



- (51) **International Patent Classification:**
H02J 17/00 (2006.01) **H04B 5/02** (2006.01)
H02J 7/00 (2006.01)
- (21) **International Application Number:**
PCT/US2011/038635
- (22) **International Filing Date:**
31 May 2011 (31.05.2011)
- (25) **Filing Language:** English
- (26) **Publication Language:** English
- (30) **Priority Data:**
61/491,844 31 May 2011 (31.05.2011) US
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- (81) **Designated States (unless otherwise indicated, for every kind of national protection available):** AE, AG, AL, AM, AO, AT, AU, AZ, BA, BB, BG, BH, BR, BW, BY, BZ, CA, CH, CL, CN, CO, CR, CU, CZ, DE, DK, DM, DO, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, GT, HN, HR, HU, ID, IL, IN, IS, JP, KE, KG, KM, KN, KP, KR, KZ, LA, LC, LK, LR, LS, LT, LU, LY, MA, MD, ME, MG, MK, MN, MW, MX, MY, MZ, NA, NG, NI, NO, NZ, OM, PE, PG, PH, PL, PT, RO, RS, RU, SC, SD, SE, SG, SK, SL, SM, ST, SV, SY, TH, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, ZA, ZM, ZW.
- (84) **Designated States (unless otherwise indicated, for every kind of regional protection available):** ARIPO (BW, GH, GM, KE, LR, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AL, AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HR, HU, IE, IS, IT, LT, LU, LV,

[Continued on next page]

- (54)
- Title:**
- AUTOMATICALLY TUNING A TRANSMITTER TO A RESONANCE FREQUENCY OF A RECEIVER

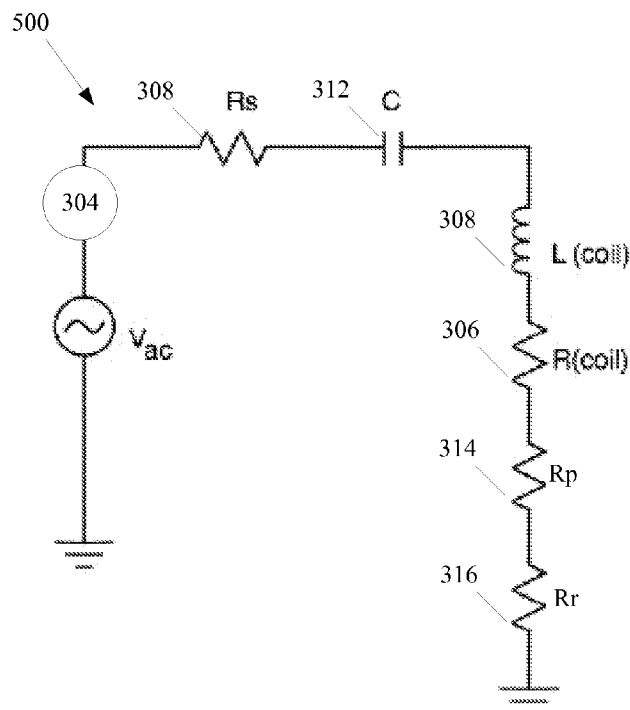


FIG. 5

(57) **Abstract:** Various embodiments of a wirelessly powered local computing environment are described. The wireless powered local computing environment includes at least a near field magnetic resonance (NFMR) power supply arranged to wirelessly provide power to any of a number of suitably configured devices. In the described embodiments, the devices arranged to receive power wirelessly from the NFMR power supply must be located in a region known as the near field that extends no further than a distance of a few times a characteristic size of the NFMR power supply transmission device. Typically, the distance can be on the order of 1 meter or so. The NFMR power supply is used to tune a transmitter in order to maintain a high Q factor.



MC, MK, MT, NL, NO, PL, PT, RO, RS, SE, SI, SK, **Published:**
SM, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, — *with international search report (Art. 21(3))*
GW, ML, MR, NE, SN, TD, TG).

Declarations under Rule 4.17:

— *of inventorship (Rule 4.17(iv))*

AUTOMATICALLY TUNING A TRANSMITTER TO A RESONANCE FREQUENCY OF A RECEIVER

TECHNICAL FIELD

[0001] The described embodiments relate generally to utilizing a wireless
5 power transmission in a portable computing environment.

BACKGROUND

[0002] Energy or power may be transferred wirelessly using a variety of
known radiative, or far-field, and non-radiative, or near-field, techniques. For
example, radiative wireless information transfer using low-directionality antennas,
10 such as those used in radio and cellular communications systems and home computer
networks, may be considered wireless energy transfer. However, this type of
radiative transfer is very inefficient because only a tiny portion of the supplied or
radiated power, namely, that portion in the direction of, and overlapping with, the
receiver is picked up. The vast majority of the power is radiated away in all the other
15 directions and lost in free space. Such inefficient power transfer may be acceptable
for data transmission, but is not practical for transferring useful amounts of electrical
energy for the purpose of doing work, such as for powering or charging electrical
devices.

[0003] One way to improve the transfer efficiency of some radiative energy
20 transfer schemes is to use directional antennas to confine and preferentially direct the
radiated energy towards a receiver. However, these directed radiation schemes may
require an unintermittible line-of-sight and potentially complicated tracking and
steering mechanisms in the case of mobile transmitters and/or receivers. In addition,
such schemes may pose hazards to objects or people that cross or intersect the beam
25 when modest to high amounts of power are being transmitted. A known non-
radiative, or near-field, wireless energy transfer scheme, often referred to as either
induction or traditional induction, does not (intentionally) radiate power, but uses an
oscillating current passing through a primary coil, to generate an oscillating magnetic
near-field that induces currents in a near-by receiving or secondary coil. Traditional
30 induction schemes have demonstrated the transmission of modest to large amounts of
power, however only over very short distances, and with very small offset tolerances
between the primary power supply unit and the secondary receiver unit. Electric

transformers and proximity chargers are examples of devices that utilize this known short range, near-field energy transfer scheme.

[0004] It is well known that useable power can be transferred wirelessly from a power source to a receiver located within a distance referred to as a near field. By
5 near field it is meant that within a distance less than a few times the geometric mean of the size of both objects involved in the transfer a relatively significant amount of power (at least on the order of tens of milli-watts) can be transferred between a wireless source device and a receiver. In this way, a realistic and practical approach to wireless transferring useable amounts of power over distances suitable for limited
10 applications can be realized. Typically, each battery powered device such as a wireless electronic device requires its own charger and power source, which is usually an alternating current (AC) power outlet. Such a wired configuration becomes unwieldy when many devices need charging.

[0005] However, the maximum power that can be wirelessly transferred from
15 a power transmitter to a receiver is dependent at least upon how closely tuned to each other resonance frequency the power transmitter and receiver are and if both the transmitter and receiver are operating at their resonance frequency. For example, any changes in the local environment can cause their resonance frequencies to change in such a way that the transmitter and receiver become de-tuned or no longer share the
20 same resonance frequency resulting in a substantial reduction in the amount of power that can be wirelessly transmitted.

[0006] Therefore, that is desired are methods, systems, and apparatus for maintaining a maximum power transfer between a wireless power transmitter and a wireless receiver in a wirelessly powered local computing environment.

25 [0007] The following are from Annals of Physics 323 (2008) 34-48 "Efficient Wireless *non-radiative mid-range* energy transfer" by Aristeidis Karalis et al., available online April 27, 2007.

1. Introduction

[0007a] In the early days of electromagnetism, before the electrical-wire grid
30 was deployed, serious interest and effort was devoted (most notably by Nikola Tesla) towards the development of schemes to transport energy over long distances without any carrier medium (e.g. wirelessly). These efforts appear to have met with little success. Radiative modes of omni-directional antennas (which work very well for

information transfer) are not suitable for such energy transfer, because a vast majority of energy is wasted into free space. Directed radiation modes, using lasers or highly-directional antennas, can be efficiently used for energy transfer, even for long distances (transfer distance $L_{\text{TRANS}} \gg L_{\text{DEV}}$, where L_{DEV} is the characteristic size of the device), but require existence of an uninterruptible line-of-sight and a complicated tracking system in the case of mobile objects. Rapid development of autonomous electronics of recent years (e.g. laptops, cell-phones, house-hold robots, that all typically rely on chemical energy storage) justifies revisiting investigation of this issue. Today, we face a different challenge than Tesla: since the existing electrical-wire grid carries energy almost everywhere, even a medium-range ($L_{\text{TRANS}} \approx \text{few}^* L_{\text{DEV}}$) wireless energy transfer would be quite useful for many applications. There are several currently used schemes, which rely on non-radiative modes (magnetic induction), but they are restricted to very close-range ($L_{\text{TRANS}} \ll L_{\text{DEV}}$) or very low-power ($\sim \text{mW}$) energy transfers.

15 **[0007b]** In contrast to all the above schemes, we investigate the feasibility of using long-lived oscillatory resonant electromagnetic modes, with localized slowly-evanescent field patterns, for efficient wireless non-radiative mid-range energy transfer. The proposed method is based on the well known principle of resonant coupling (the fact that two same-frequency resonant objects tend to couple, while
20 interacting weakly with other off-resonant environmental objects) and, in particular, resonant evanescent coupling (where the coupling mechanism is mediated through the overlap of the non-radiative near-fields of the two objects). This well known physics leads trivially to the result that energy can be efficiently coupled between objects in the extremely near field (e.g. in optical waveguide or cavity couplers and in resonant
25 inductive electric transformers). However, it is far from obvious how this same physics performs at mid-range distances and, to our knowledge, there is no work in the literature that demonstrates efficient energy transfer for distances a few times larger than the largest dimension of both objects involved in the transfer. In the present paper, our detailed theoretical and numerical analysis shows that such an
30 efficient mid-range wireless energy-exchange can actually be achieved, while suffering only modest transfer and dissipation of energy into other off-resonant objects, provided the exchange system is carefully designed to operate in a regime of “strong coupling” compared to all intrinsic loss rates. The physics of “strong

coupling'' is also known but in very different areas, such as those of light-matter interactions. In this favorable operating regime, we quantitatively address the following questions: up to which distances can such a scheme be efficient and how sensitive is it to external perturbations? The omni-directional but stationary (non-lossy) nature of the near field makes this mechanism suitable for mobile wireless receivers. It could therefore have a variety of possible applications including for example, placing a source (connected to the wired electricity network) on the ceiling of a factory room, while devices (robots, vehicles, computers, or similar) are roaming freely within the room. Other possible applications include electric-engine buses, RFIDs, and perhaps even nano-robots.

2. Range and rate of coupling

[0007c] The range and rate of the proposed wireless energy-transfer scheme are the first subjects of examination, without considering yet energy drainage from the system for use into work. An appropriate analytical framework for modeling this resonant energy-exchange is that of the well-known coupled-mode theory (CMT). In this picture, the field of the system of two resonant objects 1 and 2 is approximated by $F(r, t) \approx a_1(t)F_1(r) + a_2(t)F_2(r)$, where $F_{1,2}(r)$ are the eigenmodes of 1 and 2 alone, and then the field amplitudes $a_1(t)$ and $a_2(t)$ can be shown to satisfy, to lowest order:

$$\begin{aligned} \frac{da_1}{dt} &= -i(\omega_1 - i\Gamma_1)a_1 + i\kappa a_2 \\ \frac{da_2}{dt} &= -i(\omega_2 - i\Gamma_2)a_2 + i\kappa a_1 \end{aligned} \quad (1)$$

where $\omega_{1,2}$ are the individual eigenfrequencies, $\Gamma_{1,2}$ are the resonance widths due to the objects' intrinsic (absorption, radiation, etc.) losses, and κ is the coupling coefficient. Eqs. (1) show that at exact resonance ($\omega_1 = \omega_2$ and $\Gamma_1 = \Gamma_2$), the normal modes of the combined system are split by 2κ ; the energy exchange between the two objects takes place in time $\sim \pi/2\kappa$ and is nearly perfect, apart for losses, which are minimal when the coupling rate is much faster than all loss rates ($\kappa \gg \Gamma_{1,2}$). It is exactly this ratio $\kappa/\sqrt{\Gamma_1\Gamma_2}$ that we will set as our figure-of-merit for any system under consideration for wireless energy-transfer, along with the distance over which this ratio can be achieved. The desired optimal regime $\kappa/\sqrt{\Gamma_1\Gamma_2} \gg 1$ is called "strong-coupling" regime.

