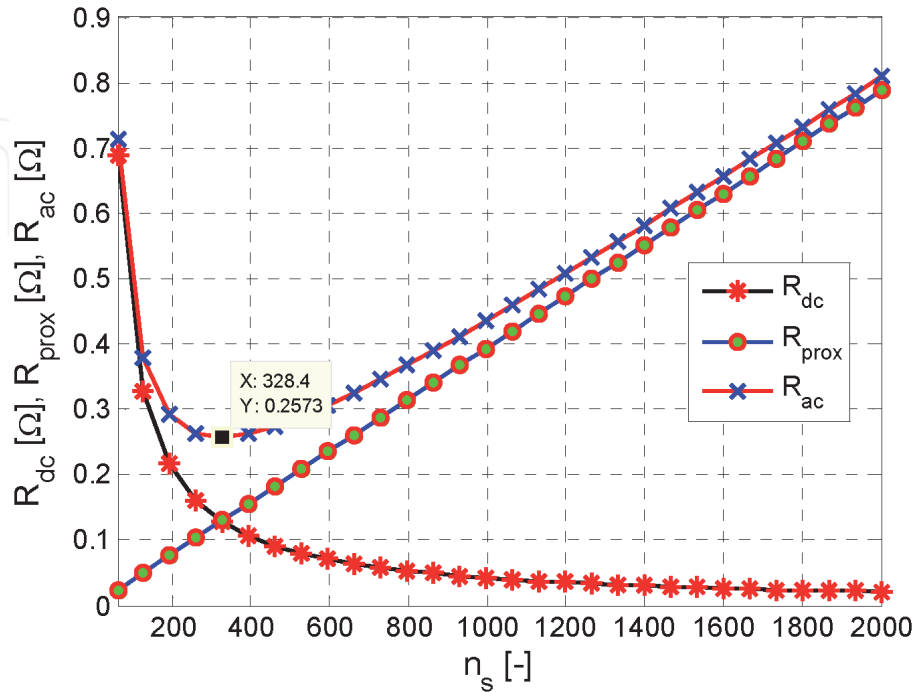


Figure 17.
Nomogram for the calculation of the optimal number of litz wire.

3.5.3 Series parasitic resistance of rectangular planar coil

The most general definition of the quality factor is based on the ratio of accumulated and lost energy in the investigated passive component. For AC supply we can write (50), where the influence of the electric field prevails in the case of a capacitor and the influence of the magnetic field in the case of a coil.

$$Q = \omega \frac{|W_{mg} - W_{el}|}{P_j} = \frac{1}{R} \sqrt{\frac{L}{C}}. \tag{50}$$

If we consider an ideal coil (R-L circuit) without parasitic capacitance, we get a quality factor such as

$$Q = \omega \frac{|W_{mg}|}{P_j} = \omega \frac{\frac{1}{2} L I_m^2}{\frac{1}{2} R I_m^2} = \frac{\omega L}{R}, \tag{51}$$

For more complicated circuits, such as components with parasitic effects, we can also use the relation to calculate the quality factor

$$Q = \frac{\left| I \left\{ \frac{\omega L - jR}{\omega C_p (R + j\omega L) - j} \right\} \right|}{R \left\{ \frac{\omega L - jR}{\omega C_p (R + j\omega L) - j} \right\}} = \frac{\left| \frac{\omega (L - C_p R^2 - \omega^2 C_p L^2)}{1 + \omega^2 C_p [C_p R^2 + L(\omega^2 C_p L - 2)]} \right|}{\frac{R}{C_p^2 \omega^2 R^2 + (\omega^2 C_p L - 1)^2}} = \left| \frac{\omega L}{R} - \frac{\omega C_p (R^2 + \omega^2 L^2)}{R} \right|. \tag{52}$$

The first part of the result of Eq. (52) corresponds to the quality factor of the individual coil, the second part then respects the effect of parasitic capacitance between the turns. A closer look reveals that there is a frequency at which both parts are equal and (52) gives zero result. This frequency is often referred to as the coil's own resonant frequency

$$f_{r-self} = \frac{1}{2\pi \sqrt{LC_p}}. \tag{53}$$

The situation is indicated in **Figure 18** on the right, where the dependency of the quality factor and the character of the resulting reactance on the frequency is plotted for selected values of the parasitic capacity of the inductance and the series

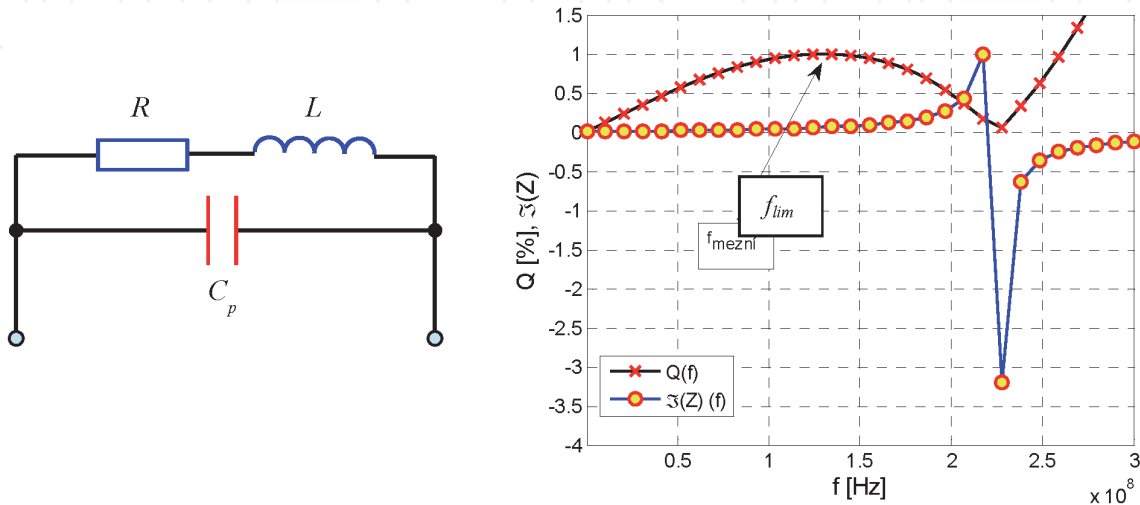


Figure 18.
Parasitic components of the coil and the quality factor characteristic.

