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The Emission Level of Medium-Range Inductive Wireless Power Transmission Systems

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Abstract — **Medium-range inductive wireless power transmission systems allow a sufficient power transfer without requiring a close proximity between a primary coil and a secondary coil. We investigate the relationships between the power gain and the distance between the coils. We then address the radiated emission, from the perspectives of EMC and human exposure requirements.**

I. Introduction

Varying electric fields and magnetic fields can be used for wireless power transmission (WPT) from a power-transmitting unit to a power-receiving unit. Three WPT configurations may be defined:

- in the "electrically small electric antenna" case, the power-receiving unit uses a receiving antenna much smaller than the half wavelength(s) of the varying electromagnetic field produced by the power-transmitting unit, the receiving antenna being mainly responsive to the electric field;
- in the "electrically small magnetic antenna" case, the power-receiving unit uses a receiving antenna much smaller than the half wavelength(s) of the varying electromagnetic field produced by the power-transmitting unit, the receiving antenna being mainly responsive to the magnetic field;
- in the "electrically large antenna" case, the power-receiving unit uses a receiving antenna having a largest dimension commensurate with or larger than the half wavelength of the varying electromagnetic field produced by the power-transmitting unit.

The electrically small electric antenna configuration is very seldom used because of the safety issue relating to the high electric field strength needed for a significant power delivery; because low frequency electric fields are easily shielded by a layer of conductive dirt or moisture; and because the impedance of a suitable receiving antenna is too high for most applications. However, wireless power-feeding using a 50 Hz electric field is used in beacons for high-voltage overhead power lines, at line voltages higher than 60 kV. The receiving antenna is a rod having a typical length of a few meters.

The electrically small magnetic antenna configuration is popular for remote power-feeding applications because there is often no safety issue relating to the high magnetic field strength needed for an adequate power delivery; because low frequency magnetic fields are unaffected by non-conductive and non-magnetic items and not easily shielded by conductive non-magnetic items; and because the impedance of the receiving antenna is convenient for many applications.

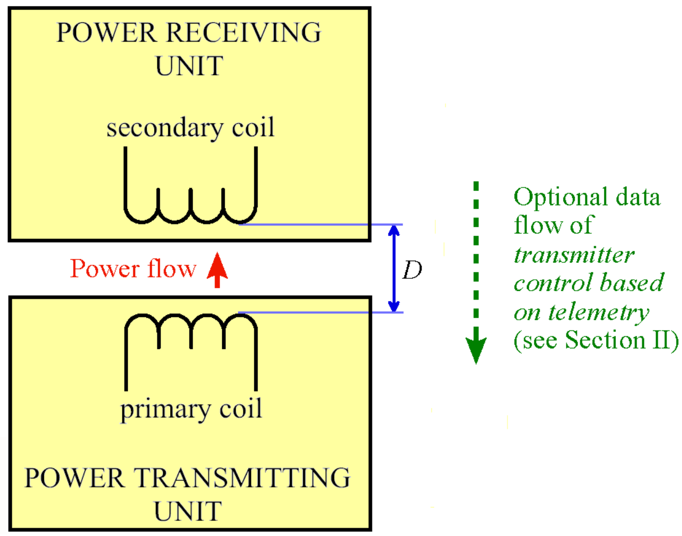

Fig. 1. An inductive WPT system.

The electrically large antenna configuration is more expensive to build, so that it is typically used where long-range power transmission is needed [1] [2].

WPT using an electrically small magnetic antenna is usually referred to as *inductive WPT*. Here, the antenna used in the power-transmitting unit is a primary coil and the receiving antenna is a secondary coil, as shown in Fig. 1. The primary coil and the secondary coil form a transformer having a primary and a secondary which are mechanically separable. When the two are placed in proper orientation and proximity, the coupling becomes sufficient to allow an adequate power transmission.

In a short-range inductive WPT system, power transmission takes place only when the power-transmitting unit and the power-receiving unit have a well-defined position with respect to each other and are in mechanical contact with each other. Thus, the primary and secondary coils are typically a few millimeters apart and a good efficiency can be obtained. Short-range WPT systems are for instance used in chargers for the rechargeable batteries of hand-held battery-operated items, such as electric tooth brushes, radio transceivers and cellular telephones [3]. This approach allows the hand-held item to be completely sealed, and uses no electrical connection between the mains and the hand-held item.