[0007d] Consequently, our energy-transfer application requires resonant modes of high $Q = \omega/2\Gamma$ for low (slow) intrinsic-loss rates Γ , and this is why we propose a

scheme where the coupling is implemented using, not the lossy radiative far-field, but the evanescent (non-lossy) stationary near-field. Furthermore, strong (fast) coupling rate κ is required over distances larger than the characteristic sizes of the objects, and therefore, since the extent of the near-field into the air surrounding a finite-sized resonant object is set typically by the wavelength (and quantified rigorously by the “radiation caustic”), this mid-range non-radiative coupling can only be achieved using resonant objects of subwavelength size, and thus significantly longer evanescent field-tails. This is a regime of operation that has *not* been studied extensively, since one usually prefers short tails to minimize interference with nearby devices. As will be seen in examples later on, such subwavelength resonances can often be accompanied with a high radiation- Q , so this will typically be the appropriate choice for the possibly-mobile resonant device-object d . Note, though, that the resonant source-object s will in practice often be immobile and with less stringent restrictions on its allowed geometry and size, which can be therefore chosen large enough that the near-field extent is not limited by the wavelength (using for example waveguides with guided modes tuned close to the “light line” in air for slow exponential decay therein).

[0007e] The proposed scheme is very general and *any* type of resonant structure (e.g. electromagnetic, acoustic, nuclear) satisfying the above requirements can be used for its implementation. As examples and for definiteness, we choose to work with two well-known, but quite different, electromagnetic resonant systems: dielectric disks and capacitively-loaded conducting-wire loops. Even without optimization, and despite their simplicity, both will be shown to exhibit acceptably good performance.

25 2.1. Dielectric disks

[0007f] Consider a 2D dielectric disk object of radius r and relative permittivity ϵ surrounded by air that supports high- Q “whispering-gallery” resonant modes (Fig. 1). The loss mechanisms for the energy stored inside such a resonant system are radiation into free space and absorption inside the disk material. High – Q^{rad} and long-tailed subwavelength resonances can be achieved, only when the dielectric permittivity ϵ is large and the azimuthal field variations are slow (namely of small principal number m). Material absorption is related to the material loss tangent: $Q^{abs} \sim \text{Re}\{\epsilon\}/\text{Im}\{\epsilon\}$. Mode-solving calculations for this type of disk resonances were

performed using two independent methods: numerically, 2D finite-difference frequency-domain (FDFD) simulations (which solve Maxwell's Equations in frequency domain exactly apart for spatial discretization) were conducted with a resolution of 30 pts/ r ; analytically, standard separation of variables (SV) in polar coordinates was used. The results for two TE-polarized dielectric-disk subwavelength modes of $\lambda/r \geq 10$ are presented in Fig. 10. The two methods have excellent agreement and imply that for a properly designed resonant low-loss-dielectric object values of $Q^{rad} \geq 2000$ and $Q^{abs} \sim 10,000$ should be achievable.

[0007g] Note that the required values of ε , shown in Fig. 10, might at first seem unrealistically large. However, not only are there in the microwave regime (appropriate for meter-range coupling applications) many materials that have both reasonably high enough dielectric constants and low losses (e.g. Titania: $\varepsilon \approx 96$, $\text{Im}\{\varepsilon\}/\varepsilon \approx 10^{-3}$; Barium tetratitanate: $\varepsilon \approx 37$, $\text{Im}\{\varepsilon\}/\varepsilon \approx 10^{-4}$; Lithium tantalite: $\varepsilon \approx 40$, $\text{Im}\{\varepsilon\}/\varepsilon \approx 10^{-4}$; etc.) [9,10], but also ε could signify instead the effective index of other known subwavelength ($\lambda/r \gg 1$) surface-wave systems, such as surface-plasmon modes on surfaces of metal-like (negative- ε) materials or metallo-dielectric photonic crystals.

[0007h] To calculate now the achievable rate of energy transfer between two disks 1 and 2, we place them at distance D between their centers (Fig. 11). Numerically, the FDFD mode-solver simulations give κ through the frequency splitting ($=2\kappa$) of the normal modes of the combined system, which are even and odd superpositions of the initial single-disk modes; analytically, using the expressions for the separation-of-variables eigenfields $\mathbf{E}_{1,2}(\mathbf{r})$ CMT gives κ through $\kappa = \omega_1/2 \cdot \int d^3\mathbf{r} \varepsilon_2(\mathbf{r}) \mathbf{E}_2^*(\mathbf{r}) \mathbf{E}_1(\mathbf{r}) / \int d^3\mathbf{r} \varepsilon(\mathbf{r}) |\mathbf{E}_1(\mathbf{r})|^2$, where $\varepsilon_j(r)$ are $\varepsilon(r)$ the dielectric functions that describe only the disk j (minus the constant ε_o background) and the whole space respectively. Then, for medium distances $D/r = 10 - 3$ and for non-radiative coupling such that $D < 2r_c$, where $r_c = m\lambda/2\pi$ is the radius of the radiation caustic, the two methods agree very well, and we find (Fig. 11) coupling-to-loss ratios in the range $\kappa/\Gamma \sim 1 - 50$. Although the achieved figure-of-merit values do not fall in the ideal "strong-coupling" operating regime $\kappa/\Gamma \gg 1$, they are still large enough to be useful for applications, as we will see later on.

2.2. Capacitively-loaded conducting-wire loops

[0007i] Consider a loop of radius r of conducting wire with circular cross-section of radius a connected to a pair of conducting parallel plates of area A spaced by distance d via a dielectric of relative permittivity ε and everything surrounded by air (Fig. 12). The wire has inductance L , the plates have capacitance C and then the system has a resonant mode, where the nature of the resonance lies in the periodic exchange of energy from the electric field inside the capacitor, due to the voltage across it, to the magnetic field in free space, due to the current in the wire. Losses in this resonant system consist of ohmic loss R_{abs} inside the wire and radiative loss R_{rad} into free space. Mode-solving calculations for this type of *RLC*-circuit resonances were performed using again two independent methods: numerically, 3D finite-element frequency-domain (FEFD) simulations (which solve Maxwell's Equations in frequency domain exactly apart for spatial discretization) were conducted, in which the boundaries of the conductor were modeled using a complex impedance $\eta_c \sqrt{\mu_c \omega / 2\sigma}$ boundary condition, valid as long as $\eta_c / \eta_o \ll 1$ [14] (10^{-5}) for copper in the microwave), where μ_o, ε_o and $\eta_o = \sqrt{\mu_o / \varepsilon_o}$ are the magnetic permeability, electric permittivity and impedance of free space and σ is the conductivity of the conductor; analytically, the formulas $L = \mu_o r [1n(8r/a) - 2]$ and $C = \varepsilon_o \varepsilon A / d$, and, in the desired subwavelength-loop ($r \ll \lambda$) limit, the quasi-static formulas $R_{\text{abs}} \approx \eta_c r / a$ (which takes skin-depth effects into account) and $R_{\text{rad}} \approx \pi / 6 \eta_o (r / \lambda)^4$ were used to determine the resonant frequency $\omega = 1 / \sqrt{LC}$ and its quality factors $Q^{\text{abs}} = \omega L / R_{\text{abs}}$ and $Q^{\text{rad}} = \omega L / R_{\text{rad}}$. By tuning the capacitance and thus the resonant frequency, the total Q becomes highest for some optimal frequency determined by the loop parameters: at low frequencies it is dominated by ohmic loss and at high frequencies by radiation. The results for two subwavelength modes of $\lambda / r \geq 70$ (namely highly suitable for near-field coupling and really in the quasi-static limit) at this optimal frequency are presented in Fig. 12. The two methods are again in very good agreement and show that expected quality factors in the microwave are $Q^{\text{abs}} \geq 1000$ and $Q^{\text{rad}} \geq 10,000$.

[0007j] For the rate of energy transfer between two loops 1 and 2 at distance D between their centers (Fig. 4): numerically, the FEFD mode-solver simulations give κ again through the frequency splitting $= 2\kappa$ of the normal modes of the combined system; analytically, κ is given by $\kappa = \omega M / 2\sqrt{L_1 L_2}$, where M is the mutual

inductance of the two loops, which, in the quasi-static limit $\ll d \ll \lambda$ and for the relative orientation shown in Fig. 13, is $M \approx \pi/2\mu_0(r_1r_2)^2/D^3$, which means that $\omega/2\kappa \sim (D/\sqrt{r_1r_2})^3$. Then, and for medium distances $D/r = 10$ -3, the two methods agree well, and we finally find (Fig. 13) coupling-to-loss ratios, which peak at a frequency between those where the single-loop $Q_{1,2}$ peak and are in the range $\kappa/\Gamma \sim 0.5 - 50$.

[0007k] It is important to appreciate the difference between such a resonant-coupling inductive scheme and the well-known non-resonant inductive scheme for energy transfer. Using CMT it is easy to show that, keeping the geometry and the energy stored at the source fixed, the resonant inductive mechanism allows for $\sim Q^2$ ($\sim 10^6$) times more power delivered for work at the device than the traditional non-resonant mechanism. This is why only close-range contact-less medium-power ($\sim W$) transfer is possible with the latter [2,3], while with resonance either close-range but large-power ($\sim kW$) transfer is allowed [4,5] or, as currently proposed, if one also ensures operation in the strongly-coupled regime, medium-range and medium-power transfer is possible. Capacitively-loaded conductive loops are actually being widely used also as resonant antennas (for example in cell phones), but those operate in the far-field regime with $D/r \gg 1, r/\lambda \sim 1$, and the radiation Q 's are intentionally designed to be small to make the antenna efficient, so they are not appropriate for energy transfer.

3. Influence of extraneous objects

[0007l] Clearly, the success of the proposed resonance-based wireless energy-transfer scheme depends strongly on the robustness of the objects' resonances. Therefore, their sensitivity to the near presence of random non-resonant extraneous objects is another aspect of the proposed scheme that requires analysis. The appropriate analytical model now is that of perturbation theory (**PT**), which suggests that in the presence of an extraneous object e the field amplitude $a_1(t)$ inside the resonant object 1 satisfies, to first order:

$$\frac{da_1}{dt} = -i(\omega_1 - i\Gamma_1)a_1 + i(\kappa_{11-e} + i\Gamma_{1-e})a_1 \quad (2)$$

where again ω_1 is the frequency and Γ_1 the intrinsic (absorption, radiation, etc.) loss rate, while κ_{11-e} is the frequency shift induced onto 1 due to the presence of e and Γ_{1-e} is the extrinsic due to e (absorption inside e , scattering from e , etc.) loss rate². The frequency shift is a problem that can be "fixed" rather easily by applying to

every device a feedback mechanism that corrects its frequency (e.g. through small changes in geometry) and matches it to that of the source. However, the extrinsic loss can be detrimental to the functionality of the energy-transfer scheme, because it cannot be remedied, so the total loss rate $\Gamma_{1[e]} = \Gamma_1 + \Gamma_{1-e}$ and the corresponding

- 5 figure-of-merit $\kappa_{[e]}/\sqrt{\Gamma_{1[e]}\Gamma_{2[e]}}$, where $\kappa_{[e]}$ the perturbed coupling rate, must be quantified.

3.1. Dielectric disks

- [0007m] In the first example of resonant objects that we have considered, namely dielectric disks, small, low-index, low-material-loss or far-away stray objects
10 will induce small scattering and absorption. In such cases of small perturbations these extrinsic loss mechanisms can be quantified using respectively the analytical first-order PT formulas $\Gamma_{1-e}^{rad} \propto \omega_1 \cdot \int d^3\mathbf{r} |\text{Re}\{\varepsilon_e(\mathbf{r})\}|^2 |\mathbf{E}_1(\mathbf{r})|^2 / U$ and $\Gamma_{1-e}^{abs} = \omega_1 / 4 \cdot \int d^3\mathbf{r} \text{Im}\{\varepsilon_e(\mathbf{r})\} |\mathbf{E}_1(\mathbf{r})|^2 / U$, where $U = 1/2 \cdot \int d^3\mathbf{r} \varepsilon(\mathbf{r}) |\mathbf{E}_1(\mathbf{r})|^2$ is the total resonate
15 electromagnetic energy of the unperturbed mode. As one can see, both of these losses depend on the *square* of the resonant electric field tails $\mathbf{E}_1$ at the site of the extraneous object. In contrast, the coupling rate from object 1 to another resonant object 2 is, as stated earlier, $\kappa = \omega_1 / 4 \cdot \int d^3\mathbf{r} \varepsilon_2(\mathbf{r}) \mathbf{E}_2^*(\mathbf{r}) \mathbf{E}_1(\mathbf{r}) / U$ and depends
linearly on the field tails $\mathbf{E}_1$ of 1 inside 2. This difference in scaling gives us confidence that, for exponentially small field tails, coupling to other resonant objects
20 should be much faster than all extrinsic loss rates ($\kappa \gg \Gamma_{1-e}$), at least for small perturbations, and thus the energy-transfer scheme is expected to be sturdy for this class of resonate dielectric disks.