resistance of the coil ($C_p = 5 \text{ pF}$, $L = 0.1 \text{ mH}$, $R = 1 \Omega$). As can be seen from the figure, when reaching the natural resonant frequency of the $f_{r\text{-self circuit}} \gg 225 \text{ MHz}$, the quality factor is equal to zero and at the same time the inductive character of the reactance changes to capacitive character. For this reason, we always try to operate the coil at a frequency much lower than the self-resonant frequency.

4. Practical design approach for industrial wireless power transfer charging system

4.1 Power electronic system configuration

Electrical engineers responsible for the design of the wireless transfer chargers must consider standard grid network connection during design process. Because many issues are nowadays address on the quality of the supply grid, the main goal during design of any power electronic system is to achieve the best performance related to the power factor parameter at any power consumption of the system. In addition to this fact, it is also required to have fully symmetrical 3-phase current with as low total harmonic distortion as possible [25–28].

$$THDi = \frac{\sqrt{\sum_{\mu=2} i_{ac(\mu)}^2}}{i_{ac(1)}} 100 \tag{54}$$

Regarding above mentioned facts, each power electronic system, which must undergo strict normative given on the qualitative indicators of the grid variables, must be equipped with input active or passive power factor corrector (PFC) and total harmonic distortion correction (THDC). These blocks are consequently followed by diode rectifier, dc/dc converter (step-up or step-down) and the voltage source inverter. Such power electronic system configuration is robust and verified by many similar applications (mostly power supplies and battery chargers). The main negative drawback of such concept lies in higher price and build-in dimensions along with the increase in power rating. This topology should therefore be recommended for low or medium power WPT chargers (Figure 19 – blue blocks).

Second group of WPT chargers considering the value of power delivery is medium to high power concepts. Here it is recommended to use the configuration composed of input filter (inductive – designed as distribution transformer for

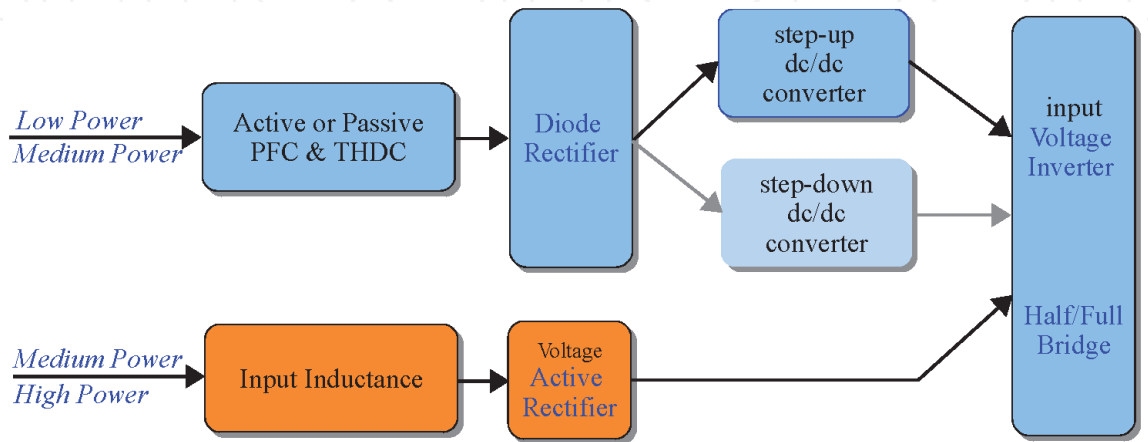


Figure 19. Power electronics configuration on the primary side of the wireless power charger indicating differences related to the level of the power transfer.

example), followed by the active PFC/THDC rectifier supplying the voltage source inverter (VSI). For both cases (low or high power) the VSI is sourcing primary/ transmitting coil with relevant compensation. This configuration of power electronic system (**Figure 19** – orange blocks) is providing low ripple input current with sinusoidal character, low THDi, excellent power factor and controllable output voltage. Therefore, it is not required to implement another dc/dc converter stage within the system [29–31].

The recommended topologies are summarized in **Figure 19** according to system dedicated power level.

The concept of power electronic system for the secondary side also differs based on the type of the load, and level of the power delivery. Basically, it consists of secondary side coil equipped by relevant compensation, passive or active rectifier and dc/dc converter stage providing required functionality of the charger.

Finally, the system connection to the grid considering all the power levels established as WPT categories by SAE TIR J2954 is seen in conceptual layout shown in **Figure 20**, valid especially for central Europe [32–35].

A more detailed example above described solution, which could meet all necessary technical requirements on high power applications and simultaneously having excellent operational properties, is seen in **Figure 21**.

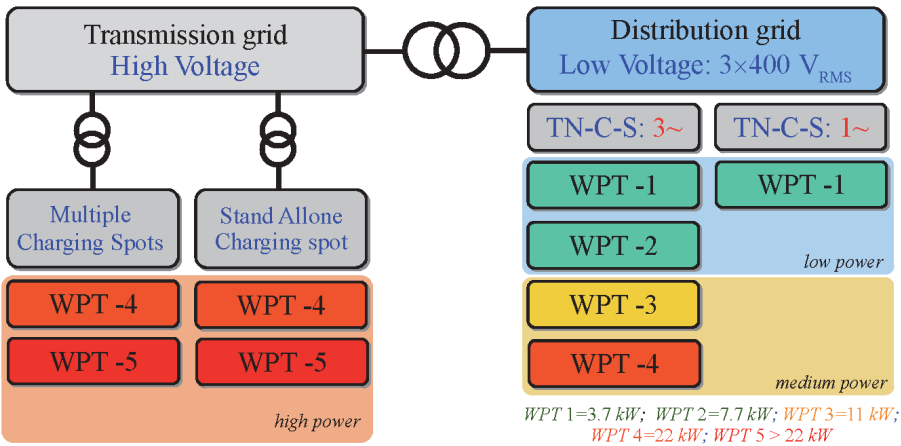


Figure 20.
WPT system categories – Connection to the grid.