This paper is about medium-range inductive WPT systems, which do not require a close proximity between the primary and

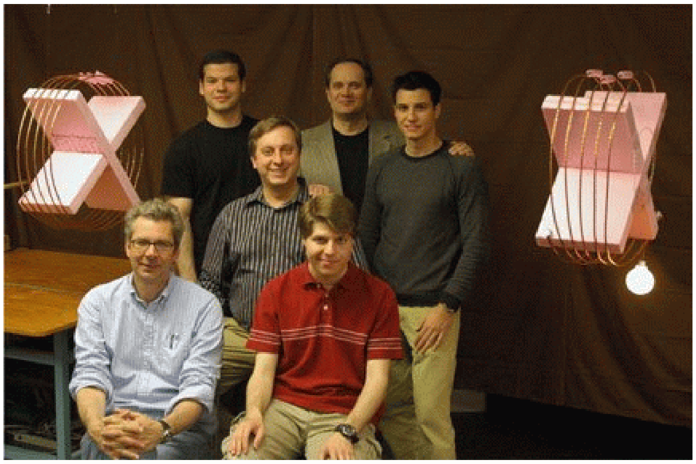

Fig. 2. The MIT experiment, showing power transmission over 2 m distance, and members of the team that performed the experiment. Courtesy of Aristeidis Karalis. Used with permission.

secondary coils during power transmission. The Fig. 2 shows an experimental medium-range inductive WPT system in operation [4] [5]. The added convenience of positioning-free operation entails an increased emission level, which can exceed EMC and human exposure requirements. Section II presents the typical characteristics of inductive WPT systems. Section III presents a computation of the power gain, which is a measure of the efficiency of a WPT system. Section IV addresses the radiated emission of medium-range WPT systems.

II. Presentation of Inductive WPT Systems

Let us use the wording *resonant coil* to designate a coil used in a series or parallel resonant circuit comprising a capacitor or a coil used at its self-resonance frequency. An inductive WPT system may use:

- non-resonant coils in the power-transmitting unit and in the power-receiving unit, in which case a reasonable efficiency requires that a magnetic circuit made of a magnetic material is used to obtain a strong coupling between the coils (i.e., a coefficient of coupling close to 1), the distance between the coils leaving only a gap much smaller than the section of the magnetic circuit;
- one resonant coil and one non-resonant coil, in which case a reasonable efficiency does not require a magnetic circuit made of a magnetic material, and the distance between the coils may be increased compared to two non-resonant coils;
- resonant coils in the power-transmitting unit and in the power-receiving unit, in which case a reasonable efficiency does not require a magnetic circuit made of a magnetic material, and the distance between the coils may be further increased compared to a device using only one resonant coil, up to a few times the largest dimension of one of the coils.

In some cases, the primary coil is driven directly by the a.c. voltage of the mains [3]. Typically, such a design uses an iron core in the primary coil and in the secondary coil, and both coils are non-resonant. Such a design may be considered as outdated and operates properly only when the coils are in close proximity.

In current inductive WPT systems, the primary coil is excited by a high-frequency current generated by an electronic circuit, typically a switched-mode or resonant inverter operating at a frequency above 20 kHz. Such a design typically implements resonant coils in the power-transmitting unit and/or in the power-receiving unit. For instance:

- many transponders for radio frequency identification (RFID) use inductive WPT to supply enough power to sustain the operation of the transponder, typically 10 μW to 1 mW [6];
- implantable medical device typically use inductive WPT with resonant coils in the frequency range 1 MHz to 20 MHz to deliver 5 mW to 250 mW to the implant [7] [8] [9];
- a Wireless Power Consortium has issued a specification for inductive WPT systems operating in the 110 kHz to 205 kHz frequency range and capable of delivering up to 5 W at a distance of about 5 mm between the primary and secondary coils [10] and some products and integrated circuits complying with this specification are commercially available [11] [12] [13];
- WPT systems intended to be used as battery chargers for electric vehicles are being designed, for instance a 30 kW system operating at a distance of 45 mm [14].

Inductive WPT systems using resonant coils in the power-transmitting unit and in the power-receiving unit provide the best efficiency for a given size of the coils and a given distance between the coils, but they require an accurate tuning. Additionally, power is wasted if the primary coil is excited when the power-receiving unit is not present or does not need to receive power. In order to avoid this situation and to compensate the effect of component tolerances, temperature changes, component drift and the effect of nearby conducting or magnetic items, it might be desirable to use a WPT system using a *self-tuning capability* of the power-transmitting unit or a more elaborate *transmitter control based on telemetry* involving communication from the power-receiving unit to the power-transmitting unit. According to the self-tuning capability, the power-transmitting unit is capable of:

- sensing the presence of a nearby secondary coil connected to a circuit comprising only a capacitor and a resistance, and varying or activating/deactivating the power in the primary coil accordingly; and/or
- dynamically seeking resonance and optimizing power transfer with a nearby secondary coil connected to a circuit comprising only a capacitor and a resistance.