- [0007n] However, we also want to examine certain possible situations where extraneous objects cause perturbations too strong to analyze using the above first-
25 order PT approach. For example, we place a dielectric disk c close to another off-resonance object of large $\text{Re}\{\varepsilon\}$, $\text{Im}\{\varepsilon\}$ and of same size but different shape (such as a human being h), as shown in Fig. 14A, and a roughened surface of large extent but of small $\text{Re}\{\varepsilon\}$, $\text{Im}\{\varepsilon\}$ (such as a wall w), as shown in Fig. 14B. For distances $D_{h/w}/r = 10-3$ between the disk-center and the “human”-center/“wall”, the numerical
30 FDFD simulation results presented in Fig. 14A-B suggest that $Q_{c[h]}^{rad}, Q_{c[w]}^{rad} \geq 1000$ (instead of the initial $Q_c^{rad} \geq 2000$, $Q_c^{abs} \sim 10,000$ (naturally unchanged), $Q_{c-h}^{abs} \sim 10^5-10^2$, and $Q_{c-w}^{abs} \sim 10^5-10^4$, namely the disk resonance seems to be fairly robust, since it

is not detrimentally disturbed by the presence of extraneous objects, with the exception of the *very* close proximity of the high-loss objects.

[0007o] To examine the influence of the large perturbations on an entire energy-transfer system we consider two resonant disks in the close presence of both a
 5 “human” and a “wall”. The numerical FDFD simulations show that the system performance deteriorates from $\kappa/\Gamma_c \sim 1\text{-}50$ (Fig. 11) to $\kappa_{[hw]}/\Gamma_{c[hw]} \sim 0.5\text{-}10$ (Fig. 15), i.e. only by acceptably small amounts.

3.2. Capacitively-loaded conducting-wire loops

[0007p] In the second example of resonant objects that we have considered, the
 10 conducting-wire loops, the influence of extraneous objects on the resonances is nearly absent. The reason is that, in the quasi-static regime of operation ($r \ll \lambda$) that we are considering, the near field in the air region surrounding the loop is predominantly magnetic (since the electric field is localized inside the capacitor), therefore extraneous non-metallic objects e that could inter- act with this field and act as a
 15 perturbation to the resonance are those having significant magnetic properties (magnetic permeability $\text{Re}\{\mu\} > 1$ or magnetic loss $\text{Im}\{\mu\} > 0$). Since almost all every-day materials are non-magnetic, they respond to magnetic fields in the same way as free space, and thus will not disturb the resonance of a conducting-wire loop. To get only a rough estimate of this disturbance, we use the **PT** formula, stated
 20 earlier, $\Gamma_{1-e}^{abs} = \omega_1/4 \cdot \int d^3\mathbf{r} \text{Im}\{\varepsilon_e(\mathbf{r})\} |\mathbf{E}_1(\mathbf{r})|^2/U$ with the numerical results for the field example like the one shown in the plot of Fig. 13 and with a rectangular object of dimensions 30 cm $\times$ 30 cm $\times$ 1.5 m and permittivity $\varepsilon = 49 + 16i$ (human muscles) residing between the loops and almost standing on top of one capacitor (~ 3 cm away from it) and find $Q_{c-h}^{abs} \sim 10^5$ and for ~ 10 cm away $Q_{c-h}^{abs} \sim 5 \times 10^5$. Thus, for ordinary
 25 distances ($\sim 1\text{m}$) and placements (not immediately on top of the capacitor) or for most ordinary extraneous objects e of much smaller loss-tangent, we conclude that it is indeed fair to say that $Q_{c-e}^{abs} \rightarrow \infty$ and that $\kappa_{[e]} / \Gamma_{[e]} \sim \kappa / \Gamma \sim 0.5\text{-}50$. The only perturbation that is expected to affect these resonances is a close proximity of large metallic structures.

30 [0007q] An extremely important implication of this fact relates to safety considerations for human beings. Humans are also non-magnetic and can sustain strong magnetic fields with- out undergoing any risk. A typical example, where

magnetic fields $B \sim 1 \text{ T}$ are safely used on humans, is the magnetic resonance imaging (MRI) technique for medical testing. In contrast, the magnetic near-field required by our scheme in order to provide a few Watts of power to devices is only $B \sim 10^{-4} \text{ T}$, which is actually comparable to the magnitude of the Earth's magnetic field. Since, as explained above, a strong electric near-field is also not present and the radiation produced from this non-radiative scheme is minimal, it is reasonable to expect that our proposed energy-transfer method should be safe for living organisms.

[0007r] In comparison of the two classes of resonant systems under examination, the strong immunity to extraneous objects and the absence of risks for humans probably makes the conducting-wire loops the preferred choice for many real-world applications; on the other hand, systems of disks (or spheres) of high (effective) refractive index have the advantage that they are also applicable to much smaller length-scales (for example in the optical regime dielectrics prevail, since conductive materials are highly lossy).

15 **4. Efficiency of energy-transfer scheme**

[0007s] Consider again the combined system of a resonant source s and device d in the presence of a set of extraneous objects e , and let us now study the efficiency of this resonance-based energy-transfer scheme, when energy is being drained from the device at rate Γ_{work} for use into operational work. The coupled-mode-theory equation for the device field-amplitude is

$$\frac{da_d}{dt} = -i(\omega - i\Gamma_{d[e]})a_d + i\kappa_{[e]}a_s - \Gamma_{\text{work}}a_d \quad (3)$$

Where $\Gamma_{d[e]} = \Gamma_{d[e]}^{\text{rad}} + \Gamma_{d[e]}^{\text{abs}} = \Gamma_{d[e]}^{\text{rad}} + (\Gamma_d^{\text{abs}} + \Gamma_{d-e}^{\text{abs}})$ is the net perturbed-device loss rate, and similarly we define $\Gamma_{s[e]}$ for the perturbed-source. Different temporal schemes can be used to extract power from the devices (e.g. steady-state continuous-wave drainage, instantaneous drainage at periodic times and so on) and their efficiencies exhibit different dependence on the combined system parameters. Here, we assume steady state, such that the field amplitude inside the source is maintained constant, namely $a_s(t) = A_s e^{-i\omega t}$ to then the field amplitude inside the device is $a_d(t) = A_d e^{-i\omega t}$ with $A_d/A_s = i\kappa_{[e]}/(\Gamma_{d[e]} + \Gamma_{\text{work}})$. The various time-averaged powers of interest are than: the useful extracted power is $P_{\text{work}} = 2\Gamma_{\text{work}}|A_d|^2$, the radiated (including scattered) power is $P_{\text{rad}} = 2\Gamma_{s[e]}^{\text{rad}}|A_s|^2 + 2\Gamma_{d[e]}^{\text{rad}}|A_d|^2$, the power absorbed

at the source/device is $P_{\text{rad}} = 2\Gamma_{s/d}^{\text{abs}} |A_{s/d}|^2$, and at the extraneous objects $P_e = 2\Gamma_{s-e}^{\text{abs}} |A_s|^2 + 2\Gamma_{d-e}^{\text{abs}} |A_d|^2$. From energy conservation, the total time-averaged power entering the system is $P_{\text{total}} = P_{\text{work}} + P_{\text{rad}} + P_s + P_d + P_e$. Note that the reactive powers, which are usually present in a system and circulate stored energy around it, cancel at resonance (which can be proven for example in electromagnetism from Poynting's Theorem) and do not influence the power-balance calculations. The working efficiency is then:

$$\eta_{\text{work}} \equiv \frac{P_{\text{work}}}{P_{\text{total}}} = \frac{1}{1 + \frac{\Gamma_{d[e]}}{\Gamma_{\text{work}}} \left[1 + \frac{1}{\text{fom}_{[e]}^2} \left(1 + \frac{\Gamma_{\text{work}}}{\Gamma_{d[e]}} \right)^2 \right]}, \quad (4)$$

where $\text{fom}_{[e]} = \kappa_{[e]} / \sqrt{\Gamma_{s[e]} \Gamma_{d[e]}}$ is the distance-dependent figure-of-merit of the perturbed resonant energy-exchange system. Depending on the targeted application, reasonable choices for the work-drainage rate are: $\Gamma_{\text{work}} / \Gamma_{d[e]} = 1$ to minimize the required energy stored in the source, $\Gamma_{\text{work}} / \Gamma_{d[e]} = \sqrt{1 + \text{fom}_{[e]}^2} > 1$ to maximize the efficiency for some particular value of $\text{fom}_{[e]}$ or $\Gamma_{\text{work}} / \Gamma_{d[e]} \gg 1$ to minimize the required energy stored in the device. For any of these choices, η_{work} is a function of the $\text{fom}_{[e]}$ parameter only. η_{work} is shown for its optimal choice in Fig. 16 with a solid black line, and is $\eta_{\text{work}} > 17\%$ for $\text{fom}_{[e]} > 1$, namely large enough for practical applications. The loss conversion ratios depends also on the other system parameters, and the most disturbing ones (radiation and absorption in stray objects) are plotted in Fig. 16 for the two example systems of dielectric disks and conducting loops with value for their parameters within the ranges determined earlier.

[0007t] To get a numerical estimate for a system performance, take, for example, coupling distance $D/r = 5$, a “human” extraneous object at distance $D_h/r = 10$, and that $P_{\text{work}} = 10\text{W}$ must be delivered to the load. Then, for dielectric disks we have (based on Fig. 14A-B) $Q_{s[h]}^{\text{rad}} = Q_{d[h]}^{\text{rad}} \sim 10^3$, $Q_s^{\text{abs}} = Q_d^{\text{ags}} \sim 10^4$, $Q_{s-h}^{\text{abs}} = Q_{d-h}^{\text{abs}} \sim 5 \times 10^4$ and (based on Figs. 11 and 15) $\text{fom}_{[h]} \sim 3$, so from Fig. 16 we find efficiency $\eta_{\text{work}} = 52\%$ and that $P_{\text{rad}} \approx 8.3\text{W}$ will be radiated to free space, $P_s \approx 0.5\text{W}$ will be dissipated inside the source, $P_d \approx 0.3\text{W}$ inside the device, and $P_h \approx 0.2\text{W}$ inside the human. On the other hand, for conducting loops we have (based on Figs. 3 and 4) $Q_{s[h]}^{\text{rad}} = Q_{d[h]}^{\text{rad}} \sim 10^4$, $Q_s^{\text{abs}} = Q_d^{\text{ags}} \sim 10^3$, $Q_{s-h}^{\text{abs}} = Q_{d-h}^{\text{abs}} \rightarrow \infty$ and $\text{fom}_{[h]} \sim 4$, so

we find $\eta_{\text{work}} = 61\%$, $P_{\text{rad}} \approx 0.6\text{W}$, $P_s \approx 3.6\text{W}$, $P_d \approx 2.2\text{W}$, and most importantly $P_h \rightarrow 0$.

5. Conclusion

[0007u] In conclusion, we present a scheme based on “strongly-coupled” resonances for mid-range wireless non-radiative energy transfer. Although our consideration has been for a static geometry (namely κ and Γ_e were independent of time). All the results can be applied directly for the dynamic geometries of mobile objects, since the energy-transfer time κ^{-1} ($\sim 1\text{-}100\ \mu\text{s}$ for microwave application) is much shorter than any timescale associated with motions of macroscopic objects.