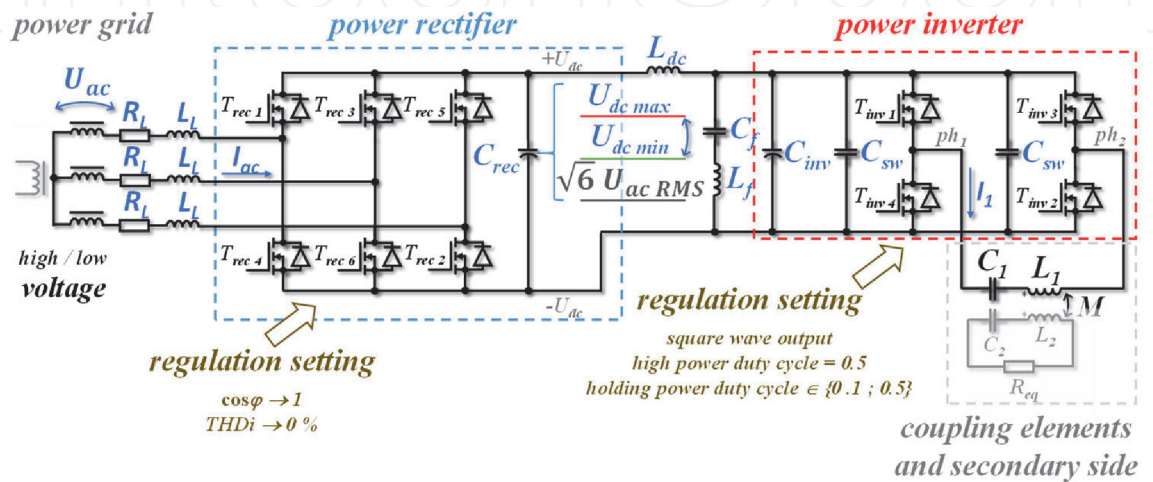


Figure 21.
Recommended system configuration for high power application.

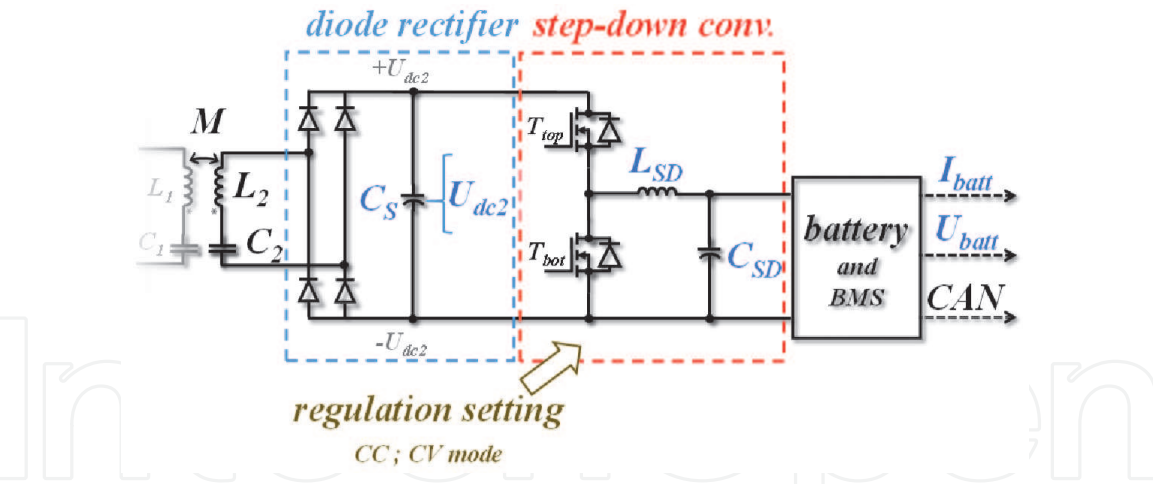


Figure 22.
Recommended system configuration of secondary side for high power application.

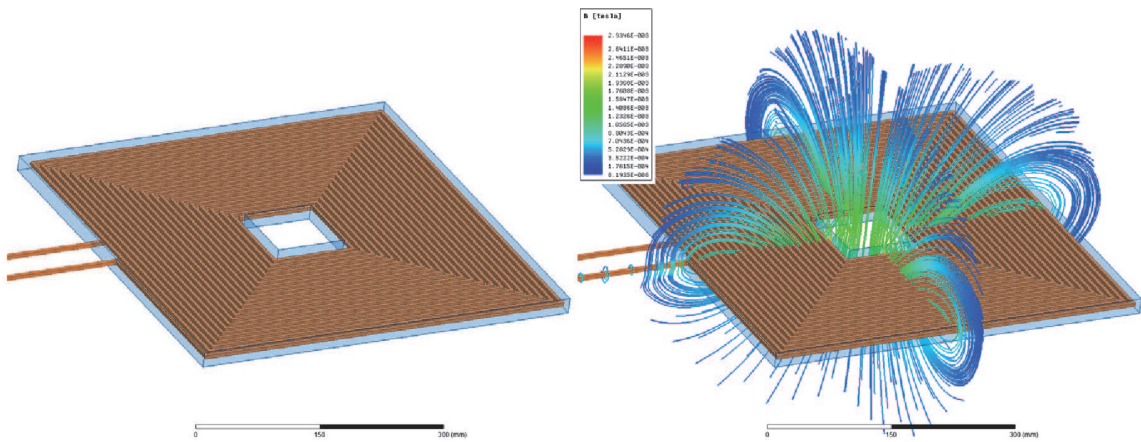


Figure 23.
Proposed coupling coil (left) and its magnetic field (right).