Of course, the self-tuning capability must also be able to operate in the presence of the intended power-receiving unit, which typically comprises a rectifier and other non-linear circuits. We see that the self-tuning capability helps to obtain a good efficiency.

The above-mentioned specification of the Wireless Power Consortium uses *transmitter control based on telemetry*, involving a digital communication using load modulation, referred to as "backscatter modulation" in [10]. This specification also uses a form of *self-tuning capability*, referred to as "analog ping" in [10].

The distance between the primary coil and the secondary coil, denoted by D, could be used to set a boundary between short-range and medium-range inductive WPTs. However, this approach does not lead to general consequences from the design standpoint. In this paper, a medium-range inductive WPT is defined by a structural attribute: the primary coil does not comprise a magnetic circuit made of a magnetic material, as in the example of Fig. 2.

III. Power Transfer of Inductive WPT Systems

We will now assess the typical efficiency which might be expected from a WPT system using resonant coils in the power-transmitting unit and in the power-receiving unit. Since the largest dimension of each coil is small compared to wavelength, we can use a lumped-element circuit model to derive the currents through the coils. We will assume a simple equivalent linear circuit for the power-transmitting unit and the power-receiving unit. This is not a severe limitation since we only want to assess the behavior at the nominal frequency.

In line with Hochmair [7], we can use the model shown in Fig. 3 to investigate a WPT system using a parallel resonant circuit comprising the primary coil (in the power-transmitting unit) and a parallel resonant circuit comprising the secondary coil (in the power-receiving unit). In Fig. 3, L_1 is the primary coil, and L_2 is the secondary coil. Since the circuit of the secondary coil is floating, we have not introduced the capacitance between the windings in the model.

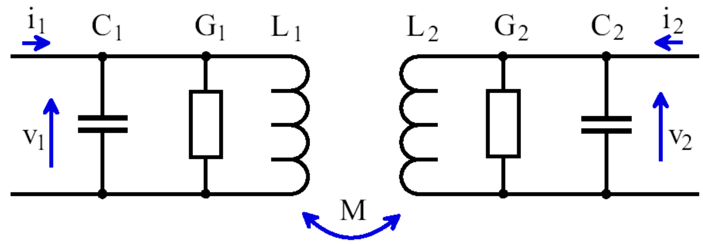

Fig. 3. Model for the case of parallel resonance.

The equations describing the operation of this circuit are

$$\begin{pmatrix} v_1 \\ v_2 \end{pmatrix} = \frac{1}{\Delta}\begin{pmatrix} j\omega L_{1EQ} & j\omega M \\ j\omega M & j\omega L_{2EQ} \end{pmatrix}\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} \tag{1}$$

where ω is the radian frequency and where

$$L_{1EQ} = L_1 - j\omega\left(M^2 - L_1 L_2\right)\left(j\omega C_2 + G_2\right) \tag{2}$$

$$L_{2EQ} = L_2 - j\omega\left(M^2 - L_1 L_2\right)\left(j\omega C_1 + G_1\right) \tag{3}$$

and

$$\begin{aligned}\Delta = 1 &+ j\omega L_1\left(j\omega C_1 + G_1\right) + j\omega L_2\left(j\omega C_2 + G_2\right) \\ &+ \omega^2\left(M^2 - L_1 L_2\right)\left(j\omega C_1 + G_1\right)\left(j\omega C_2 + G_2\right)\end{aligned} \tag{4}$$

We introduce the coefficient of coupling k, the quality factors Q_1 and Q_2 and the radian frequencies ω_{01} and ω_{02}, given by

$$k = \frac{M}{\sqrt{L_1 L_2}} \tag{5}$$

and

$$\begin{cases} Q_1 = \dfrac{1}{G_1}\sqrt{\dfrac{C_1}{L_1}} \\ \omega_{01} = \dfrac{1}{\sqrt{C_1 L_1}} \end{cases} \qquad \begin{cases} Q_2 = \dfrac{1}{G_2}\sqrt{\dfrac{C_2}{L_2}} \\ \omega_{02} = \dfrac{1}{\sqrt{C_2 L_2}} \end{cases} \tag{6}$$