Analyses of very simple implementation geometries provide encouraging performance characteristics and further improvement is expected with serious design optimization. Thus the proposed mechanism is promising for many modern applications. For example, in the macroscopic world, this scheme could potentially be used to deliver power to roots and/or computers in a factory room, or electric buses on a highway (source-cavity would in this case be a “pile” running above the highway). In the micorscopic world, where much smaller wavelengths would be used and smaller powers are needed, one could use it to implement optical inter-connects for CMOS electronics, or to transfer energy to autonomous nano-objects (e.g. MEMS or nan-robots) without worrying much about the relative alignment between the sources and the devices.

[0007v] As a venue of future scientific research, enhanced performance should be pursued for electromagnetic systems either by exploring different materials, such as plasmonic or metallodielectric structures of large effective refractive index, or by fine-tuning the system design, for example by exploiting the earlier mentioned interference effects between the radiation fields of the coupled object. Furthermore, the range of applicability could be extended to acoustic systems, where the source and device are connected via a common condensed-matter object.

SUMMARY

[0008] The present invention provides a system and method for utilizing wireless near field magnetic resonance (NFMР) power transmission in a computing environment. In particular, techniques for maintaining a shared resonance frequency in the transmitter and receiver in a magnetic resonance power transfer system are described.

[0009] In one embodiment, a wireless power transmission unit is described. The wireless power transmission unit includes at least a varying frequency excitation source arranged to provide a variable circuit excitation at frequency f and a first resonator structure coupled to the varying frequency f excitation source, the first resonator structure having a resonant frequency ω_1 , a first Q factor Q_1 , and a characteristic size L_1 . Non-radiative energy transfer between the first resonator structure and a second resonator structure positioned apart from the wireless power transmission unit, the second resonator structure having a resonant frequency ω_2 , a second Q factor Q_2 , and a characteristic size L_2 is mediated through a magnetic resonant coupling associated with evanescent tails between the first resonator structure and the second resonator structure when the first resonator structure and the second resonator structure are in a tuned state. The first and second resonator structures are in the tuned state when the first resonant frequency ω_1 and the second resonant frequency ω_2 are within the narrower of first resonance width Γ_1 or second resonance width Γ_2 such that the first resonant frequency ω_1 and the resonant frequency ω_2 are separated by a bandwidth of no more than 3db. The first resonator structure maintains the first resonator structure and the second resonator structure in the tuned state by varying the first resonant frequency ω_1 in order to maintain the first resonant frequency ω_1 and the second resonant frequency ω_2 such that the first resonant frequency ω_1 and the resonant frequency ω_2 are separated by the bandwidth of no more than 3db in accordance with any changes in the second resonant frequency ω_2 .

[0010] In another embodiment, a method performed by a wireless power transmission unit for automatically tuning a wireless power transmission system is described. In the described embodiment, the wireless power transmission unit includes at least a first resonator structure coupled to a varying frequency f voltage supply. The method can be performed by establishing a magnetic resonant inductive

coupling between the first resonator structure to a second resonator structure positioned apart from the wireless power transmission unit in a tuned state, non-radiatively transferring useable power from the first resonator structure to the second resonator structure using the magnetic resonant inductive coupling between the first resonator structure and the second resonator structure. In the tuned state, a first resonant frequency ω_1 of the first resonator structure and the second resonant frequency ω_2 of the second resonator structure are separated by a bandwidth of no more than 3db. The first resonator structure maintains the first resonator structure and the second resonator structure in the tuned state by varying a first resonant frequency ω_1 of the first resonant structure in order to maintain the first resonant frequency ω_1 and a second resonant frequency ω_2 are separated by the bandwidth of no more than 3db in accordance with any change in the second resonant frequency ω_2 .

[0011] In yet another embodiment, non-transitory computer readable medium for storing a computer program executable by a processor in a wireless power transmission unit for automatically tuning a wireless power transmission system is described. In the described embodiment, the wireless power transmission unit includes at least a first resonator structure coupled to a varying frequency f voltage supply. The computer readable medium includes at least computer code for establishing a magnetic resonant inductive coupling between the first resonator structure to a second resonator structure positioned apart from the wireless power transmission unit in a tuned state, computer code for non-radiatively transferring useable power from the first resonator structure to the second resonator structure using the magnetic resonant inductive coupling between the first resonator structure and the second resonator structure, wherein in the tuned state, a first resonant frequency ω_1 of the first resonator structure and the second resonant frequency ω_2 of the second resonator structure are separated by a bandwidth of no more than 3db; and computer code for maintaining the first resonator structure and the second resonator structure in the tuned state by first resonator structure by varying a first resonant frequency ω_1 of the first resonant structure in order to maintain the first resonant frequency ω_1 and a second resonant frequency ω_2 are separated by the bandwidth of no more than 3db in accordance with any change in the second resonant frequency ω_2 .

[0012] In still another embodiment, a wirelessly powered local computing environment is described that includes at least a variable frequency power supply

arranged to provide power at a frequency, a resonance frequency-adjustable near field magnetic resonance (NFMR) wireless transmitter coupled to the power supply and arranged to create a resonance channel used to transfer useable energy from the power supply to a receiver within a near field distance d , and a central processing unit, the

5 central processing unit providing processing resources to the variable frequency power supply and the resonance frequency-adjustable NFMR transmitter. In the described embodiment, when the NFMR transmitter establishes a resonance coupling with the receiver, the central processing unit maintains the resonance frequency by adjusting the NFMR transmitter.

10 **[0013]** In another embodiment, a method performed by a processor for auto-tuning the resonance frequency of a near field magnetic resonance (NFMR) transmitter to its receiver is described. In the described embodiment, the NFMR transmitter is arranged to create a resonance channel used to transfer useable energy from a variable frequency power supply to a receiver within a near field distance d of

15 the transmitter, the variable frequency power supply is arranged to provide the power to the NFMR transmitter at a frequency. The method is carried out by performing at least the following operations: providing power to the NFMR transmitter at a frequency by the variable frequency power supply, adjusting the NFMR transmitter to resonate at the power supply frequency, monitoring peak real power supplied to the

20 resonating NFMR transmitter by the power supply, and setting the frequency of the variable frequency power supply to the current frequency when the monitored peak real power supplied is less than peak real power supplied at other frequency when the NFMR is also resonating by at least a pre-determined amount.

[0014] In yet another embodiment, a non-transitory computer readable

25 medium for storing computer code executed by a processor for auto-tuning a near field magnetic resonance (NFMR) transmitter is described. In the described embodiment, the NFMR transmitter is arranged to create a resonance channel used to transfer useable energy from a variable frequency power supply to a receiver within a near field distance d of the transmitter, the variable frequency power supply is

30 arranged to provide the power to the NFMR transmitter at a frequency. The computer readable medium includes at least computer code for providing power to the NFMR transmitter at the frequency by the variable frequency power supply, computer code for maintaining the NFMR transmitter in resonance mode, computer code for

changing the frequency of the variable frequency power supply to a current frequency, computer code for monitoring a current generated in response to the provided power at the current frequency, and computer code for setting the frequency of the variable frequency power supply to the current frequency when the monitored
5 current is less than an average current by at least a pre-determined amount.

[0015] Other apparatuses, methods, features and advantages of the described embodiments will be or will become apparent to one with skill in the art upon examination of the following figures and detailed description. It is target that all such additional apparatuses, methods, features and advantages be included within this
10 description be within the scope of and protected by the accompanying claims.

BRIEF DESCRIPTION OF THE DRAWINGS

[0016] The included drawings are for illustrative purposes and serve only to provide examples of possible structures and arrangements for the disclosed embodiments. These drawings in no way limit any changes in form and detail that
15 may be made to the described embodiments by one skilled in the art without departing from the spirit and scope of the embodiments.

[0017] FIG. 1 shows an exemplary system diagram of an auto-tuning wireless power transfer system in accordance with the described embodiments.

[0018] FIGS. 2A – 2B shows embodiments of an RLC equivalent circuit of a
20 transmitter having adjustable capacitive and inductive elements, respectively.

[0019] FIG. 3 shows representative P_{real} family of curves for a transmitter in accordance with the described embodiments.

[0020] FIG. 4 represents an RLC equivalent circuit of a transmitter with parasitic resistance in accordance with the described embodiments.

25 [0021] FIG. 5 shows another embodiment of RLC equivalent circuit that models a receiver in proximity to a transmitter in accordance with the described embodiments.

[0022] FIG. 6 shows P_{real} curve representing presence or resonating receiver based upon RLC equivalent circuit of FIG. 5.

30 [0023] FIG. 7 shows a flow chart detailing a process for automatically tuning a transmitter in a wireless circuit.

[0024] FIG. 8 shows representative virtual charging area in accordance with the described embodiments.

[0025] FIG. 9 shows a computing system in accordance with the described embodiments.

DETAILED DESCRIPTION

[0026] Various embodiments of a wirelessly powered local computing environment are described. The wireless powered local computing environment includes at least a near field magnetic resonance (NFMR) power supply arranged to wirelessly provide power to any of a number of suitably configured devices. In the described embodiments, the devices arranged to receive power wirelessly from the NFMR power supply can be located in a region known as the near field that extends about a distance D that can be a few times a characteristic size of the NFMR power supply transmission device. Typically, the distance D can be on the order of 1 meter or so. The local wireless environment can be affected by many things such as the relative positions of the NFMR power transmitter and receivers, obstacles that can affect the distribution of magnetic energy in the NFMR field that can adversely affect the resonance frequency of the system as a whole and those of the NFMR power transmitter and receiver specifically. This effective de-tuning of resonance frequency can effectively reduce the overall efficiency of the system (i.e., the Q_{system} factor) which can result in a concomitant reduction in the amount of useable power that can ultimately be transferred.

[0027] In the context of this discussion, it is well known that useable power can be wirelessly transmitted by way of a wireless transmitter and receiver transiently coupled by way of a magnetic field. More specifically, a system in accordance with the described embodiments can include a wireless power receiver incorporated within or electrically coupled to a peripheral device that can wirelessly receive useful amounts of power from an external power supply. In the described system, the wireless power receiver can include a first resonator structure having a first resonant frequency ω_1 , a first Q factor Q_1 (a measure of power transfer efficiency), and a first characteristic size L_1 . For example, in the context of a computing system where the peripheral device takes the form of a computer mouse or other input device, the characteristic size L_1 can be on the order of a few inches or centimeters. The system can also include a power supply coupled to at least a second resonator structure positioned a variable distance d from the first resonator structure having a second resonant frequency ω_2 and a second Q factor Q_2 and second characteristic size L_2 .

For example, the second resonator structure can be incorporated within a computer such as a desktop or laptop computer. In this way, a charging region can be formed around the computer in which the peripheral device (or any other appropriately configured device) can wirelessly receive useful amounts of power from the power supply via the second resonator structure.

[0028] When first and second resonant frequencies ω_1 and ω_2 are close together, a coupling region between the first and second resonator structures can be formed. Within this coupling region, useful power can be transferred by way of a non-radiative energy transfer mechanism that utilizes resonant-field evanescent tails of the magnetic field provided by the second resonator structure. For example, when the first resonator structure is incorporated into a peripheral device such as a computer mouse, the computer mouse can be at least partially supplied with power from the second resonator structure coupled to the power supply mediated by the magnetic field. In this way, the peripheral device can operate in a wireless mode without the need for a power supply other than that provided in the computing device. The amount and duration of power that can be delivered to the peripheral device can be dependent on a number of factors. For example, the coupling between the first resonator structure and the second resonator structure can be highly dependent upon maintaining a tuned state between the first and second resonant structures.

[0029] In a magnetic resonance power transfer system, at least a transmitter and a receiver are resonantly coupled in such a way that useable power can be transferred from a power supply by way of the transmitter to at least one receiver. In order to transfer useable power, both the transmitter and receiver are tuned to a resonance frequency characteristic of the magnetic resonance power transfer system. A quality factor or Q factor is a dimensionless parameter that can be used as a measure of the ability of the resonant system to transfer energy from the transmitter to the receiver. More specifically, the Q factor is the ratio of power stored to power dissipated in reactance and resistance elements in the circuit according to Eq. (1):

$$\frac{P_{\text{stored}}}{P_{\text{dissipated}}} = I^2 X / I^2 R \quad \text{Eq. (1)}$$

where X is the capacitive/inductive reactance, and R is series resistance.