Here the distribution transformer is presented as the grid source, followed by active rectifier, which is responsible for regulation of PF and THDi. Then full bridge inverter is used as VSI and supplies primary side coupling section.

The secondary side of the system shown in **Figure 21** is drawn in more detail in **Figure 22**. The secondary side coupling system is followed by full-bridge diode rectifier with filtering capacitor CS. Then the dc/dc step-down converter (SD) providing required charging algorithm (mostly CC&CV) is supplying the on-board battery pack.

Previously described concepts are representing the mostly used configurations of power electronic systems required for the design of the wireless power chargers suited for industrial and/or automotive applications.

4.2 Coupling elements design

The most important parameter in the design of coupling coils is undoubtedly the product of quality factors Q and coupling k . Its operating size is strongly dependent on various parameters (e.g. circuit topology, load size, coil distance, etc.) and therefore cannot be optimized directly. One option is to maximize the quality factor. To achieve maximum inductance, we make the coil as planar with square turns (**Figure 23**). In addition, due to the limitation of parasitic capacitance (we now neglect), we keep constant spacings between individual turns of size $\delta_v = 4$ mm. Geometric dimensions allow to wind about 26 turns. In addition, if we know the

operating frequency, we can determine the voltage drop and the current through the coil from the required power. It is necessary to design an effective winding cross section for this.

Regarding the available conductor cross-sections, a copper wire (2200 mutually insulated conductors) with a total cross-section of $S_v = 19.63 \text{ mm}^2$ was selected. The winding produced in this way eliminates the effect of the skin effect and the resulting resistance of the coil is therefore only affected by the phenomenon of proximity.

The coil has 22 turns, the calculation parameters being as follows. The self-inductance has a value $L = 147 \text{ } \mu\text{H}$ and the active resistance is $R = 0.19 \text{ } \Omega$.

4.3 Experimental set-up

In this case, we will focus on the WPT 1 category with an output of 3.7 kW. The experimental workplace consists of a programmable power supply, electronic load, precision power analyzer, oscilloscope, input inverter, output rectifier, additional resistors and the compensated LC circuit WPT itself. The measuring workplace is connected according to the functional diagram, see **Figure 24**. The determining factor in the selection of power components was the ability to work with a switching frequency from 200 kHz upwards. For this reason, a solution based entirely on SiC elements was chosen. The inverter is built on 1200 V JFET modules FF45R12J1W1_B11 (Infineon) with a type current of 45 A. Due to the low values of switching times of these modules, which are actually in the order of tens of nano-seconds, it is possible to minimize the effect of inverter dead times. The rectifier is based on a 1200 V diode SiC module APTDC20H1201G (Microsemi) with a type current of 20 A.

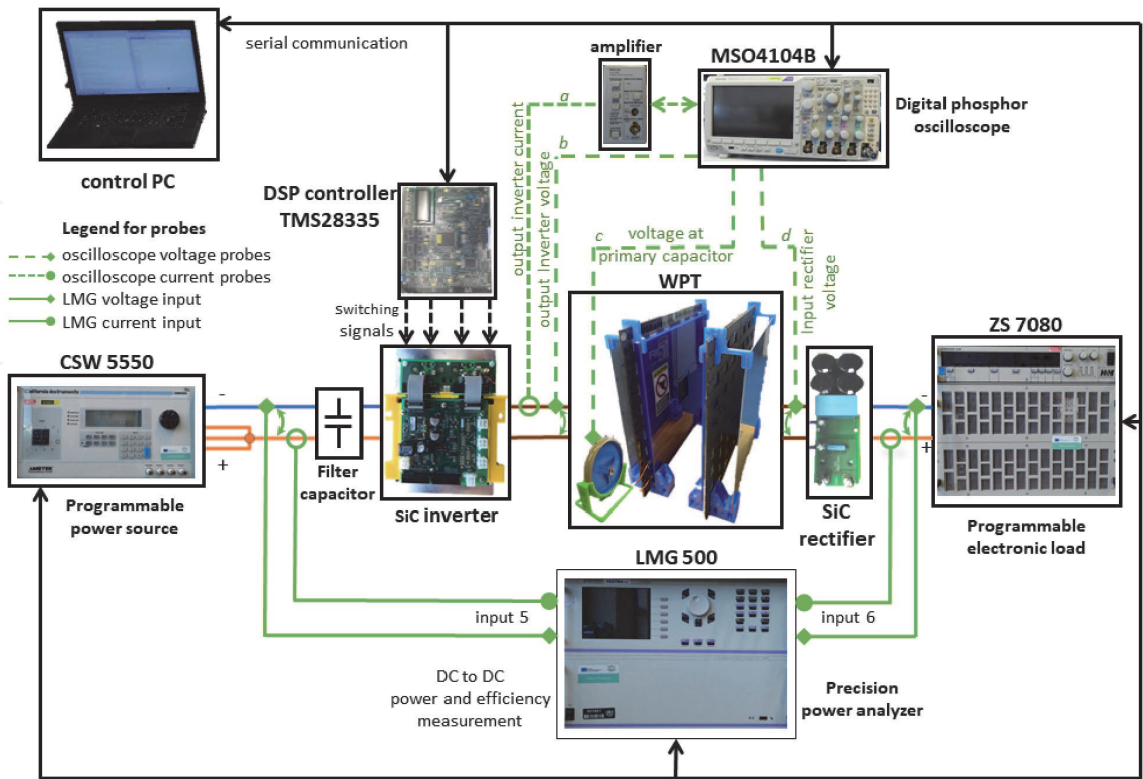


Figure 24.
Block diagram of the laboratory experimental set-up.

