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Enhanced Wireless Power Transmission Using Strong Paramagnetic Response

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Abstract

A method of quasi-static magnetic resonant coupling has been presented for improving the power transmission efficiency (PTE) in near-field wireless power transmission, which improves upon the state of the art. The traditional source resonator on the transmitter side is equipped with an additional resonator with a resonance frequency that is tuned substantially higher than the magnetic field excitation frequency. This additional resonator enhances the magnetic dipole moment and the effective permeability of the power transmitter, owing to a phenomenon known as the strong paramagnetic response. Both theoretical calculations and experimental results show increased PTE due to amplification of the effective permeability. In measurements, the PTE was improved from 57.8% to 64.2% at the nominal distance of 15 cm when the effective permeability was 2.6. The power delivered to load was also improved significantly, with the same 10 V excitation voltage, from 0.38 to 5.26 W.

Index Terms

Magnetically coupled resonance; metamaterial; paramagnetic; power transfer efficiency (PTE); wireless power transfer

I. Introduction

Wireless transmission of electrical energy has received great attention in the last decade. Applications span over a broad range from submicrowatt radio-frequency identification transponders [1], milliwatt integrated circuit [2], and biomedical devices [3]–[5] to several watts for charging portable electronics [6]–[8] and a few kilowatts for charging batteries in electric vehicles [9]–[11].

Since the concept of nonradiative transmission was restated in [12], the resonant coupled quasi-static wireless power transmission has become mainstream among various methods of wireless power transmission. However, the basic operating principle has not been changed: 1) one resonator is placed on the transmitter and another on the receiver sides, as shown in Fig. 1(a); 2) the two resonators are tuned at the same resonance frequency, ω_0 , as the magnetic field excitation; and 3) a resonant coupling path is formed between the two

resonators. While impedance matching with the help of additional resonators on both sides of the wireless link increased the power transfer efficiency (PTE) compared with traditional resonators [12], the demand for achieving even higher PTE calls for employing more advanced principles.

It has recently been shown that using different resonance frequencies among coupled resonators can lead to higher PTE [13]–[15]. An additional resonator was introduced close to the original source or load resonators and the resonators were tuned at different resonance frequencies from the magnetic field excitation frequency. The resonance frequencies in this case were determined solely based on the frequency splitting of the two resonators within the transmitter or receiver. In other words, the criterion for selecting different resonance frequencies in previous works was merely to compensate for the interferences between adjacent resonators. The two resonators still shared the same resonance frequency, and the resonance frequency was not used as an active design parameter. Frequency adjustments in previous works, therefore, did not have the ability to modulate the transmitter–receiver coupled resonance toward an optimal performance, where PTE is maximized.

We discovered that the magnetic resonance of the source resonator can be enhanced even further by taking advantage of strong paramagnetic response of an additional resonator L_M , shown in Fig. 1(b), whose resonance frequency ω_M is substantially higher than the magnetic field excitation frequency. This will improve the PTE from transmitter to receiver.

II. Principles of Operation

The magnetic field normal to the added resonator is initially generated by the source resonator. Since the added resonator is in immediate vicinity of the source resonator on the transmitter side, a substantial portion of the magnetic flux is linked between the two resonators. The flux linkage induces a current in the added resonator, which in turn creates a magnetic dipole moment.

When the magnetic field excitation frequency, ω_0 , is substantially lower than the resonance frequency of the added resonator, ω_M , the dipole moment of the added resonator is in phase with that of the source resonator. Therefore, the real part of transmitter effective permeability, which consists of a combination of both resonators, becomes higher than unity. If the magnetic field excitation frequency is gradually increased, the real part of the effective permeability also increases. Simultaneously, the imaginary part of it, which is the loss of the added resonator, also increases. Therefore, there is an optimal frequency that maximizes the PTE. This optimal frequency is determined based on the Q-factors of the two resonators and the couplings from each resonator in the transmitter to the resonator in the receiver. If the magnetic field excitation frequency is further increased to be the same with the resonance frequency of the added resonator, the magnetic dipole moments of the two resonators will have 90° phase difference. At this frequency, the imaginary part of the effective permeability is quite large, although the real part of permeability is also high. If the excitation frequency is increased above the resonance frequency of added resonator, the magnetic dipole moment of the added resonator turns out of phase with respect to the source resonator, resulting in reduction in the real part of the transmitter effective permeability.