The Q factor can also be used to characterize a resonator's bandwidth (BW) relative to its center frequency in accordance with Eq. (2):

$$BW = \Delta\omega = \omega_h - \omega_l = \omega_c/Q \quad \text{Eq. (2)}$$

where ω_h = high band edge, ω_l = low band edge, and

ω_c = center frequency (resonant frequency)

$$\omega_l = \omega_c - \Delta\omega/2$$

$$\omega_h = \omega_c + \Delta\omega/2.$$

[0030] Therefore, systems having high Q factors have necessarily narrow bandwidths (i.e., $\Delta\omega/2$ decreases as Q increases). For example, a wireless power transfer system with a system quality factor Q_{system} of 100 or more can be achieved by properly tuning the transmitter and receiver. However, the higher the Q factor, the narrower the associated frequency range. For example, a wireless power transfer system can only maintain system quality factor Q_{system} of at least 100 by tuning the resonant frequencies of the transmitter and receiver to within an extremely tight (such as 3dB) bandwidth of each other. In other words, the transmitter and receiver must be very closely tuned to the same frequency for significant power transfer to occur. If the transmitter and receiver become de-tuned beyond the 3dB bandwidth, the ability of the wireless power transfer system to magnetically transfer power exponentially decays. The detuning of the system can occur if either a) the transmitter or receiver detunes (i.e not resonating) or b) the resonance frequency of the transmitter and receiver do not align resulting from interference created by extraneous objects (typically ferro-magnetic) that capacitively couple with either or both the transmitter or receiver.

[0031] For example, if either the transmitter and/or receiver are placed in an environment in which the transmitter and/or receiver are in proximity to metallic or ferromagnetic objects, the resonance frequency of either can deviate significantly from design due to magnetic or capacitive coupling to the objects in the environment. Therefore, without any modifications, the transmitter and receiver can de-tune to such a degree that the system Q factor drops precipitously resulting in a substantial, if not complete, loss in power transfer between the transmitter and receiver.

[0032] Accordingly, the following discussion describes techniques for automatically maintaining a NFMR power transfer system in a tuned state thereby maintaining a high system Q factor. In one embodiment, the system can be

maintained in a tuned state by automatically tuning a transmitter to match the resonant frequency of an associated receiver. The high system Q factor, in turn, facilitates an efficient transfer of power from the transmitter to the receiver having the effect of providing useful amounts of power wirelessly to a device coupled to the receiver. In one embodiment, the transmitter is actively tuned in order to maintain resonance coupling between the transmitter and receiver. Tuning the transmitter is an advantageous technique since the transmitter is typically tethered to a power source that can be frequency adjusted.

[0033] In one embodiment, the transmitter can be coupled to an oscillator circuit that can generate a frequency used to excite the transmitter. The oscillator circuit can take the form of a variable oscillator circuit that can be adjusted via a variable clock circuit or a VCO (voltage controlled oscillator). Moreover, a capacitance or inductance of the resonant system can be modified in such a way that the resonance frequency of the transmitter is within an acceptable range of the resonance frequency of the system. In this way, maximum power can be transferred from the power supply via the transmitter to the receiver that can be used to provide power to a device. In one embodiment, a variable capacitance circuit can be used to change the system capacitance leaving the system inductance unchanged. Other techniques can be used in other embodiments.

[0034] These and other embodiments are discussed below with reference to Figs. 1-9. However, those skilled in the art will readily appreciate that the detailed description given herein with respect to these figures is for explanatory purposes only and should not be construed as limiting.

[0035] FIG. 1 shows an exemplary system diagram of an auto-tuning wireless power transfer system in accordance with the described embodiments. Magnetic resonance power transfer system 10 can include transmitter 100 configured to supply power wirelessly to receiver 200 by way of magnetic field B_{NFMR} . Transmitter 100 can include at least transmitter resonator or resonator configured as coil 102 (Tx Coil). Similarly, receiver 200 can be configured as having a receiver resonator or resonator as coil 202 (Rx Coil). In some aspects, transmitter coil 102 (Tx Coil) and/or the receiver coil 202 (Rx Coil) can be presumed to be essentially a two-dimensional structure. In the described embodiment, Tx coil 102 can be coupled to transmitter impedance-matching circuit 104 that can be used to match the impedances between

Tx coil 102 and Rx coil 202. In the described embodiment, Rx coil 202 can be coupled to receiver load device 204 that receives useful amounts of power from power supply 106 connected to transmitter 100. Transmitter 100 can wirelessly transfer power received from power supply 106 using Tx Coil 102 resonantly coupled with Rx Coil 202. Power received at Rx Coil 202 can be passed to load device 204. In the described embodiment, tuning circuit 108 can be used to automatically adjust characteristics of transmitter 100 in order that a system quality factor remains greater than a pre-determined threshold value. In one embodiment, tuning circuit 108 can detect the presence of a resonating receiver unit and adjust the resonance frequency of transmitter 100 accordingly.

[0036] FIGS. 2A – 2B shows embodiments of an RLC equivalent circuit of a transmitter having adjustable capacitive and inductive elements, respectively. For example, as shown in FIG. 2A and FIG. 2B, RLC circuit 300 can include various reactive elements such as capacitors and inductors. RLC circuit 300 can also include various resistive elements representing various aspects of transmitter 100. For example, various parasitic power losses inherent in transmitter 100 can be characterized as a series resistor. It should be further noted that the overall impedance Z of RLC circuit 300 can be a combination of resistance values of the various resistive elements as well as reactive values of constituent inductors and capacitors. It should be noted, however, that the electrical characteristics of inductors and capacitors and therefore their respective reactance values are frequency dependent. At a resonance frequency, the inductance and capacitance reactive values are equal in magnitude but 180° out of phase having the effect that the reactive values of the inductors and capacitors cancel each other out resulting a sharp drop in overall circuit impedance Z . This sharp drop in overall circuit impedance Z can be used to automatically and periodically detect a resonant frequency of RLC circuit 300 in a manner described below in more detail.

[0037] More particularly, as shown in FIG. 2A, RLC circuit 300 can include voltage source (V_s) 302 in the form of a variable excitation source along the lines of an AC power supply arranged to provide variable voltage V_{AC} . It should be noted that in terms of this discussion, alternating voltage V_{AC} need not be sinusoidal in nature in which case alternating voltage V_{AC} can have fundamental frequency f Hz represented as $\omega = 2\pi f$. In the described embodiment, V_s 302 can be connected in series with

power detector 304. Power detector 304 can be used to detect current I_{rms} as well as real power developed in circuit 300 in response to the application of alternating voltage V_{AC} and calculate the real power supplied by the power source 302. Those skilled in the art will realize that 302 can also be also an alternating current source instead of alternating voltage source with similar end goal of supply power to circuit 300 with variable frequency.

[0038] RLC circuit 300 can also include coil resistor R_{coil} 306 that can represent the resistance or power loss inherent in Tx Coil 102. Series resistor R_s 308 can represent the resistance or power loss in the remainder of circuit 300 associated with, for example, PCB traces, internal resistance of V_s 302 and so on. Coil inductance can be represented as coil inductor 310, or more simply L_{coil} 310 having reactance $X_L (= \omega L_{coil})$ and variable capacitor (or capacitor bank) C 312 having reactance $X_C (= 1/\omega C_{coil})$ that can be varied, or tuned, such that circuit 300 resonates at which point $X_L = X_C$ resulting in a sharp drop in circuit impedance Z as shown by Eq. (3a) and Eq. (3b) and Eq. (3c):

$$I_{rms} = V_{rms} / Z \quad \text{Eq. (3a)}$$

$$\text{where } Z = \sqrt{R^2 + (X_L - X_C)^2} \quad \text{Eq. (3b)}$$

$$Z = R_{circuit} = R_s + R_{coil} \quad \text{Eq. (3c)}$$

[0039] Therefore at resonance (i.e., $X_L = X_C$), circuit impedance Z simply reduces to $Z = R_{circuit}$. In this way by incrementing frequency f of V_s 302 and tuning capacitor C 312 to maintain RLC circuit 300 in resonance, power detector 304 will measure real power as approximately a straight line vs resonance frequency of circuit 300 as illustrated in FIG. 3 as line 402 (presuming little or no frequency dependence of the resistive elements) and according to Eq. (4) showing the relationship between P_{Real} , I_{RMS} and V_{RMS} that can be used to characterize transmitter 100 in resonance:

$$P_{Real} = (I_{rms})^2 R_{circuit} = (V_{rms})^2 / R_{circuit} \quad \text{Eq. (4).}$$

In place of or in addition to variable capacitor C 312, RLC circuit 300 can include variable inductor L 312 as shown in FIG. 2B.

[0040] It should be noted that parasitic lossy elements along the lines of metallic or magnetic objects in the vicinity of Tx Coil 102 can be represented as an additional resistive element along the lines of R_p 314 shown in FIG. 4 as circuit 400. The addition of parasitic resistance R_p increases circuit resistance $R_{circuit}$ resulting in a reduction in P_{Real} as shown in FIG. 3 as curve 404. In this way, transmitter 100 can be

characterized by a family of P_{Real} curves. In some cases, the P_{Real} curves can be frequency independent such as shown in FIG. 3 as curve 402, 404 or frequency dependent shown in FIG. 3 as curve 406. In the frequency dependent scenario, some of the resistive elements can be sensitive to frequency f in that their respective

5 resistive values can change as frequency f changes as a result of, for example, changing frequency f during a frequency sweep operation. For example, in the scenario represented by FIG. 3 with regards to curve 406, at least some of the resistive elements can be exhibit higher resistance values when frequency f is high as opposed to low resulting in curve 406 having a non-zero slope (in the example, shown the

10 linearity of curve 406 is maintained by presuming that all resistive elements are equally sensitive to frequency, which may not be true in all cases).

[0041] FIG. 5 shows another embodiment of circuit 300 that models a receiver in proximity to transmitter 100 as circuit 500. The receiver load can be modeled as receiver resistance R_r such that at resonance the circuit impedance Z_{circuit} satisfies Eq.

15 (6):

$$Z_{\text{circuit}} = R_s + R_{\text{coil}} + R_p + R_r \quad \text{Eq. (6).}$$

[0042] In this way, the value of Z_{circuit} will increase due to the presence of the resonant receiver causing P_{Real} to drop around where the resonance frequency of transmitter 100 lines up with the resonance frequency of receiver 200 as shown in

20 FIG. 6 indicating the presence of a resonating receiver.

[0043] Accordingly, by using both frequency sweeping and maintaining the transmitter equivalent circuit in resonance, a receiver resonance frequency can be automatically determined and the transmitter tuned in accordance with the following procedure. Firstly, transmitter 100 can be configured to include a variable reactance

25 element such as a variable capacitor along the lines of capacitor C 312. Transmitter 100 can be provided power by a variable power supply that can supply a variable frequency voltage $V(f)$ that can be incremented from a low frequency value f_{LOW} to a high frequency value f_{HIGH} during a frequency sweep operation. A power detector coupled in series with the variable frequency voltage supply can be used to detect the

30 circuit real power P_{Real} .

[0044] During the frequency sweep operation, the power detector provides a periodic value of P_{Real} which is measured against a previous value of P_{Real} in order to determine a value of frequency f corresponding to a pre-determined drop in value of

P_{Real} with respect to a previous value of P_{Real} . In the described embodiment, the frequency or range of frequencies corresponding to the detected pre-determined drop 408 in value of P_{Real} can be associated with a resonant frequency ω_R of receiver 200 as that can be used to tune transmitter resonant frequency ω_T . In one embodiment,

5 transmitter resonant frequency ω_T can be determined using an iterative process in which a range of frequencies corresponding to drop 408 can be used to determine a center frequency corresponding to a desired transmitter resonant frequency ω_T .

[0045] FIG. 7 shows a flow chart detailing a process 700 for automatically tuning a transmitter in a wireless circuit. Process 700 can be carried out by applying a

10 varying voltage (or current) source to a transmitter at 702. At 704, a reactive element coupled to the transmitter is used to adjust the transmitter in resonance in accordance with the applied voltage. At 706, a first transmitter real power is measured at a first frequency. At 708, the frequency is incremented and at 710 a second transmitter real power is measured at the incremented frequency. At 712, the second transmitter real

15 power at the incremented frequency is compared to the first transmitter real power at the first frequency. If at 714 the comparison indicates that the second real power is less than the first real power by an amount greater than a pre-determined amount, then the frequency of the voltage source is set to the incremented frequency at 716, otherwise control is passed back to 708. It should be noted that I_{rms} and V_{rms} can also

20 be used to analogize the change in real power as a function of transmitter frequency in accordance with Eq. (4).