4.3.1 Measurement of the operational characteristics of series: series compensated system

On the primary side, a total of three quantities are measured with an oscilloscope. Probe “a” (THDP 0200) measures the output voltage of the inverter, probe “b” (current probe TCP 404 XL and amplifier TCPA 400) measures the primary current and probe “c” (P6015A) senses the voltage on the compensation capacitor. The secondary side is not measured by the oscilloscope at all in this configuration. Also, no resistors are connected here, and the system works directly into the ZS 7080. The applied oscilloscopic measurements on the primary side are rather indicative and do not serve to calculate the efficiency [35–37].

Figure 25 shows an oscillograph at a load power of 2678 W. The purple waveform represents the inverter output voltage, the light blue waveform the primary current waveform, and the blue waveform represents the voltage on the primary compensation capacitor (scale 1: 1000). The real elements (influenced by parasitics) of the WPT system are the main reason why the phase shift of voltage and current is non-zero (according to theoretical assumptions it should be close to zero).

A comparison of power (**Figure 26**) and efficiency (**Figure 27**) shows that the analytical models accurately describe the behavior of the system in a wide range of frequencies and loads.

4.4 Electromagnetic shielding application

Although the current system achieves very high efficiency even over long working distances, it is unsatisfactory due to hygienic limits and standards for EV charging. The main weaknesses are mainly the high switching frequency and the large intensities of the EM field. The magnetic field in the vicinity of both (optimally coupled) coils at a transmitted power of approx. 4000 W is plotted in **Figure 28**. The distribution of the field changes over time, and therefore each time point must be evaluated separately.

The picture shows a large scattering of the field into the surroundings, which must be avoided. Exact induction values at a specific distance from the center of the

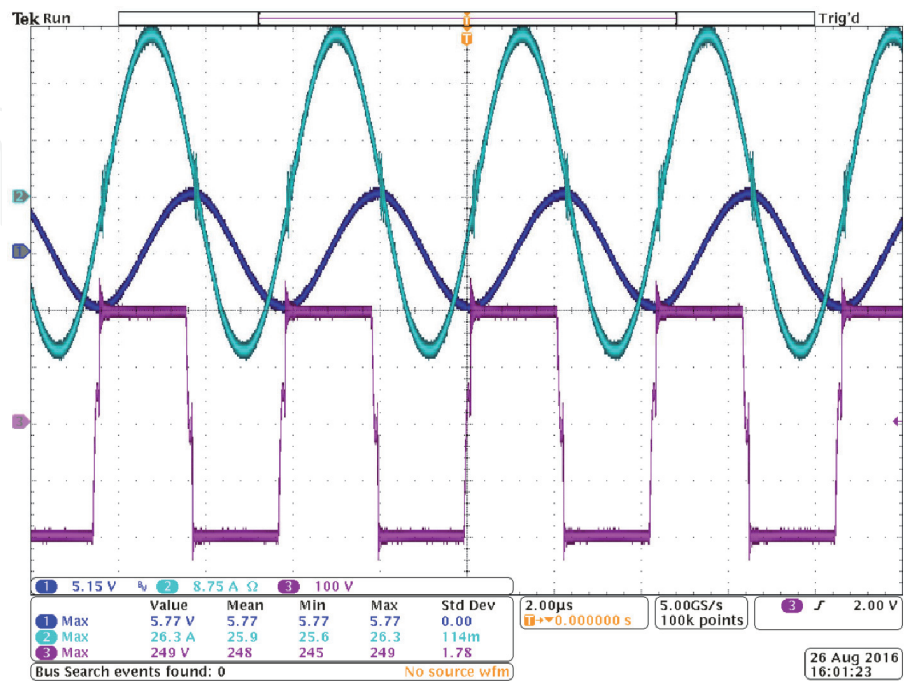


Figure 25.
Time waveforms of the primary side of tested WPT system during full load operation.

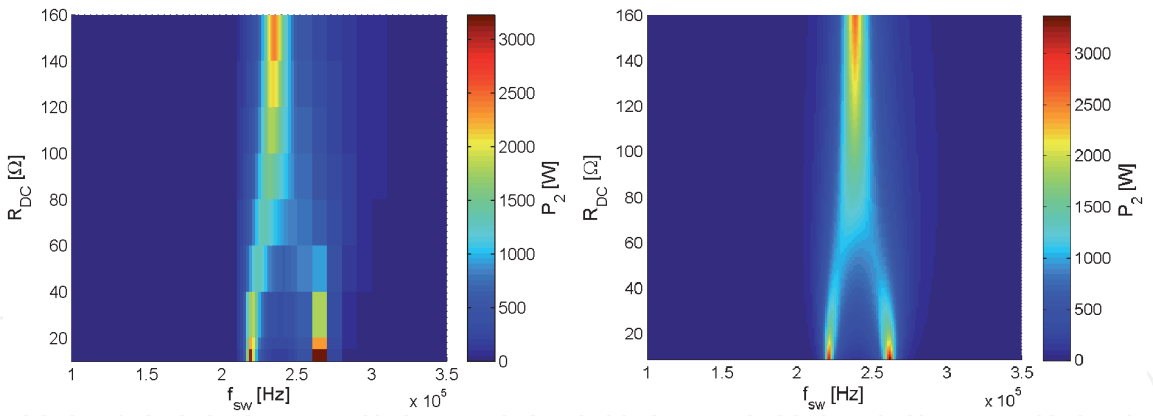


Figure 26.
Output power characteristic in dependency on load and operation frequency for measurement (left) and simulation (right).