Note that such magnetic responses that result from tuning the resonance frequency of the added resonator higher or lower than the excitation frequency are analogous to the paramagnetic and diamagnetic phenomena, respectively. In this light, the additional resonator can be viewed as a single-cell magnetic metamaterial that is capable of manipulating the effective permeability of the transmitter. The conception and formulation of the original split-ring magnetic metamaterial were also based on the single-cell approach [16]. The operating region of the proposed wireless power transmission link is in the in-phase frequency region, i.e., in the paramagnetic region.

Using the circuit model of magnetic dipole coupling [17], [18], the effective permeability of the transmitter can be found as

$$\begin{aligned}\mu_{\text{eff}}(\omega) &= 1 + \frac{k_{\text{SRC},M}^2}{\omega_M^2 - 1 + \frac{j}{Q_M}} \\ &= \mu_{\text{real}}(\omega) - j\mu_{\text{imag}}(\omega)\end{aligned}\quad (1)$$

where $j = \sqrt{-1}$, $k_{\text{SRC},M}$ is the magnetic flux coupling ratio between the source resonator and the added resonator or metamaterial, ω_M is the resonance frequency of the added resonator, Q_M is the quality factor of the additional resonator, and μ_{real} and μ_{imag} are the real and imaginary parts of the effective permeability as functions of the excitation frequency, ω , respectively. $k_{\text{SRC},M}$ is analogous to the fractional volume F in [16]. The sign of the imaginary part in (1) is negative due to the time convention of $\exp(j\omega t)$. A similar approach, using a lumped circuit model, can be found in [19]. Circuit simulations using the advanced design system by Agilent Technologies (Santa Clara, CA) in Fig. 2 show that at 13.56 MHz excitation frequency, the effective permeability has been multiplied by a factor of 2.6.

A key advantage of the proposed scheme over conventional metamaterials is that the additional resonator does not occupy the free space between transmitter and receiver coils. The additional resonator can be embedded into the transmitter housing or placed coplanar with respect to the source resonator. On the other hand, the metamaterial slabs in most of the previous literatures were placed within the free space between transmitter and receiver [18], [20], [21], compromising the true wireless power transmission distance.

Fig. 3 shows simulated magnetic field distribution at different relative permeability values. It can be seen that the magnetic field around the transmitter and its resonant coupling with the receiver have been increased with higher effective permeability without increasing the excitation amplitude. Hence, this can be considered a method to control the magnetic field amplification. However, increasing the permeability too much will increase the power loss in the added resonator, as can be observed in the imaginary part of μ_{eff} in Fig. 2. There exists an optimal value for the effective permeability, and therefore for ω_M , which maximizes the PTE.

The resonance frequency of the source resonator, ω_{SRC} , should be determined by the effective permeability of the transmitter. The effective resonance frequency of entire transmitter is lowered by a factor of $\sqrt{\mu_{\text{real}}}$ as a result of the added resonator because the

effective source inductance is increased by a factor of μ_{real} . Therefore, the required resonance frequency of source resonator is

$$\omega_{\text{SRC}} = \omega_0 \sqrt{\mu_{\text{real}}} \quad (2)$$

where ω_0 is the magnetic field excitation frequency.

This is in contrast to the conventional paradigm where all the resonators in the wireless link are tuned to resonate at the same frequency as the magnetic field excitation. By manipulating the resonance frequency of the added resonator, ω_M , the degree of μ_{eff} enhancement can be controlled to maximize the PTE under various operating conditions. Simultaneously, resonance frequency of the source resonator is also adjusted such that the effective resonance frequency of the transmitter is tuned to ω_0 .