[0046] FIG. 8 shows representative virtual charging area 800 in accordance with the described embodiments. Virtual charging area 800 provides region R of charging for suitably configured devices placed within the region R. NFMR power

25 supply can be placed in central unit such as desktop computer. In this way, the desktop computer can provide the NFMR power supply with computing resources. It should be noted that the near field magnetic resonance (NFMR) power supply can include high Q circuit that relies upon near field magnetic coupling by way of a resonance channel formed between resonances of the power source and sink to

30 transfer power. The NFMR power supply can be a standalone unit such as, for example, included in a desk top computer, laptop computer, tablet computer, and so on. In other embodiments, the NFMR power supply can take the form of a portable type unit such as a dongle that can be connected to a legacy device such as a desktop

computer thereby providing the ability to retrofit devices. In still other embodiments, housing or a portion of a housing used to enclose the NFMR power source can act to extend a useful range of the NFMR power supply.

[0047] As shown in Fig. 8 virtual charging area 800 includes central unit 802

5 (desktop computer) that can include the NFMR power supply, keyboard 804, mouse 806, and portable media player 808. In one embodiment, keyboard 804 can be configured to receive power directly from the NFMR power supply included in desktop computer 802 as can mouse 806 and portable media player 808 (when located within range R).

10 **[0048]** In some cases, the ability of desktop computer 802 to provide power directly to mouse 806, for example, can be reduced due to any number of factors. Such factors can include, for example, the addition of other devices into region R that require power from the NFMR power supply, obstacles interfering with the direct power channel formed between the NFMR and mouse 806, and so on. In this case,
15 keyboard 804 can act as a re-resonator such that a portion of the power delivered to keyboard 804 from the NFMR power supply can be passed on by way of a re-resonator transmission unit (not shown) in keyboard 804. In this way, any power loss experienced by mouse 806 can be ameliorated by the power received from keyboard 804. This arrangement can be transitory or can last for as long as mouse 806 is not
20 able to receive adequate power directly from the NFMR power supply. In other cases, the locating of portable media player 808 within region R can reduce the amount of power available to keyboard 804 and mouse 806. In this case, if a battery in keyboard 806 is fully charged (or additional charge is not necessary) then keyboard 806 can decouple a charging circuit while still maintaining a re-resonator circuit providing
25 power to mouse 806.

[0049] In some embodiments, dongle 810 can be connected to desktop computer 802 (by way of a USB port or cable, for example). So connected, dongle 810 can, in turn, act as a range extender for the NFMR power supply. In this way, dongle 810 can extend a range that power can be provided by the NFMR power
30 supply included in desktop computer 802. In some cases, dongle 810 can re-resonate power already received from the NFMR power supply while in other cases, dongle 810 can include its own NFMR power supply. By having its own NFMR power supply, dongle 810 can provide additional power wirelessly to those devices within

virtual charging region 800 separate from the power provided by the NFMR power supply included in desktop 802. It should be noted that in some embodiments, the housing of desktop computer 802 (or a portion thereof) can be used as a resonator as part of the NFMR power supply.

5 **[0050]** FIG. 9 is a block diagram of computing system 900 in accordance with the described embodiments. Computing system 900 includes processor 902 that pertains to a microprocessor or controller for controlling the overall operation of computing system 900. Computing system 900 stores data pertaining to media items, for example, in a file system 904 and a cache 906. The file system 904 is, typically, a
10 storage disk or a plurality of disks. The file system typically provides high capacity storage capability for computing system 900. However, since the access time to the file system 904 is relatively slow, computing system 900 also includes a cache 906. The cache 906 is, for example, Random-Access Memory (RAM) provided by semiconductor memory. The relative access time to the cache 906 is substantially
15 shorter than for the file system 904. However, the cache 906 does not have the large storage capacity of the file system 904.

[0051] Further, the file system 904, when active, consumes more power than does the cache 906. The power consumption is particularly important when the computing system 900 is a portable media player that is powered by a battery (not
20 shown).

[0052] Computing system 900 also includes a user input device 908 that allows a user of computing system 900 to interact with computing system 900. For example, the user input device 908 can take a variety of forms, such as a button, keypad, dial, etc. Still further, the computing system 900 includes data bus 910 can
25 facilitate data transfer between at least the file system 904, the cache 906, the processor 902, and the CODEC 912.

[0053] In one embodiment, computing system 900 serves to store a plurality of media items (e.g., songs) in the file system 904. When a user desires to have the media player play a particular media item, a list of available media items is displayed
30 on the display 910. Then, using the user input device 908, a user can select one of the available media items. The processor 502, upon receiving a selection of a particular media item, supplies the media data (e.g., audio file) for the particular media item to a coder/decoder (CODEC) 912.

[0054] The various aspects, embodiments, implementations or features of the described embodiments can be used separately or in any combination. Various aspects of the described embodiments can be implemented by software, hardware or a combination of hardware and software. The described embodiments can also be embodied as computer readable code on a computer readable medium for controlling manufacturing operations or as computer readable code on a computer readable medium for controlling a manufacturing line. The computer readable medium is any data storage device that can store data which can thereafter be read by a computer system. Examples of the computer readable medium include read-only memory, random-access memory, CD-ROMs, DVDs, magnetic tape, and optical data storage devices. The computer readable medium can also be distributed over network-coupled computer systems so that the computer readable code is stored and executed in a distributed fashion.

[0055] The foregoing description, for purposes of explanation, used specific nomenclature to provide a thorough understanding of the described embodiments. However, it will be apparent to one skilled in the art that the specific details are not required in order to practice the invention. Thus, the foregoing descriptions of specific embodiments are presented for purposes of illustration and description. They are not intended to be exhaustive or to limit the invention to the precise forms disclosed. It will be apparent to one of ordinary skill in the art that many modifications and variations are possible in view of the above teachings.

WHAT IS CLAIMED IS:

1. A wireless power transmission unit comprising:
 - a varying frequency excitation source arranged to provide a variable circuit excitation at frequency f ; and
 - 5 a first resonator structure coupled to the varying frequency f power supply, the first resonator structure having a resonant frequency ω_1 , a first Q factor Q_1 , and a characteristic size L_1 wherein non-radiative energy transfer between the first resonator structure and at a second resonator structure positioned apart from the transmission unit, the second resonator structure having a resonant frequency ω_2 , a second Q factor
 - 10 Q_2 , and a characteristic size L_2 is mediated through a magnetic resonant coupling associated with evanescent tails between the first resonator structure and the second resonator structure when the first resonator structure and the second resonator structure are in a tuned state when the first resonant frequency ω_1 and the second resonant frequency ω_2 are within the narrower of first resonance width Γ_1 or second
 - 15 resonance width Γ_2 such that the first resonant frequency ω_1 and the resonant frequency ω_2 are separated by a bandwidth of no more than 3db, wherein the first resonator structure maintains the first resonator structure and the second resonator structure in the tuned state by varying the first resonant frequency ω_1 such that the first resonant frequency ω_1 and the resonant frequency ω_2 are separated by the
 - 20 bandwidth of no more than 3db or are within the narrower of first resonance width Γ_1 or second resonance width Γ_2 in accordance with any changes in the second resonant frequency ω_2 .
2. The wireless power transmission unit as recited in claim 1, wherein the varying frequency excitation source is a varying voltage supply arranged to provide
- 25 frequency dependent voltage V_f .
3. The wireless power transmission unit as recited in claim 2, wherein the first resonator structure comprises:
 - a transmitter coil coupled to the varying frequency f power supply, wherein in the tuned state, the transmitter coil wirelessly provides real power P_{real} at the first
 - 30 resonant frequency ω_1 to the second resonator structure by magnetic resonant inductive coupling, the transmitter coil having a representative transmitter coil resistance R_{coil} .

4. The wireless power transmission unit as recited in claim 3, wherein the transmitter coil further comprising:
 - a variable impedance element coupled in series with the transmitter coil resistance R_{coil} .
5. The wireless power transmission unit as recited in claim 4, further comprising
 - an average current (I_{ave}) indicator connected in series to the varying voltage supply, the transmitter coil resistance R_{coil} and the variable impedance element, the I_{ave} indicator arranged to provide an indication of an average current I_{ave} in real time in the wireless power transmission unit;
 - a parasitic resistance R_p connected in series with the variable impedance element and the coil resistance R_{coil} ; and
 - a receiver load R_r corresponding to a receiver load representing a load on the wireless power transmission unit caused by the coupling of the first resonator structure and the second resonator structure
6. The wireless power transmission unit as recited in claim 5, wherein the wireless power transmission unit has an effective circuit impedance Z_{circuit} at resonance in accordance with $Z_{\text{circuit}} = R_s + R_{\text{coil}} + R_p + R_r$.
7. The wireless power transmission unit as recited in claim 6, further comprising:
 - a processor coupled with the varying frequency voltage supply and with the average current I_{ave} indicator arranged to calculate a real power (P_{real}) value corresponding to an amount of real power P_{real} wirelessly transmitted by the first resonator structure to the second resonator structure in real time in the tuned state.
8. The wireless power transmission unit as recited in claim 7, wherein the processor calculates the amount of real power P_{real} wirelessly transmitted by the first resonator structure to the second resonator structure in accordance with $P_{\text{Real}} = (I_{\text{ave}})^2 Z_{\text{circuit}}$.
9. The wireless power transmission unit as recited in claim 8, wherein the processor automatically determines second resonant frequency ω_2 by:
 - setting the frequency f of the variable voltage supply to a current frequency f_{current} ;
 - performing a frequency sweep by incrementing the current frequency f_{current} in a range of frequencies having a low frequency f_{low} and a high frequency f_{high} ;

- adjusting the variable impedance element to maintain the wireless power transmission unit in resonance at the current frequency f_{current} ;
- instructing the I_{ave} indicator to periodically measure the I_{ave} as a current I_{ave} at the current frequency f_{current} ;
- 5 calculating a current real power P_{real} at the current frequency f_{current} ;
- comparing the current real power P_{real} with a previously calculated real power P_{real} ;
- if the current real power P_{real} is less than the previously calculated real power P_{real} by at least a pre-determined amount, then setting the frequency f of the variable
- 10 voltage supply to the current frequency f_{current} , otherwise,
- adjusting the current frequency f_{current} within the range of frequency low f_{low} to frequency high f_{high} .
10. The wireless power transmission unit as recited in claim 9, comprising:
- a variable capacitance element having a variable capacitance value C , the
- 15 variable capacitance element coupled to the processor, wherein the processor maintains the wireless power transmission unit in resonance during the frequency sweep by adjusting the variable capacitance value C such that the overall impedance of the wireless power transmission unit is essentially zero.
11. The wireless power transmission unit as recited in claim 9, comprising:
- 20 a variable inductance element having a variable inductance value L , the variable inductance element coupled to the processor, wherein the processor maintains the wireless power transmission unit in resonance during the frequency sweep by adjusting the variable inductance value L such that the overall impedance of the wireless power transmission unit is essentially zero.
- 25 12. A method performed by a wireless power transmission unit for automatically tuning a wireless power transmission system, comprising:
- wherein the wireless power transmission unit includes at least a first resonator structure coupled to a varying frequency f voltage supply,
- in a tuned state, establishing a magnetic resonant inductive coupling between
- 30 the first resonator structure to a second resonator structure positioned apart from the wireless power transmission unit;
- non-radiatively transferring useable power from the first resonator structure to the second resonator structure using the magnetic resonant inductive coupling

between the first resonator structure and the second resonator structure, wherein in the tuned state, a first resonant frequency ω_1 of the first resonator structure and the second resonant frequency ω_2 of the second resonator structure are separated by a bandwidth of no more than 3db;

5 wherein the first resonator structure maintains the first resonator structure and the second resonator structure in the tuned state by varying a first resonant frequency ω_1 of the first resonant structure in order to maintain the first resonant frequency ω_1 and a second resonant frequency ω_2 are separated by the bandwidth of no more than 3db in accordance with any change in the second resonant frequency ω_2 .

10 13. The method as recited in claim 12, wherein the varying frequency excitation source is a varying voltage supply arranged to provide frequency dependent voltage V_f .