We can replace the variables M, C_1, C_2, G_1 and G_2 in (1) to (4), with k, Q_1, Q_2, ω_{01} and ω_{02} defined by (5) and (6). We get:

$$L_{1EQ} = L_1\left(1 + \frac{\omega}{\omega_{02}}\left(1-k^2\right)\left(\frac{j}{Q_2} - \frac{\omega}{\omega_{02}}\right)\right) \tag{7}$$

$$L_{2EQ} = L_2\left(1 + \frac{\omega}{\omega_{01}}\left(1-k^2\right)\left(\frac{j}{Q_1} - \frac{\omega}{\omega_{01}}\right)\right) \tag{8}$$

$$M = k\sqrt{L_1 L_2} \tag{9}$$

and

$$\begin{aligned}\Delta = 1 &- \frac{\omega^2}{\omega_{01}^2} - \frac{\omega^2}{\omega_{02}^2} + j\left(\frac{\omega}{Q_1\omega_{01}} + \frac{\omega}{Q_2\omega_{02}}\right) \\ &-\left(1-k^2\right)\left(\frac{j\omega^2}{\omega_{01}^2} + \frac{\omega}{Q_1\omega_{01}}\right)\left(\frac{j\omega^2}{\omega_{02}^2} + \frac{\omega}{Q_2\omega_{02}}\right)\end{aligned} \tag{10}$$

At this stage, no simplifying assumption has been made. In practice, we can assume that the circuits are tuned to the same frequency ω_0, hence $\omega_0 = \omega_{01} = \omega_{02}$ and that the coupling is weak, that is for $k \ll 1$. In this case, an acceptable efficiency is only obtained for $Q_1 \gg 1$ and $Q_2 \gg 1$, so that we also make this assumption. At the resonance, that is for $\omega = \omega_0$, these assumptions lead us to

$$\begin{pmatrix} v_1 \\ v_2 \end{pmatrix} \approx \frac{1}{K^2+1}\begin{pmatrix} \dfrac{1}{G_1} & \dfrac{-jK}{\sqrt{G_1 G_2}} \\ \dfrac{-jK}{\sqrt{G_1 G_2}} & \dfrac{1}{G_2} \end{pmatrix}\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} \tag{11}$$

where, following Terman [15, section 3, § 5] and Hochmair [7], we have used the dimensionless coupling parameter K defined by

$$K = k\sqrt{Q_1 Q_2} \tag{12}$$

If we consider the problem for which i_2 is determined by a load of conductance G_{2L}, in such a way that $i_2 = -G_{2L}\, v_2$, we define the power gain as

$$a_P = \frac{\mathrm{Re}\left(v_2\, \bar{i}_2\right)}{\mathrm{Re}\left(v_1\, \bar{i}_1\right)} \tag{13}$$

where the bar denotes the complex conjugate.

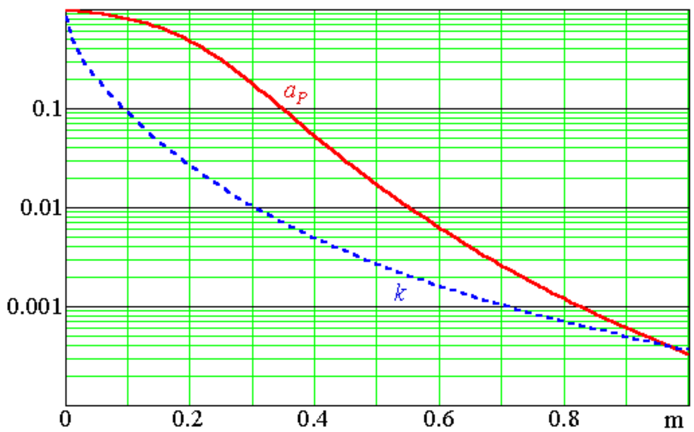

Fig. 4. coefficient of coupling k and power gain a_P versus D.

Using (11), we find that, at $\omega = \omega_0$,

$$a_P \approx \frac{K^2 G_2 G_{2L}}{\left(G_{2L} + G_2\left[K^2+1\right]\right)\left(G_{2L}+G_2\right)} \tag{14}$$

In order to maximize the power gain for G_2 and K fixed, we must use $G_{2L} = G_{2L\,MAX}$ with

$$G_{2L\,MAX} = G_2\sqrt{1+K^2} \tag{15}$$

for which the maximum gain satisfies

$$a_{P\,MAX} \approx \frac{K^2}{\left[1+\sqrt{1+K^2}\right]^2} \tag{16}$$

The Fig. 4 shows the coefficient of coupling k and the exact power gain a_P as a function of the distance D between the primary and secondary coils in a coaxial configuration, for $Q_1 = Q_2 = 100$ and parallel resonance at the frequency f_0 = 149 kHz. This computation considers 2 identical circular coils of mean radius r = 0.1 m each having a 36-turn winding of square cross-section, of side 3.6 mm, in free space. We assume that, at each D, the power-receiving unit absorbs a sinusoidal current and adjusts G_{2L} so that it takes on the value given by (15). The computation of k uses [16, p. 175, eq. 92] or [17, p. 358, Problem 5-29] which involve complete elliptic integrals, and [18, ch. 12, eq. 85]. In Fig. 4, the exact value of the power gain a_P obtained for $G_{2L} = G_{2L\,MAX}$ given by (15) is obtained using (1) to (4).