III. PTE Improvement

Due to amplification of the effective permeability, the near-field reactive energy stored within the transmitter is increased, which in turn enhances the resonant coupling with the receiver. The PTE can be found from

$$\text{PTE} = \frac{\frac{R_{L,\text{refl}}}{R_{\text{RX}} + R_{L,\text{refl}}}}{\left\{ \frac{1}{k_{\text{SRC,RX}}^2 Q_{\text{SRC}} Q_{\text{RX,L}}} \left| \frac{1}{1+\text{AF}} \right|^2 + \frac{1}{k_{\text{M,RX}}^2 Q_{\text{M}} Q_{\text{RX,L}}} \left| \frac{\text{AF}}{1+\text{AF}} \right|^2 + 1 \right\}} \quad (3)$$

under these conditions

$$\left(\frac{\omega_M^2}{\omega_0^2} - 1 \right)^2 \gg \frac{1}{Q_M^2}, \quad \left(\frac{\omega_M^2}{\omega_0^2} - 1 \right)^2 \gg (k_{\text{M,RX}}^2 Q_{\text{RX,L}})^2$$

and

$$k_{\text{SRC,M}}^2 \gg (k_{\text{SRC,RX}} k_{\text{M,RX}} Q_{\text{RX,L}})^2$$

where

$$\text{AF} = \frac{k_{\text{M,RX}}}{k_{\text{SRC,RX}}} \frac{\mu_{\text{real}}(\omega_0) - 1}{k_{\text{SRC,M}}}$$

Q_{SRC} and Q_{M} are the quality factors of the source and added resonators, respectively, $Q_{\text{RX,L}} = \omega_0 L_{\text{RX}} / (R_{\text{RX}} + R_{L,\text{refl}})$, and $R_{L,\text{refl}}$ is the reflected resistance from the load coil onto the receiver resonator. Please refer to appendix for the detailed derivations. The first term in the

denominator is the energy loss in source resonator relative to the transferred power, and the second term is the loss in the added resonator relative to the transferred power. As μ_{real} is increased, loss in the source resonator is reduced while loss in the added resonator is increased. This is also apparent from Fig. 2 by noting that μ_{imag} is significantly increased if μ_{real} is to be increased at frequencies below ω_M . Therefore, there exists an optimal value of μ_{real} , which maximizes the PTE, as

$$\mu_{\text{real}}(\omega_0) = 1 + k_{\text{SRC},M} \frac{k_{M,\text{RX}} Q_M}{k_{\text{SRC},\text{RX}} Q_{\text{SRC}}}. \quad (4)$$

From the optimal condition in (4) and the definition of μ_{real} , the optimal resonance frequency of the added resonator is

$$\omega_M = \omega_0 \sqrt{1 + k_{\text{SRC},M} \frac{k_{\text{SRC},\text{RX}} Q_{\text{SRC}}}{k_{M,\text{RX}} Q_M}}. \quad (5)$$

As a conservative approximation, the source and the added resonators can be assumed to have nearly the same magnetic coupling coefficient to the receiver resonator and identical Q factors. From the generalized PTE equation in (3) and the optimal condition in (4), the simplified PTE can be obtained as

$$\text{PTE} \cong \frac{1}{\frac{1}{2k_{\text{SRC},\text{RX}}^2 Q_{\text{SRC}} Q_{\text{RX}}} + 1} \frac{R_{L,\text{refl}}}{R_{\text{RX}} + R_{L,\text{refl}}}. \quad (6)$$

Note that the Q -factor of the added resonator is typically higher than that of the source resonator because it is not connected to other components, such as output resistance of the power amplifier (PA). Moreover, $k_{M,\text{RX}}$ is slightly larger than $k_{\text{SRC},\text{RX}}$ due to being on the outer side. Hence, the actual PTE is typically higher than the conservative estimate in (6).