14. The method as recited in claim 13, wherein the first resonator structure comprises:

15 a transmitter coil coupled to the varying frequency f power supply, wherein in the tuned state, the transmitter coil wirelessly provides real power P_{real} at the first resonant frequency ω_1 to the second resonator structure by magnetic resonant inductive coupling, the transmitter coil having a representative transmitter coil resistance R_{coil} .

20 15. The method as recited in claim 14, the transmitter coil further comprising:
a variable impedance element coupled in series with the transmitter coil resistance R_{coil} .

16. The method as recited in claim 15, the wireless power transmission unit further comprising:

25 an average current (I_{ave}) indicator connected in series to the varying voltage supply, the transmitter coil resistance R_{coil} and the variable impedance element, the I_{ave} indicator arranged to provide an indication of an average current I_{ave} in real time in the wireless power transmission unit;

30 a parasitic resistance R_p connected in series with the variable impedance element and the coil resistance R_{coil} ; and

 a receiver load R_r corresponding to a receiver load representing a load on the wireless power transmission unit caused by the coupling of the first resonator structure and the second resonator structure

17. The method as recited in claim 16, wherein the wireless power transmission unit has an effective circuit impedance Z_{circuit} at resonance in accordance with $Z_{\text{circuit}} = R_s + R_{\text{coil}} + R_p + R_r$.
18. The method as recited in claim 17, wireless power transmission unit further comprising:
- 5 a processor coupled with the varying frequency voltage supply and with the average current I_{ave} indicator arranged to calculate a real power (P_{real}) value corresponding to an amount of real power P_{real} wirelessly transmitted by the first resonator structure to the second resonator structure in real time in the tuned state.
- 10 19. The method as recited in claim 18, wherein the processor calculates the amount of real power P_{real} wirelessly transmitted by the first resonator structure to the second resonator structure in accordance with $P_{\text{Real}} = (I_{\text{ave}})^2 Z_{\text{circuit}}$.
20. The method as recited in claim 19, wherein the processor automatically determines the second resonant frequency ω_2 by:
- 15 setting the frequency f of the variable voltage supply to a current frequency f_{current} ;
- performing a frequency sweep by incrementing the current frequency f_{current} in a range of frequencies having a low frequency f_{low} and a high frequency f_{high} ;
- adjusting the variable impedance element to maintain the wireless power transmission unit in resonance at the current frequency f_{current} ;
- 20 instructing the I_{ave} indicator to periodically measure the I_{ave} as a current I_{ave} at the current frequency f_{current} ;
- calculating a current real power P_{real} at the current frequency f_{current} ;
- comparing the current real power P_{real} with a previously calculated real power P_{real} ;
- 25 if the current real power P_{real} is less than the previously calculated real power P_{real} by at least a pre-determined amount, then setting the frequency f of the variable voltage supply to the current frequency f_{current} , otherwise,
- adjusting the current frequency f_{current} within the range of frequency low f_{low} to frequency high f_{high} .
- 30 21. The method as recited in claim 20, wireless power transmission unit further comprising:

a variable capacitance element having a variable capacitance value C , the variable capacitance element coupled to the processor, wherein the processor maintains the wireless power transmission unit in resonance during the frequency sweep by adjusting the variable capacitance value C such that the overall impedance of the wireless power transmission unit is essentially zero.

22. The method as recited in claim 20, wireless power transmission unit further comprising:

a variable inductance element having a variable inductance value L , the variable inductance element coupled to the processor, wherein the processor maintains the wireless power transmission unit in resonance during the frequency sweep by adjusting the variable inductance value L such that the overall impedance of the wireless power transmission unit is essentially zero.

23. Non-transitory computer readable medium for storing a computer program executable by a processor in a wireless power transmission unit for automatically tuning a wireless power transmission system, the wireless power transmission unit includes at least a first resonator structure coupled to a varying frequency f voltage supply, the computer readable medium comprising:

computer code for establishing a magnetic resonant inductive coupling between the first resonator structure to a second resonator structure positioned apart from the wireless power transmission unit in a tuned state;

computer code for non-radiatively transferring useable power from the first resonator structure to the second resonator structure using the magnetic resonant inductive coupling between the first resonator structure and the second resonator structure, wherein in the tuned state, a first resonant frequency ω_1 of the first resonator structure and the second resonant frequency ω_2 of the second resonator structure are separated by a bandwidth of no more than 3db;

computer code for maintaining the first resonator structure and the second resonator structure in the tuned state by first resonator structure by varying a first resonant frequency ω_1 of the first resonant structure in order to maintain the first resonant frequency ω_1 and a second resonant frequency ω_2 are separated by the bandwidth of no more than 3db in accordance with any change in the second resonant frequency ω_2 .

24. The computer readable medium as recited in claim 23, wherein the varying frequency excitation source is a varying voltage supply arranged to provide frequency dependent voltage V_f .
25. The computer readable medium as recited in claim 24, wherein the first resonator structure comprises:
- a transmitter coil coupled to the varying frequency f power supply, wherein in the tuned state, the transmitter coil wirelessly provides real power P_{real} at the first resonant frequency ω_1 to the second resonator structure by magnetic resonant inductive coupling, the transmitter coil having a representative transmitter coil resistance R_{coil} .
26. The computer readable medium as recited in claim 25, the transmitter coil further comprising:
- a variable impedance element coupled in series with the transmitter coil resistance R_{coil} .
27. The computer readable medium as recited in claim 26, the wireless power transmission unit further comprising:
- an average current (I_{ave}) indicator connected in series to the varying voltage supply, the transmitter coil resistance R_{coil} and the variable impedance element, the I_{ave} indicator arranged to provide an indication of an average current I_{ave} in real time in the wireless power transmission unit;
- a parasitic resistance R_p connected in series with the variable impedance element and the coil resistance R_{coil} ; and
- a receiver load R_r corresponding to a receiver load representing a load on the wireless power transmission unit caused by the coupling of the first resonator structure and the second resonator structure
28. The computer readable medium as recited in claim 27, wherein the wireless power transmission unit has an effective circuit impedance Z_{circuit} at resonance in accordance with $Z_{\text{circuit}} = R_s + R_{\text{coil}} + R_p + R_r$.
29. The computer readable medium as recited in claim 28, wireless power transmission unit further comprising:
- a processor coupled with the varying frequency voltage supply and with the average current I_{ave} indicator arranged to calculate a real power (P_{real}) value

corresponding to an amount of real power P_{real} wirelessly transmitted by the first resonator structure to the second resonator structure in real time in the tuned state.

30. The computer readable medium as recited in claim 29, wherein the processor calculates the amount of real power P_{real} wirelessly transmitted by the first resonator structure to the second resonator structure in accordance with $P_{\text{Real}} = (I_{\text{ave}})^2 Z_{\text{circuit}}$.

31. The computer readable medium as recited in claim 30, further comprising:
computer code for automatically determining the second resonant frequency ω_2 comprising:

computer code for setting the frequency f of the variable voltage supply to a current frequency f_{current} ,

computer code for performing a frequency sweep by incrementing the current frequency f_{current} in a range of frequencies having a low frequency f_{low} and a high frequency f_{high} ,

computer code for adjusting the variable impedance element to maintain the wireless power transmission unit in resonance at the current frequency f_{current} ,

computer code for instructing the I_{ave} indicator to periodically measure the I_{ave} as a current I_{ave} at the current frequency f_{current} ,

computer code for calculating a current real power P_{real} at the current frequency f_{current} ,

comparing the current real power P_{real} with a previously calculated real power P_{real} ,

computer code for setting the frequency f of the variable voltage supply to the current frequency f_{current} if the current real power P_{real} is less than the previously calculated real power P_{real} by at least a pre-determined amount, then, otherwise,

computer code for adjusting the current frequency f_{current} within the range of frequency low f_{low} to frequency high f_{high} .

32. The computer readable medium as recited in claim 31, wireless power transmission unit further comprising:

a variable capacitance element having a variable capacitance value C , the variable capacitance element coupled to the processor, wherein the processor maintains the wireless power transmission unit in resonance during the frequency sweep by adjusting the variable capacitance value C such that the overall impedance of the wireless power transmission unit is essentially zero.

33. The computer readable medium as recited in claim 32, wireless power transmission unit further comprising:
- a variable inductance element having a variable inductance value L , the variable inductance element coupled to the processor, wherein the processor
- 5 maintains the wireless power transmission unit in resonance during the frequency sweep by adjusting the variable inductance value L such that the overall impedance of the wireless power transmission unit is essentially zero.
34. A wirelessly powered local computing environment, comprising:
- a variable frequency power supply arranged to provide power at a frequency;
- 10 an resonance frequency adjustable near field magnetic resonance (NFMR) wireless transmitter coupled to the power supply and arranged to create a resonance channel used to transfer useable energy from the power supply to a receiver within a near field distance d ; and
- a central processing unit, the central processing unit providing processing
- 15 resources to the variable frequency power supply and the resonance frequency adjustable NFMR transmitter, wherein when the NFMR transmitter establishes a resonance coupling with the receiver, the central processing unit maintains at least a minimum system Q factor Q_{system} by adjusting only the resonance frequency of the NFMR transmitter.
- 20 35. The wirelessly powered local computing environment as recited in claim 34, further comprising:
- a power detector connected in series with the variable frequency power supply arranged to monitor a current in the NFMR wireless transmitter and provide a corresponding current value.
- 25 36. The wirelessly powered local computing environment claim 35, wherein during a frequency sweep operation, the central processing unit adjusts the frequency of the variable frequency power supply from a low frequency value to a high frequency value and the instructs the power detector to concurrently monitor the power in the NFMR transmitter.
- 30 37. The wirelessly powered local computing environment as recited in claim 36, further comprising:
- an adjustable reactive element connected in series with variable frequency power supply forming an RLC equivalent circuit corresponding to the NFMR

transmitter, wherein during the frequency sweep operation the central processing unit adjusts the adjustable reactive element in order to maintain the NFMR transmitter in resonance.

38. The wirelessly powered local computing environment as recited in claim 37,
5 wherein the adjustable reactive element is an adjustable capacitor having an adjustable capacitance C or adjustable inductance L .

39. The wirelessly powered local computing environment as recited in claim 38,
wherein when the monitored power in the NFMR transmitter at a current frequency is less than an average power value by a pre-determined amount, then the frequency of
10 the adjustable power supply is set to the current frequency.

40. A method performed by a processor for auto-tuning a near field magnetic resonance (NFMR) transmitter arranged to create a resonance channel used to transfer useable energy from a variable frequency power supply to a receiver within a near field distance d of the transmitter, wherein the variable frequency power supply is
15 arranged to provide the power to the NFMR transmitter at a frequency, the method comprising:

providing power to the NFMR transmitter at the frequency by the variable frequency power supply;

maintaining the NFMR transmitter in resonance mode by adjusting the
20 reactive element;

monitoring a power generated in response to the provided power at the current frequency; and

setting the frequency of the variable frequency power supply to the current frequency when the monitored power is less than an average power by at least a pre-
25 determined amount.

41. The method as recited in claim 40, wherein the average current is a root mean square (RMS) value of the monitored current.

42. The method as recited in claim 41, wherein the frequency of the variable frequency power supply is changed during a frequency sweep operation during which
30 processor adjusts the frequency of the variable frequency power supply from a low frequency value to a high frequency value and the instructs the current detector to concurrently monitor the current.