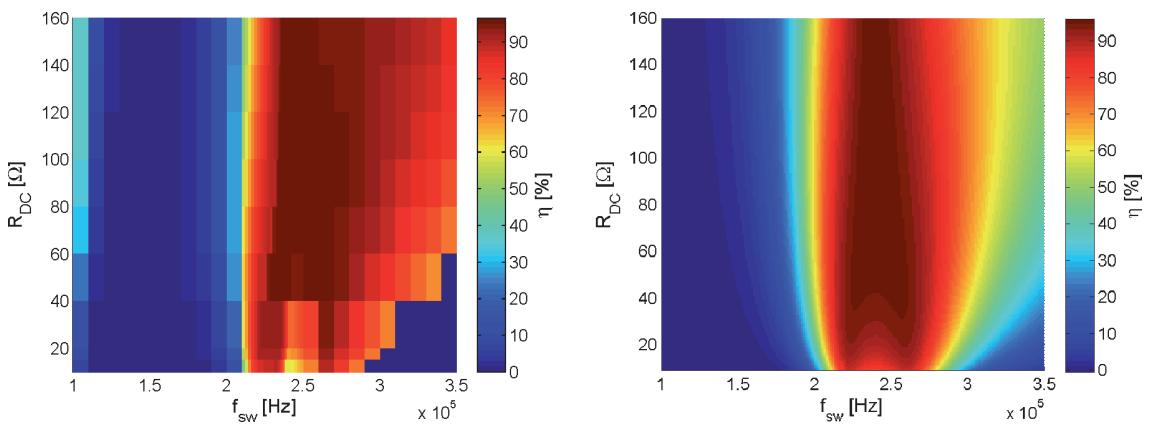


Figure 27.
Efficiency characteristic in dependency on load and operation frequency for measurement (left) and simulation (right).

coils can be obtained by introducing a spherical surface to which the EM field results are mapped. The radius of this area must be defined regarding the dimensions of the vehicle and the location of the coupling coils on its chassis. The key is especially the space in which exposed persons can normally occur. For practical reasons, therefore, it does not make sense to monitor the magnetic induction near both coils. For the sake of clarity, we state here (see **Figure 28**) the magnitude of

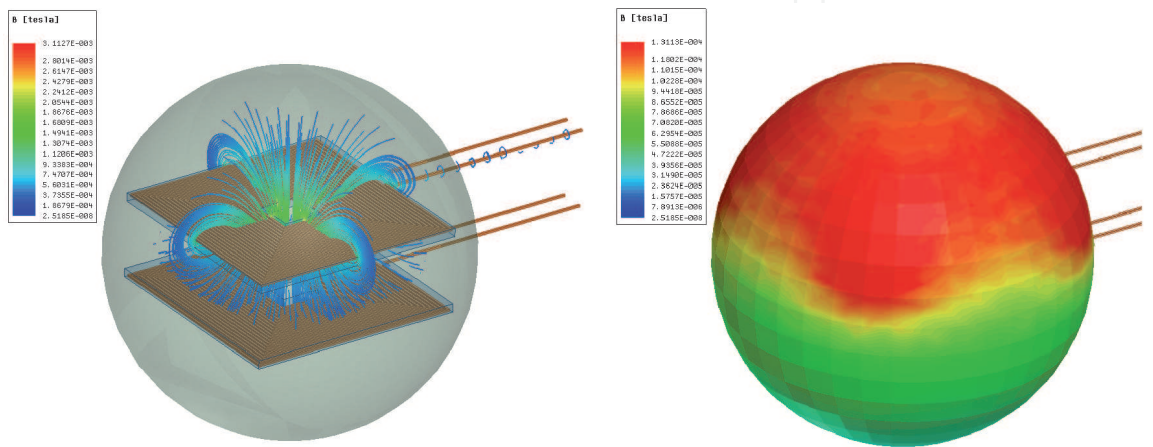


Figure 28.
Magnetic induction around system of unshielded coils.

induction on the sphere surface with a radius of 450 mm at the time ($j = 0^\circ$), when the current passes through only one coil.

Shielding can be realized by a matrix arrangement of ferrite cores lying on the back sides of both coils. The resulting magnetic field is directed into the main coupling space, while the interior of the vehicle remains protected. The material of the cores must correspond to the operating frequency and especially to the saturation at full load. Material N87 with relative permeability >1450 and operating frequency up to 500 kHz was selected for prototype. The size of the cores is 20x30x3 mm. Due to the high price and weight of the ferrite shield, it is reasonable to lighten its resulting pattern (not to occupy the full area of the coils). The finite element method will be used for this enabling to determine the intrinsic and mutual inductances of coupling coils, ferrite saturation and losses for any arrangement of ferrite cores.

Shielding consists of two functional elements (steps). The first is a ferrite array (plate) that holds the maximum amount of coupled flux and directs it for better bonding to the second coil. The second degree of shielding is an aluminum plate offset over a ferrite field. In the case of supersaturation of the ferrite core, this creates eddy currents that keep the field in the active space of both coils. The situation is indicated in **Figure 29** (left), the ferrite barrier (core) is drawn in gray. The aluminum shield is then shown by a solid plate near the ferrite core.

From **Figure 30** we can see the beneficial effect of shielding even better. Ferrite shielding almost completely shields the field above and below the coils. In this area, the hygienic limits are fully met and without the need for additional shielding.

The magnetic field of the coupling coils (**Figure 30** on the right) is now much better concentrated in the coupling space, which increases the probability of meeting the hygienic limits many times over.

4.4.1 Experimental analysis of the impact of shielding system

In order to verify the theoretical assumptions, an experimental prototype of a previously designed shielding was created. The photograph of the experimental workplace is evident from **Figure 31**.

The aim was to significantly reduce the switching frequency of the supply voltage and to suppress the emission of the EM field to meet the hygienic limits according to “ICNIRP 2010” [38, 39]. The operating parameters of the newly implemented prototype are quantified in **Table 4**. The values are valid for a working distance of 20 cm.