Comparing (6) with the PTE of the conventional scheme in Fig. 1(a) from [17]

$$\text{PTE}_{\text{conventional}} \cong \frac{1}{\frac{1}{k_{\text{SRC},\text{RX}}^2 Q_{\text{SRC}} Q_{\text{RX}}} + 1} \frac{R_{L,\text{refl}}}{R_{\text{RX}} + R_{L,\text{refl}}} \quad (7)$$

it is clear that the PTE has been increased.

IV. PA Output Power Versus Permeability

By amplifying the effective permeability, the reflected resistance from receiver to L_{SRC} is found as

$$Z_{\text{reflected,RX}} \cong k_{\text{SRC,RX}}^2 \omega_0 L_{\text{SRC}} Q_{\text{RX,L}} \times \left(1 + \frac{k_{\text{M,RX}}}{k_{\text{SRC,RX}}} \frac{\mu_{\text{real}} - 1}{k_{\text{SRC,M}}} \right)^2. \quad (8)$$

In addition to (8), the loss from the added resonator L_M is also reflected onto L_{SRC}

$$Z_{\text{reflected,LM}} \cong \frac{\omega_0 L_{\text{SRC}}}{Q_M} \left(\frac{\mu_{\text{real}} - 1}{k_{\text{SRC,M}}} \right)^2. \quad (9)$$

The total reflected resistance observed by L_{SRC} is the sum of (8) and (9). Comparing (8) with those of two- or three-coil transmitter in [23], the reflected resistance is increased by squared parenthesis. While this term improves the PTE, it may lower the PA output power, depending on the PA configuration, if the excitation voltage does not change. For example, the output power of series-resonant Class-D or Class-E PAs is inversely proportional to the load resistance. This may reduce the power delivered to load (PDL). The PA output power can be easily increased by increasing the PA supply voltage as long as the PA components can handle it.

V. Experimental Results

Fig. 4 shows the measurement setups. Solid magnet wires made of copper with 1.5 [15 American wire gauge (AWG)] and 0.5 mm (24 AWG) diameters were used to fabricate the transmitter and receiver coils, respectively. Low-loss resonating ceramic capacitors were added to turn them into resonators. For the receiver resonator, a conventional impedance matching loop, L_L , was used to maximize the loaded-Q of the receiver. The outer diameter of the receiver coil and its nominal distance from the transmitter were chosen to be 8 and 15 cm, respectively, based on the size of a smartphone to be charged when it is randomly placed on a desk with an embedded power transmitter. The load resistance of the receiver was considered as 50 Ω , and target operating frequency was 13.56 MHz, within the industrial–scientific–medical (ISM) band. Under these conditions, the optimal geometry of the conventional transmitter was calculated using the method in [22] and [23]. It was found that 19 cm diameter of transmitter yields maximum PTE. The conventional transmitter was then substituted with the proposed permeability-amplified transmitter under the same conditions to compare the two links. In measurements, to compensate for the additional loss in L_{SRC} , the theoretical optimal μ_{real} of two was increased to 2.6 to achieve the highest PTE. Table I provides a summary of the inductive link specifications.

Although there seems to be two resonators in the conventional four-coil transmitter of [12] and [23], and Fig. 4(a), the driver coil L_D is not resonantly coupled to the receiver coil L_{RX} . Rather, the function of the driver coil is limited to excitation of the source resonator, L_{SRC} , with proper impedance transformation [15], [17]. The actual energy transfer between transmitter and receiver is only through the source resonator, L_{SRC} [15], [17]. Therefore, the conventional four-coil link is single resonant, and the dimension and position of L_D have a little effect on the power exchange from transmitter to receiver.