We can use the model shown in Fig. 5 to investigate a WPT system using a series resonant circuit comprising the primary coil L_1 and a series resonant circuit comprising the secondary coil L_2. Since the circuit of the secondary coil is floating, we have not introduced the capacitance between the windings in the model. This WPT system requires a separate analysis because, contrary to appearances, the circuits shown in Fig. 3 and Fig. 5 are not dually related, since each of these circuits comprises no dual element for the mutual inductances in the other circuit.

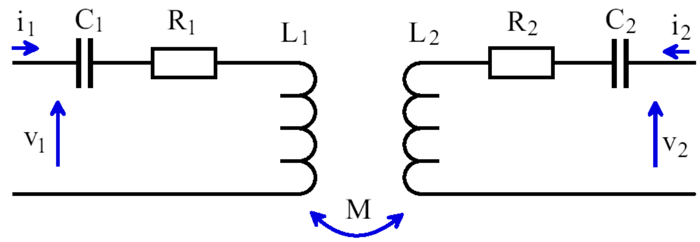

Fig. 5. Model for the case of series resonance.

The equations describing the operation of this circuit are

$$\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} = \frac{1}{\Delta}\begin{pmatrix} R_2 + \dfrac{1}{j\omega C_2} + j\omega L_2 & -j\omega M \\ -j\omega M & R_1 + \dfrac{1}{j\omega C_1} + j\omega L_1 \end{pmatrix}\begin{pmatrix} v_1 \\ v_2 \end{pmatrix} \tag{17}$$

where the new definition of Δ is

$$\begin{aligned}\Delta = &\left(j\omega L_1 + R_1\right)\left(j\omega L_2 + R_2\right) + \omega^2 M^2 \\ &+ \frac{1}{j\omega C_1}\left(j\omega L_2 + R_2\right) + \frac{1}{j\omega C_2}\left(j\omega L_1 + R_1\right) \\ &- \frac{1}{\omega^2 C_1 C_2}\end{aligned} \tag{18}$$

We introduce the quality factors Q_1 and Q_2 and the radian frequencies ω_{01} and ω_{02}, given by

$$\begin{cases} Q_1 = \dfrac{1}{R_1}\sqrt{\dfrac{L_1}{C_1}} \\ \omega_{01} = \dfrac{1}{\sqrt{C_1 L_1}} \end{cases} \qquad \begin{cases} Q_2 = \dfrac{1}{R_2}\sqrt{\dfrac{L_2}{C_2}} \\ \omega_{02} = \dfrac{1}{\sqrt{C_2 L_2}} \end{cases} \tag{19}$$

We now replace the five variables M, L_1, L_2, R_1 and R_2 in (17) and (18), with k, Q_1, Q_2, ω_{01} and ω_{02} defined by (5) and (19). We get:

$$\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} = \frac{1}{\Delta}\begin{pmatrix} \dfrac{1 + \dfrac{j\omega}{Q_2\omega_{02}} - \dfrac{\omega^2}{\omega_{02}^2}}{j\omega C_2} & \dfrac{-j\omega k}{\omega_{01}\omega_{02}\sqrt{C_1 C_2}} \\ \dfrac{-j\omega k}{\omega_{01}\omega_{02}\sqrt{C_1 C_2}} & \dfrac{1 + \dfrac{j\omega}{Q_1\omega_{01}} - \dfrac{\omega^2}{\omega_{01}^2}}{j\omega C_1} \end{pmatrix}\begin{pmatrix} v_1 \\ v_2 \end{pmatrix} \tag{20}$$

and

$$\Delta = \frac{-1}{\omega^2 C_1 C_2}\left[1 - \frac{\omega^2}{\omega_{01}^2} - \frac{\omega^2}{\omega_{02}^2} + j\left(\frac{\omega}{Q_1\omega_{01}} + \frac{\omega}{Q_2\omega_{02}}\right)\right.$$
$$\left. - \frac{\omega^2}{\omega_{01}\omega_{02}}\left(\frac{1}{Q_1 Q_2} + \frac{j\omega}{Q_1\omega_{02}} + \frac{j\omega}{Q_2\omega_{01}}\right) + \left(1-k^2\right)\frac{\omega^4}{\omega_{01}^4}\right] \tag{21}$$