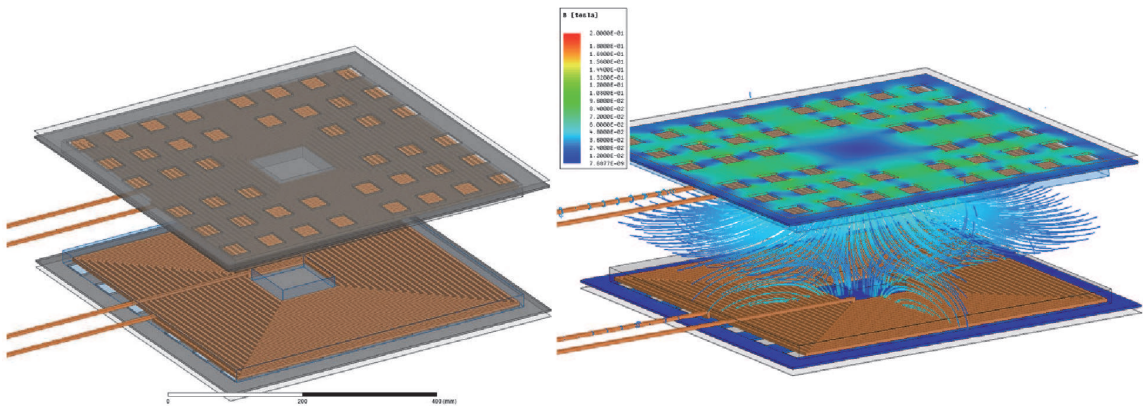


Figure 29.
Proposed electromagnetic shielding (left) and EM field distribution (right).

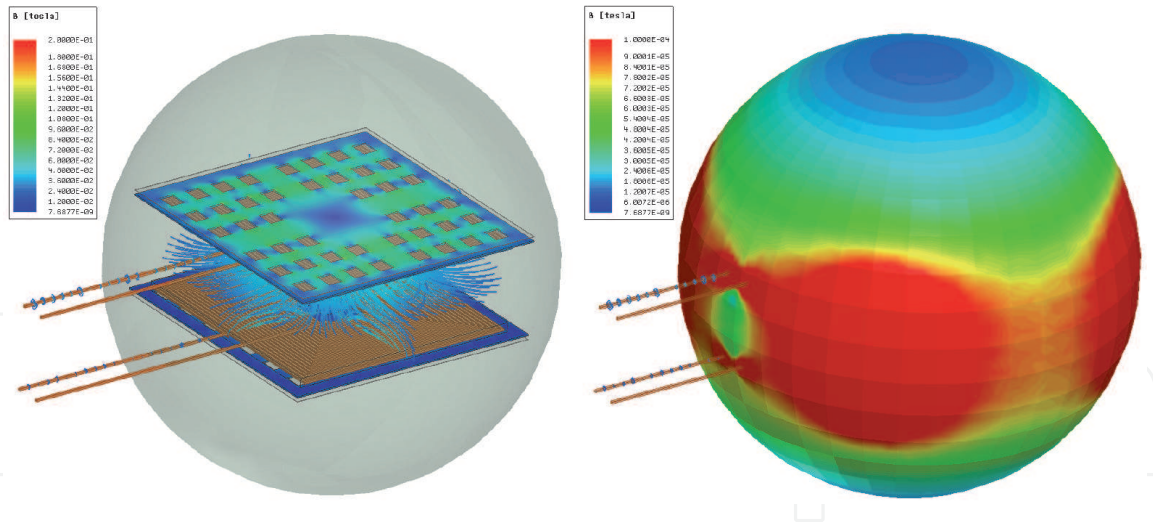


Figure 30.
Magnetic induction around shielded system.

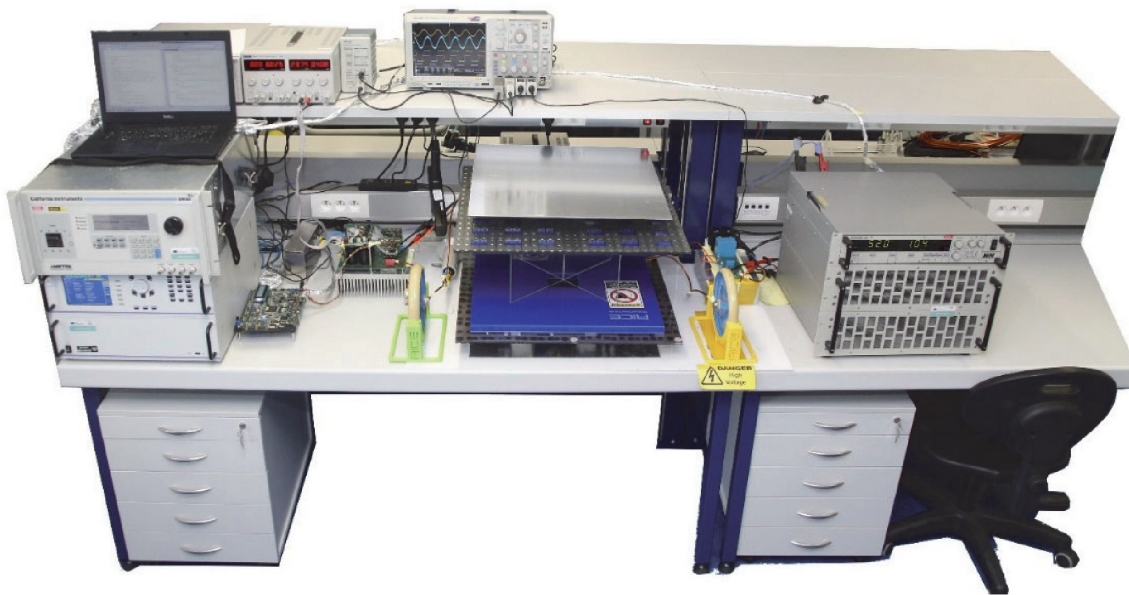


Figure 31.
Laboratory set-up for evaluation of the EM shielding impact.