Instead, the dimension and the position of the driver coil, L_D , affect the transmitter internal efficiency, which is defined by the delivered power to L_{SRC} divided by the total excitation power. In the experiment setup of Fig. 4(a), the dimension of L_D and the $k_{D,SRC}$ were chosen large enough to saturate the transmitter internal efficiency [23]. In other words, the PTE of four-coil configuration in the experiment was maximized with the selected geometry of L_D .

Unlike the conventional four-coil link, the proposed permeability-amplified transmitter is excited directly by the source resonator. Both the source and the added resonators are resonantly coupled to the receiver, while the added resonator enhances the resonance coupling.

Wireless power transmission resonance links, either conventional or the proposed one, can be abstracted as linear two-port networks. Unlike semiconductor devices, the efficiency of these resonators is not affected by the power level as long as the core materials (air in this case) are not saturated. Under these conditions, the VNA-based method, described and used here, provides an accurate method of measurement. The Vector Network Analyzer (VNA) excites port 1 at a known voltage, and senses the received signal across port 2, and vice versa. It measures the ratio between the input and output voltages. The measurement results are used to extract S-parameters. The S-parameters can be converted to Z-parameters, and finally, the PTE can be found from the equation in [24].

Fig. 5(a) compares the PTEs with and without the proposed utilization of the strong paramagnetic response. It can be seen that the paramagnetic response in the proposed link has resulted in 11% improvement in the PTE at 15 cm coil separation. Fig. 5(b) shows the PDL at 10 V PA supply voltage. It can be seen that PDL with the conventional four-coil link around the nominal distance of 15 cm is very small. This is mainly due to the excessive reflected resistance resulting from the impedance transformation of L_{SRC} and L_D . Although reducing $k_{D,SRC}$ will increase the PDL, it will slightly degrade the PTE.

VI. Conclusion

We presented a new method for coupled resonance enhancement using strong paramagnetic response of transmitter resonator. The coupled resonance is increased due to the paramagnetic response once an added resonator, which resonance frequency is higher than magnetic field excitation frequency, is introduced. This enhances the PTE compared with the traditional single resonant system, whose resonators are tuned at the excitation frequency.

One advantage of the paramagnetic effect that results from the added resonator is its ability to control the effective permeability to compensate for various operating conditions. For example, in many wireless power applications, the PA that drives the source resonator causes additional losses. Moreover, each resonator within the transmitter has a different magnetic coupling coefficient to the receiver resonators. Under these conditions the PTE can be maximized by simply adjusting the resonance frequency and, therefore the permeability of transmitter.

Although the conventional four-coil link improves the PTE by inserting an impedance transformation coil, L_D , it suffers from very low PDL at large distances due to high $k_{D, \text{SRC}}$, which reflects an excessively large resistance onto the PA output. This reduces the PA output power at a constant excitation voltage, and therefore the PDL. Small PDL can be avoided by reducing $k_{D, \text{SRC}}$ at the cost of PTE degradation, because in that case, more of the output power will be dissipated in L_D instead of being delivered to L_{SRC} . Therefore, as pointed out in [23], the conventional four-coil link is not the optimal choice when both high PTE and PDL are required.

On the other hand, in the proposed method, permeability amplification is able to increase the PTE without sacrificing PDL when compared with the conventional four-coil link. It should be pointed out that the reflected resistance is increased by a factor of the squared parenthesis in (8) when compared with the two- and three-coil links, and therefore the PA output power would be lower with the same excitation voltage. Nevertheless, the PTE of the proposed method is higher than that of the two- and three-coil links. Therefore, to draw a clearer picture as to where the proposed method should be used most effectively to strike a balance between PTE and PDL, one should use the figure-of-merit introduced in [25].

The presented method can also be used in magnetic resonance imaging (MRI). The field strength and the sensitivity of MRI have been increased using metamaterial slabs near the magnetic field source [26], [27]. The proposed additional resonator is capable of reinforcing the MRI field strength by amplifying the source permeability, which can be controlled according to (1). Aside from wireless power transmission and MRI applications, the concept of adding a paramagnetic resonator near the magnetic field source is expected to be applicable where high-intensity resonant magnetic fields are needed.