At this stage, no simplifying assumption has been made. In practice, we can assume that the circuits are tuned to the same frequency ω_0, hence $\omega_0 = \omega_{01} = \omega_{02}$. At the resonance, that is for $\omega = \omega_0$, this assumption leads us to the exact result

$$\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} = \frac{1}{K^2+1}\begin{pmatrix} \dfrac{1}{R_1} & \dfrac{-jK}{\sqrt{R_1 R_2}} \\ \dfrac{-jK}{\sqrt{R_1 R_2}} & \dfrac{1}{R_2} \end{pmatrix}\begin{pmatrix} v_1 \\ v_2 \end{pmatrix} \tag{22}$$

where the dimensionless coupling parameter K is defined by

$$K = k\sqrt{Q_1 Q_2} \tag{23}$$

We note that (22) is the dual of (11), but (22) is exact whereas (11) is an approximate result. If we consider the problem for which i_2 is determined by a load of resistance R_{2L}, in such a way that $v_2 = -R_{2L}\, i_2$, we find that, at $\omega = \omega_0$, the power gain defined by (13) is given by

$$a_P = \frac{K^2\, R_2\, R_{2L}}{\left(R_{2L} + R_2\left[K^2+1\right]\right)\left(R_{2L} + R_2\right)} \tag{24}$$

In order to maximize the power gain for R_2 and K fixed, we must use $R_{2L} = R_{2L\,MAX}$ with

$$R_{2L\;MAX} = R_2\sqrt{1+K^2} \tag{25}$$

The maximum gain, obtained for $R_{2L} = R_{2L\,MAX}$, is

$$a_{P\,MAX} = \frac{K^2}{\left[1+\sqrt{1+K^2}\right]^2} \tag{26}$$

We can compute the exact power gain a_P as a function of the distance D between the primary and secondary coils, for the geometry and the coils defined above in the explanations for Fig. 4, using series resonance at the frequency f_0 = 149 kHz. Assuming that, at each D, the power-receiving unit absorbs a sinusoidal current and adjusts R_{2L} so that it takes on the value given by (25), the exact value of the power gain a_P obtained using (17) and (18) differs by less than 0.5 % from the results obtained for parallel resonance. Thus, Fig. 4 also applies to series resonance.

There are two other simple configurations which may be used to obtain resonant coils in the power-transmitting unit and in the power-receiving unit: parallel resonance in the power-transmitting unit with series resonance in the power-receiving unit and vice-versa [14]. The latter seems to be the most popular scheme. However, more complex structures can also be used, for instance involving two coupled coils in the power-transmitting unit and in the power-receiving unit [5], or a dual resonant circuit [10].

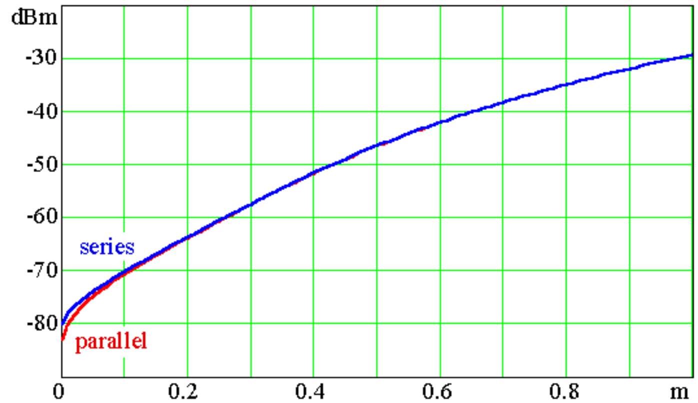

Fig. 6. Total radiated power $\mathscr{W}$ as a function of the distance D.

IV. Emission of Inductive Medium-Range WPT Systems

We now want to determine the radiated emission of the WPT system, to investigate electromagnetic compatibility and possible human exposure to the electromagnetic field. This emission is mainly characterized by the magnetic field intensity H, which will be computed in free space. We assume that the excitation of the resonant circuit comprising the primary coil is adjusted in such a way that a power of 1 W is delivered to the load of the power-receiving unit, in addition to the earlier assumption that the power-receiving unit absorbs the maximum power from the resonant circuit comprising the secondary coil.