Optimization to	U_1 [V]	P_1 [W]	P_2 [W]	η [%]	f_r [kHz]	f [kHz]	R_s [Ω]
Power	312	5116	4865	95.1	121.1	121	35
Efficiency	312	4056	3886	95.8	121.1	125	29
		R_1/R_2 [Ω]		L_1/L_2 [μ H]		C_1/C_2 [nF]	
		0.21/0.21		172/167		10.444	

Table 4.
Operational parameters of the system after application of the shielding.

Full-scale maps measured at reduced power (maximum efficiency) can be seen in **Figure 32**. The resonance is around 121 kHz, with the high efficiency range more than 10 kHz wide.

The results confirm the ability of the systems to deliver 4 kW to the load at an efficiency of >95%, which, apart from the higher supply frequency, places it in the

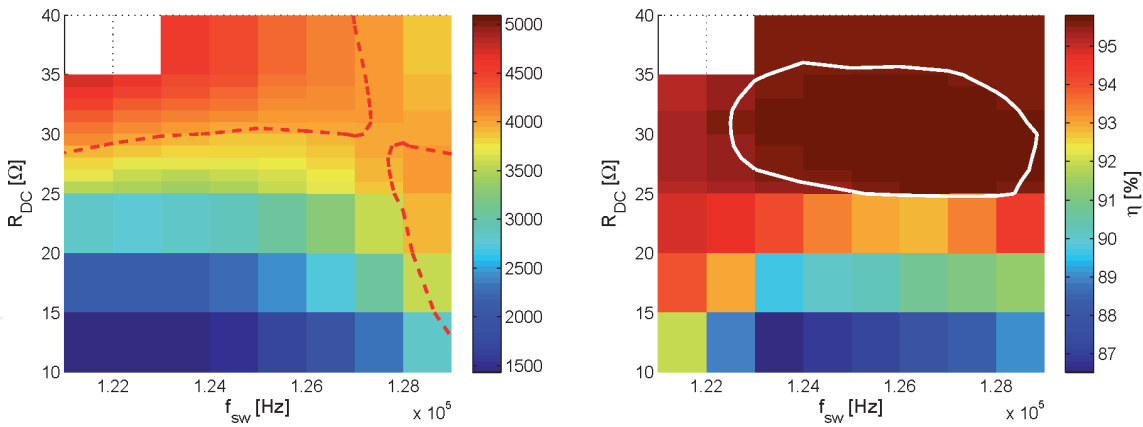


Figure 32.
Output power characteristic (left) and efficiency characteristic (right) for shielded system and 20 cm power transfer distance.

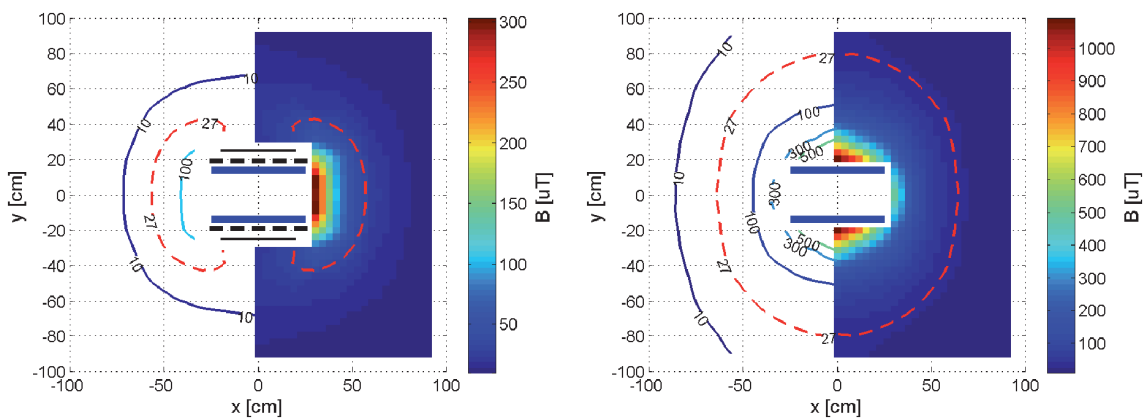


Figure 33.
Evaluation of the values of magnetic field around shielded (system) and non-shielded system (right) during experimental measurement at full power of proposed system.

“WPT 1” category according to the “SAE TIR J2954” wireless charging station standard.

To verify the shielding efficiency, a scattering magnetic field was also measured (measurement uncertainty <2%) around the coupling coils using a calibrated Narda ELT 400 probe. The values were recorded in the cutting plane with a regular step of 10 cm in length (**Figure 33**). The values of the magnetic induction relevant for hygienic limits are boarded by red dashed line (**Figure 33** left). It is seen, that specified limits are achieved approximately 20 cm from the top surface of the coils. Compared to unshielded system (**Figure 33** right), it is reduction of approximately 60 cm considering the spherical distance.

Based on the received and verified results it was achieved, that with the use of presented methodology, it is possible to design wireless charger, whose characteristics will meet standards and normative defined by regulatory companies.

5. Conclusions

The paper has given a brief recapitulation of most important standards and regulations relating to the high-power wireless charging systems. It has proposed the magnetic couplers to be designed exactly according to optimal operation to the specific load.

For medium or high-power wireless chargers, we have recommended to compose the system of input inductance, the active rectifier and the voltage source inverter, which can provide low THDi, excellent power factor and controllable output voltage. Thus, no additional dc/dc converters are needed.

The experimental prototype has proven the validity of presented physical principles and confirmed the proposed conceptual design strategies. It has also shown and discussed the comparison between ac-ac and dc-dc system efficiency relating to losses-to-power transfer ratio.

Additionally, the measurement of leakage magnetic field has shown the real flux density distribution observed around the circular-shaped coupling coils. This could be used for further optimization.

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Conflict of interest

The authors declare no conflict of interest.

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