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Appendix

A. Derivation of the Effective Permeability in (1)

The effective permeability, μ_{eff} , in this paper is defined as the increase in the source inductance in the presence of strong coupling from an additional resonator. This is based on the fact that the value of any lumped inductor is proportional to its permeability. By solving the Kirchhoff's voltage law (KVL) equations for the lumped circuit model in Fig. 6, the input impedance Z_{in} can be found from

$$Z_{\text{in}} = j\omega L_{\text{SRC}} \left(1 + \frac{k_{\text{SRC},M}^2}{\frac{\omega_M^2}{\omega^2} - 1 + \frac{j}{Q_M}} \right) \quad (\text{A1})$$

where $\omega_M = 1/\sqrt{L_M C_M}$ and $Q_M = \omega L_M / R_M$. The input impedance due to the source resonator is $j\omega L_{\text{SRC}}$. However, the impedance is multiplied by the factor in parentheses by introduction of the additional resonator. Therefore, we can consider the term in parenthesis as the effective permeability of source inductor with the added resonator.

B. Derivation of PTE in (3)

The equivalent circuit schematic of Fig. 1(b) is shown in Fig. 7. At resonance, the effect of the load coil on the receiver resonator is a reflected resistance, $R_{L,\text{refl}} = k_{\text{RX},L}^2 \omega_0^2 L_{\text{RX}} L_L / R_L$, as shown in Fig. 8.

The PTE can be defined as [12]

$$\begin{aligned} \text{PTE} &= \frac{(R_{L,\text{refl}} + R_{\text{RX}}) |I_{\text{RX}}|^2}{R_{\text{SRC}} |I_{\text{SRC}}|^2 + R_M |I_M|^2 + (R_{L,\text{refl}} + R_{\text{RX}}) |I_{\text{RX}}|^2} \times \frac{R_{L,\text{refl}}}{(R_{\text{RX}} + R_{L,\text{refl}})} \\ &= \frac{\frac{R_{\text{SRC}}}{R_{L,\text{refl}} + R_{\text{RX}}} \left| \frac{I_{\text{SRC}}}{I_{\text{RX}}} \right|^2 + \frac{R_M}{R_{L,\text{refl}} + R_{\text{RX}}} \left| \frac{I_M}{I_{\text{RX}}} \right|^2 + 1}{1} \times \frac{R_{L,\text{refl}}}{(R_{\text{RX}} + R_{L,\text{refl}})} \end{aligned} \quad (\text{A2})$$

where I_{SRC} , I_M , and I_{RX} are the currents in L_{SRC} , L_M , and L_{RX} coils, respectively. R_{SRC} , R_M , and R_{RX} are the parasitic resistances of L_{SRC} , L_M , and L_{RX} , respectively. The KVL equations can be written as

$$\left(\frac{R_{\text{SRC}}}{j\omega L_{\text{SRC}}} + 1 - \frac{\omega_{\text{SRC}}^2}{\omega^2} \right) I_{\text{SRC}} + k_{\text{SRC},M} \sqrt{\frac{L_M}{L_{\text{SRC}}}} I_M + k_{\text{SRC,RX}} \sqrt{\frac{L_{\text{RX}}}{L_{\text{SRC}}}} I_{\text{RX}} = \frac{V_s}{j\omega L_{\text{SRC}}} \quad (\text{A3})$$

$$k_{\text{SRC},M} \sqrt{\frac{L_{\text{SRC}}}{L_M}} I_{\text{SRC}} + \left(\frac{R_M}{j\omega L_M} + 1 - \frac{\omega_M^2}{\omega^2} \right) I_M + k_{M,\text{RX}} \sqrt{\frac{L_{\text{RX}}}{L_M}} I_{\text{RX}} = 0 \quad (\text{A4})$$