If we measure H at a distance d of the primary coil, and if d is much larger than D and the largest dimension of the primary and secondary coils, the free space emission is close to the emission of a magnetic dipole, which is characterized by the total radiated power $\mathscr{W}$ given by [19, p. 438]

$$\mathscr{W} = \frac{\mathscr{k}^4 \eta_0}{6\pi}\left|\mathscr{M}\right|^2 \tag{27}$$

where $\mathscr{M}$ is the r.m.s. magnetic moment of the dipole, $\eta_0 \approx 376.7\ \Omega$ is the intrinsic impedance of free space and $\mathscr{k}$ is the wave number ω/c_0 where c_0 is the velocity of light. For the configuration used above, comprising two coaxial circular coils, we have

$$\left|\mathscr{M}\right| = \pi\, r^2\left(i_{L1} + i_{L2}\right) \tag{28}$$

where $i_{L\,1}$ and $i_{L\,2}$ are the r.m.s. currents flowing in the primary coil and in the secondary coil, respectively.

The Fig. 6 shows $\mathscr{W}$ computed as a function of D, in the case of the configurations shown in Fig. 3 and Fig. 5. In the case of parallel resonance shown in Fig. 3, $i_{L\,1}$ is the sum of the currents flowing through L_1 and G_1, and $i_{L\,2}$ is the sum of the currents flowing through L_2 and G_2. In the case of series resonance shown in Fig. 5, we have $i_{L\,1} = i_1$ and $i_{L\,2} = i_2$. Once $\mathscr{W}$ is computed, the

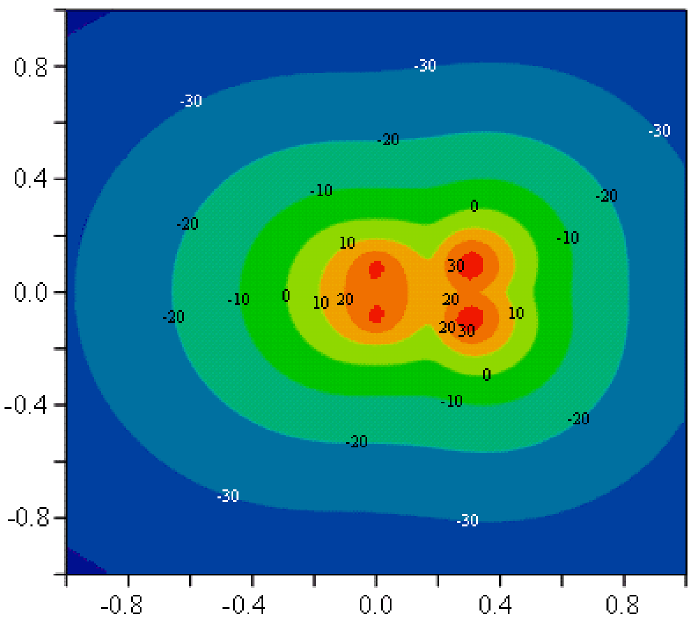

Fig. 7. Map of H in a plane containing the common axis of the coils, the origin being the center of the primary coil. Horizontal and vertical axis: cartesian coordinates in meters. Colors in 10 dB steps, 0 dB corresponding to about 134 dB(μA/m) or 5 A/m.

maximum r.m.s. magnetic field strength at the distance d, denoted by H_{MAX}, is given by [20]:

$$H_{MAX} = \frac{1}{d}\sqrt{\frac{3\mathscr{W}}{8\pi\eta_0}\frac{\mathscr{f}}{(\mathscr{k}d)^2}} \tag{29}$$

where

$$\mathscr{f} = \begin{cases} 2\sqrt{(\mathscr{k}d)^2+1} & \text{if } \mathscr{k}d \le (\mathscr{k}d)_C \\ \sqrt{(\mathscr{k}d)^4-(\mathscr{k}d)^2+1} & \text{if } \mathscr{k}d > (\mathscr{k}d)_C \end{cases} \tag{30}$$

and

$$(\mathscr{k}d)_C \equiv \sqrt{\frac{5+\sqrt{37}}{2}} \approx 2.354 \tag{31}$$

At a distance d of the primary coil which is not much larger than D and the largest dimension of the primary and secondary coils, we need to compute H at each point in space. The Fig. 7 shows a map of H, in the case of the configuration shown in Fig. 3, for coaxial coils and $D = 0.3$ m. Here, we have $D/2r = 1.5$ and $a_P = 0.174$. To obtain this map, we have used the analytical formula for the field intensity produced by a circular loop [16, p. 36, eq. 175] [19, p. 263, Problem 4]. The reference level relating to the limitation of exposure of the general public defined in [21] at the frequency $f_0 = 149$ kHz, namely 5 A/m, is 0 dB in Fig. 7.