$$k_{\text{SRC,RX}} \sqrt{\frac{L_{\text{SRC}}}{L_{\text{RX}}}} \mathbf{I}_{\text{SRC}} + k_{\text{M,RX}} \sqrt{\frac{L_{\text{M}}}{L_{\text{RX}}}} \mathbf{I}_{\text{M}} + \left(\frac{R_{\text{L,refl}} + R_{\text{RX}}}{j\omega L_{\text{RX}}} + 1 - \frac{\omega_0^2}{\omega^2} \right) \mathbf{I}_{\text{RX}} = 0 \quad (\text{A5})$$

where $\omega_{\text{SRC}} = 1/\sqrt{L_{\text{SRC}} C_{\text{SRC}}}$, $\omega_{\text{M}} = 1/\sqrt{L_{\text{M}} C_{\text{M}}}$, and $\omega_0 = 1/\sqrt{L_{\text{RX}} C_{\text{RX}}}$. The current ratios can be found from

$$\left| \frac{\mathbf{I}_{\text{SRC}}}{\mathbf{I}_{\text{RX}}} \right| = \frac{(R_{\text{RX}} + R_{\text{L,refl}})}{k_{\text{SRC,RX}} \omega_0 \sqrt{L_{\text{SRC}} L_{\text{RX}}}} \frac{1}{1 + \frac{k_{\text{M,RX}} \mu_{\text{real}}(\omega_0) - 1}{k_{\text{SRC,RX}} k_{\text{SRC,M}}}} \quad (\text{A6})$$

$$\left| \frac{\mathbf{I}_{\text{M}}}{\mathbf{I}_{\text{RX}}} \right| = \frac{(R_{\text{RX}} + R_{\text{L,refl}})}{k_{\text{M,RX}} \omega_0 \sqrt{L_{\text{M}} L_{\text{RX}}}} \frac{\frac{k_{\text{M,RX}} \mu_{\text{real}}(\omega_0) - 1}{k_{\text{SRC,RX}} k_{\text{SRC,M}}}}{1 + \frac{k_{\text{M,RX}} \mu_{\text{real}}(\omega_0) - 1}{k_{\text{SRC,RX}} k_{\text{SRC,M}}}} \quad (\text{A7})$$

under these conditions

$$\left(\frac{\omega_{\text{M}}^2}{\omega_0^2} - 1 \right)^2 \gg \frac{1}{Q_{\text{M}}^2} \quad \left(\frac{\omega_{\text{M}}^2}{\omega_0^2} - 1 \right)^2 \gg \left(k_{\text{M,RX}}^2 \frac{\omega_0 L_{\text{RX}}}{(R_{\text{RX}} + R_{\text{L,refl}})} \right)^2$$

and $k_{\text{SRC,M}}^2 \gg (k_{\text{SRC,RX}} k_{\text{M,RX}} (\omega_0 L_{\text{RX}}) / (R_{\text{RX}} + R_{\text{L,refl}}))^2$, where $Q_{\text{M}} = \omega_0 L_{\text{M}} / R_{\text{M}}$ is the quality factor of L_{M} . Substituting (A6) and (A7) into (A2) finally yields the PTE as in (3).

Fig. 1.

(a) Conventional four-coil wireless power transmission link. Although there are four coils in the link, L_{SRC} and L_{RX} coils are the resonators that are responsible for the resonant coupling between transmitter and receiver. L_D and L_L coils only perform impedance transformation, and do not contribute to the resonant coupling. The resonance frequency of all the resonators is tuned at the excitation frequency, ω_0 . (b) Proposed wireless power transmission link. The transmitter consists of a source resonator, L_{SRC} , and an additional resonator, L_M . The resonance frequencies of the source, ω_{SRC} , and the additional resonators, ω_M , are different from ω_0 . The magnetic dipole $\mathbf{m}$ can be increased by adjusting ω_M and subsequently μ_{real} .

Fig. 2.

Simulated effective permeability and PTE versus excitation frequency. (Inset: log-scale magnified vertical window between 11 and 15.5 MHz). At the excitation frequency of 13.56 MHz, the real part of the effective permeability was 2.6 when the resonance frequency of the added resonator was set to 15.4 MHz. It is possible to maximize the PTE at the desired excitation frequency by adjusting the resonance frequency of the added resonator, ω_M . The optimal ω_M is determined based on the Q-factors of the two resonators and the couplings from each resonator in the transmitter to the resonator in the receiver.

Fig. 3.

Simulated magnetic field strength using the high-frequency structure simulator by ANSYS (Canonsburg, PA). The effective permeability can be controlled by adjusting the resonance frequency of the added resonator, ω_M . Higher effective permeability yields higher resonant coupling at the same excitation amplitude. (a) $\mu_{\text{real}} = 1.7$. (b) $\mu_{\text{real}} = 2.6$.

Fig. 4.

Measurement setup. (a) Conventional four-coil wireless power link, similar to [12]. The transmitter in this configuration was coupled to a driver coil, L_D . The driver coil does not contribute to the resonant coupling with the receiver and interacts only with the source resonator. The driver coil was connected to the VNA excitation port. The source resonator was tuned at 13.56 MHz and unloaded. (b) Proposed permeability enhanced wireless power link. The transmitter consists of the source (L_{SRC}) and added (L_M) resonators. These two resonators were mounted in the same plane, in a coplanar fashion. The source resonator was connected to the excitation port of the VNA. The added resonator was tuned at 15.4 MHz and unloaded.

Fig. 5.

(a) PTE as a function of the distance between transmitter and receiver. The proposed system has improved the PTE by 11% at 15 cm distance. (b) PDL at 10 V Tx supply voltage.

Fig. 6.
Equivalent circuit model used to calculate the effective permeability.

Fig. 7.
Equivalent circuit model for proposed system.

Fig. 8.

Effect of load coil in Fig. 7 can be reflected to the receiver coil as $R_{L,refl}$.

TABLE I

Conventional and Paramagnetic Inductive Link Specifications

Parameters	Symbols	Conventional 4-coil	Proposed μ -amplified
L_D/L_{SRC}	Inductance	L_D/L_{SRC}	0.31 / 1.97 μ H
	Outer diameter	D_o	12/18.7 cm
	Turn number	n	1/2
	Quality factor	Q_D/Q_{SRC}	240/307
	Wire diameter	W_D/W_{SRC}	1.5 mm
L_M	Inductance	L_M	-
	Outer diameter	D_o	-
	Turn number	n	-
	Quality factor	Q_M	-
	Wire diameter	W_M	-
L_{RX}/L_L	Inductance	L_{RX}/L_L	1.52/ 0.344 μ H
	Outer diameter	D_o	8 / 4.2 cm
	Turn number	n	3/2
	Quality factor	Q_{RX}/Q_L	185/145
	Wire diameter	W_{RX}/W_L	0.5 mm
Tx - Rx coupling distance		d	15 cm
Power transfer efficiency	Calculated		56.6 %
	Simulated	PTE	60.1 %
	Measured		57.8 %
Source-to-receiver coupling ratio		$k_{SRC,RX}$	0.0153
Added-to-receiver coupling ratio		$k_{M,RX}$	-
Source-to-added coupling ratio		$k_{SRC,M}$	-
Receiver-to-load coupling ratio		$k_{RX,L}$	0.2
Driver-to-source coupling ratio		$k_{D,SRC}$	0.21
Added resonator resonance frequency		f_M	-
Source resonator resonance frequency		f_{SRC}	13.56 MHz
Real part of effective permeability		μ_{real}	-
Magnetic field excitation frequency		f_0	13.56 MHz
Load resistance		R_L	50 Ω