V. Conclusion

For a medium-range inductive WPT system providing a reasonable efficiency, the ratio of the shortest distance between the coils to the largest dimension of the largest coil will always be less than 4,0 in an ideal experiment and less than 2,0 in a real-world application. From the safety standpoint, the results of Fig. 7 show that the reference levels relating to the limitation of exposure of the general public may be exceeded near the coils, for a transmitted power of only 1 W and a distance of 0.3 m between the coils. If we consider that the EMC directive is applicable to the WPT system, it should be regarded as a group 1 ISM [22]. However, where the above-defined transmitter control based on telemetry is used, the R&TTE directive is likely to be applicable.

References

[1] N. Shinohara, H. Matsumoto, "Experimental Study of Large Rectenna Array for Microwave Energy Transmission", *IEEE Trans. Microw. Theory Tech.*, vol. 46, No. 3, March 1998, pp. 261-268.

[2] J.O. McSpadden, Lu Fan, Kai Chang, "Design and Experiments of a High-Conversion-Efficiency 5.8 GHz Rectenna", *IEEE Trans. Microw. Theory Tech.*, vol. 46, No. 12, Dec. 1998, pp. 2053-2060.

[3] U.S. Patent number 5,680,028, "Charger for hand-held rechargeable electric apparatus with reduced magnetic field". Inventor: McEachern. Application filed: Sept. 21, 1994. Date of patent: Oct. 21, 1997.

[4] International patent application number PCT/US2006/026480 of 5 July 2006 (WO 2007/008646), "Wireless non-radiative energy transfer". Inventors: Joannopoulos *et al.*

[5] A. Kurs, A. Karalis, R. Moffatt, D.J. Joannopoulos, P. Fisher, M. Sojacic, "Wireless Power Transfer via Strongly Coupled Magnetic Resonnances", *Science*, Vol. 317, 6 July 2007, pp. 83-86.

[6] R. Want, "An introduction to RFID technology", *Pervasive Computing*, January-March 2006, pp. 25-33.

[7] E. Hochmair, "System Optimization for Improved Accuracy in Transcutaneous Signal and Power Transmission", *IEEE Trans. Biomed. Eng.*, vol. BME-31, No. 2, pp. 177-186, Feb. 1984.

[8] G.A. Kendir, W. Liu, G. Wang, M. Sivaprakasam, R. Bashirullah, M.S. Humayun, J.D. Weiland, "An Optimal Design Methodology for Inductive Power Link With Class-E Amplifier", *IEEE Trans. Circuits Syst. I: Regular Papers*, Vol. 52, No. 5, May 2005, pp. 857-866.

[9] M. Sawan, Y. Hu, J. Coulombe, "Wireless Smart Implants Dedicated to Multichannel Monitoring and Microstimulation", *IEEE Signal Processing Magazine*, vol. 5, No. 1, pp. 21-39, First quarter 2005.

[10] *System Description — Wireless Power Transfer — Volume I: Low Power — Part 1: Interface Definition*, version 1.0.3, Wireless Power Consortium, Sept. 2011.

[11] P. Corvisier, "Vers un standard unique de recharge sans fil des appareils nomades", *Electroniques*, No. 23, pp. 60-61, Jan. 2012.

[12] *bq500210 — Qi Compliant Wireless Power Transmitter Manager*, Texas Instruments, Aug. 2011.

[13] *bq51010, bq51011, bq51013 — Integrated Wireless Power Supply Receiver, Qi (Wireless Power Consortium) Compliant*, Texas Instruments, Aug. 2011.

[14] C.S. Wang, O.H. Stielau, G.A. Covic, "Design Considerations for a Contactless Electric Vehicle Battery Charger", *IEEE Trans. Ind. Applicat.*, Vol. 52, No. 5, Oct. 2005, pp. 1308-1314.

[15] F.E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, 1943.

[16] E. Durand, *Magnétostatique*, Masson, 1968.

[17] C.T.A. Johnk, *Engineering Electromagnetic Fields and Waves*, John Wiley & Sons, 1975.

[18] F.W. Grover, *Inductance Calculations*, Instrument Society of America, 1973.

[19] J.A. Stratton, *Electromagnetic Theory*, McGraw-Hill, 1941.

[20] F. Broydé, E. Clavelier, "Maximum electric and magnetic field strengths at a given distance from some ideal antennas", *IEEE Antennas & Propagation Magazine*, Vol. 50, No. 4, Aug. 2008, pp. 66-71, and "Correction", *IEEE Antennas & Propagation Magazine*, Vol. 50, No. 6, Dec. 2008, p. 38.

[21] Council Recommendation 1999/519/EC of 12 July 1999 on the limitation of exposure of the genral public to electromagnetic fields (0 Hz to 300 GHz).

[22] EN 55011:2009 + Amendment A1:2010. Industrial, scientific and medical (ISM) radio-frequency equipment — Radio disturbance characteristics — Limits and methods of measurement.