43. The method as recited in claim 42, further comprising:
during the frequency sweep operation, adjusting an adjustable reactive element in order to maintain the NFMR transmitter in resonance, the adjustable reactive element connected in series with the variable frequency power supply forming an
- 5 RLC equivalent circuit corresponding to the NFMR transmitter.
44. The method as recited in claim 43, wherein the adjustable reactive element is an adjustable capacitor having an adjustable capacitance C .
45. The method as recited in claim 44, wherein when the monitored current in the NFMR transmitter at a current frequency is less than the monitored current at another
- 10 frequency by more than a pre-determined amount, then the frequency of the adjustable power supply is set to the current frequency.
46. A non-transitory computer readable medium for storing computer code executed by a processor for auto-tuning a near field magnetic resonance (NFMR) transmitter arranged to create a resonance channel used to transfer useable energy
- 15 from a variable frequency power supply to a receiver within a near field distance d of the transmitter, wherein the variable frequency power supply is arranged to provide the power to the NFMR transmitter at a frequency, the method comprising:
- computer code for providing power to the NFMR transmitter at the frequency by the variable frequency power supply;
- 20 computer code for maintaining the NFMR transmitter in resonance mode;
computer code for changing the frequency of the variable frequency power supply to a current frequency;
computer code for monitoring a current generated in response to the provided power at the current frequency; and
- 25 computer code for setting the frequency of the variable frequency power supply to the current frequency when the monitored current is less than an average current by at least a pre-determined amount.
47. The computer readable medium as recited in claim 46, wherein the average current is a root mean square (RMS) value of the monitored current.
- 30 48. The computer readable medium as recited in claim 47, wherein the frequency of the variable frequency power supply is changed during a frequency sweep operation during which processor adjusts the frequency of the variable frequency

power supply from a low frequency value to a high frequency value and the instructs the current detector to concurrently monitor the current.

49. The computer readable medium as recited in claim 48, further comprising:
computer code for adjusting an adjustable reactive element in order to
- 5 maintain the NFMR transmitter in resonance, the adjustable reactive element connected in series with the variable frequency power supply forming an RLC equivalent circuit corresponding to the NFMR transmitter during the frequency sweep.
50. The computer readable medium as recited in claim 49, wherein the adjustable
- 10 reactive element is an adjustable capacitor having an adjustable capacitance C.
51. The computer readable medium as recited in claim 50, wherein when the monitored current in the NFMR transmitter at a current frequency is less than the monitored current at another frequency by more than a pre-determined amount, then the frequency of the adjustable power supply is set to the current frequency.
- 15 52. The computer readable medium as recited in claim 51, wherein the average current value is a root mean square value (RMS).

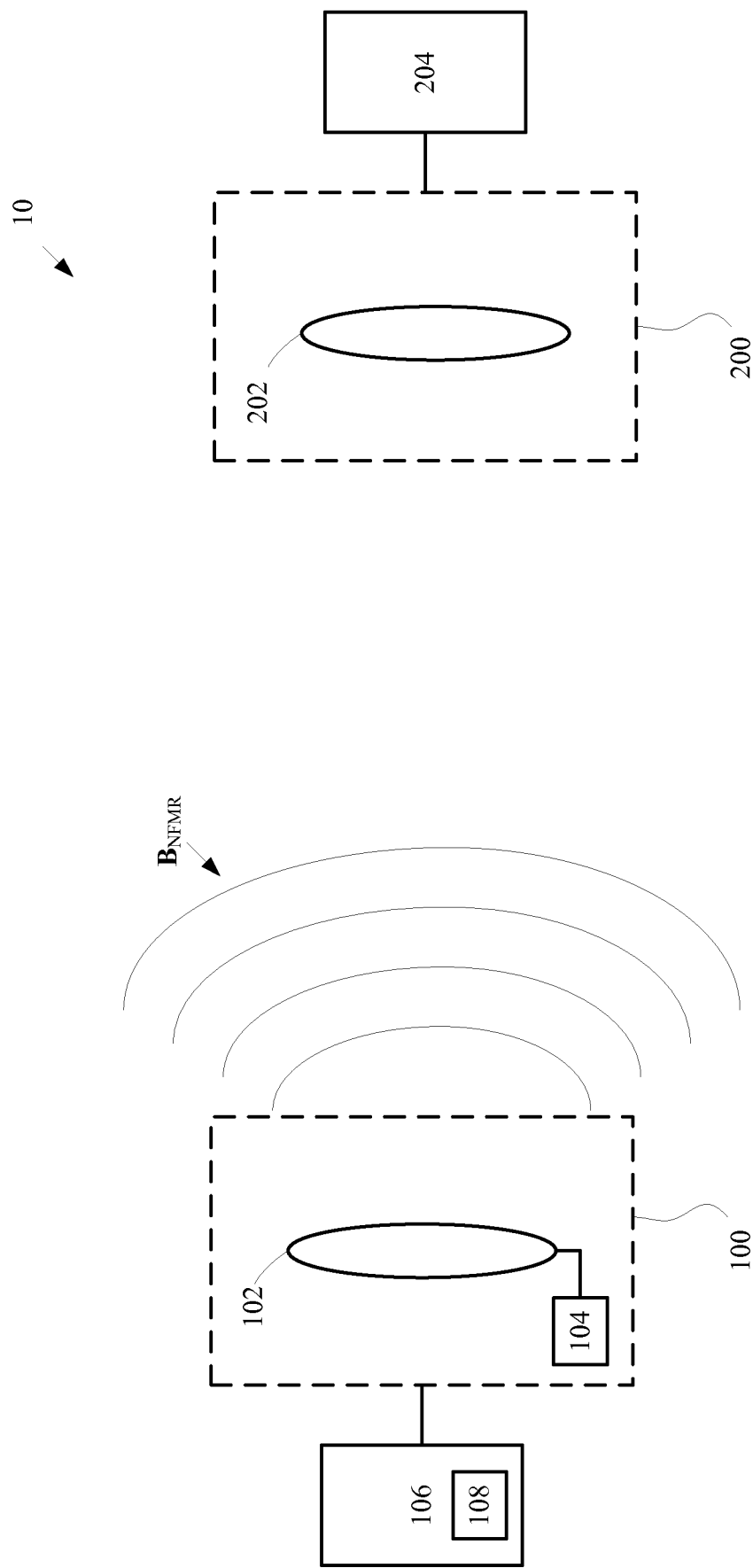


FIG. 1

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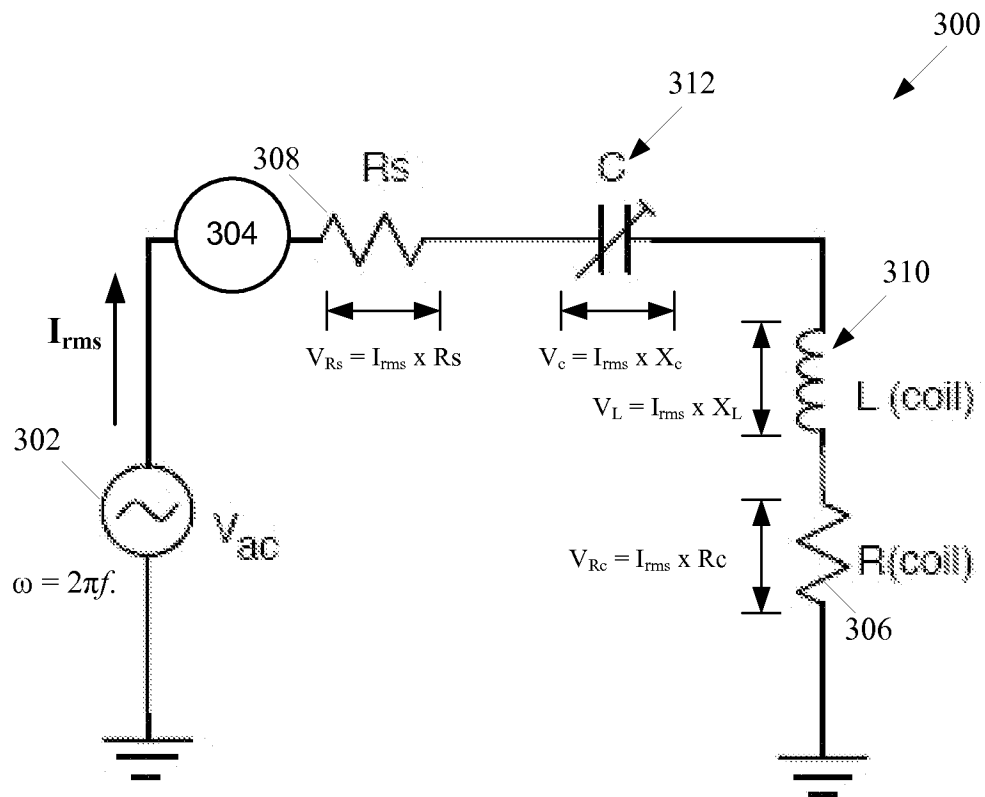


FIG. 2A

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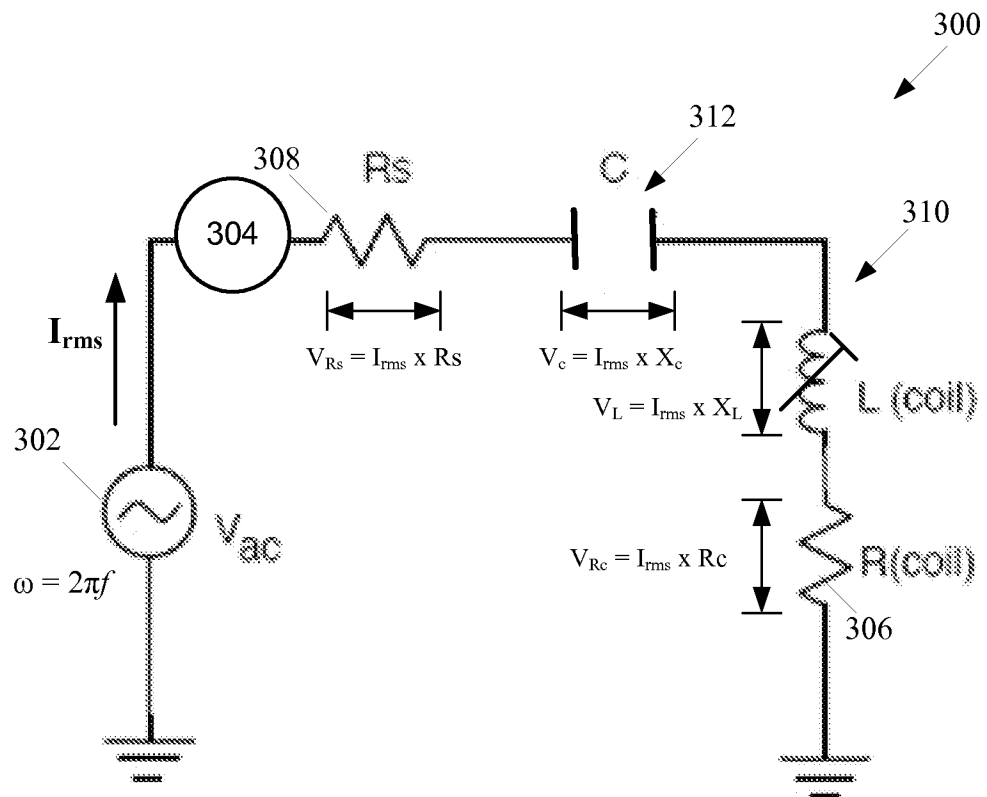


FIG. 2B

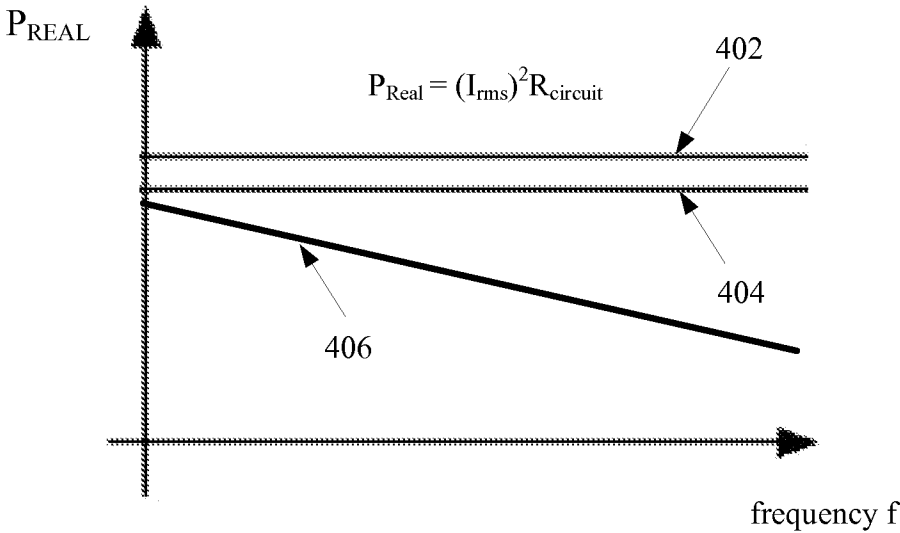


FIG. 3

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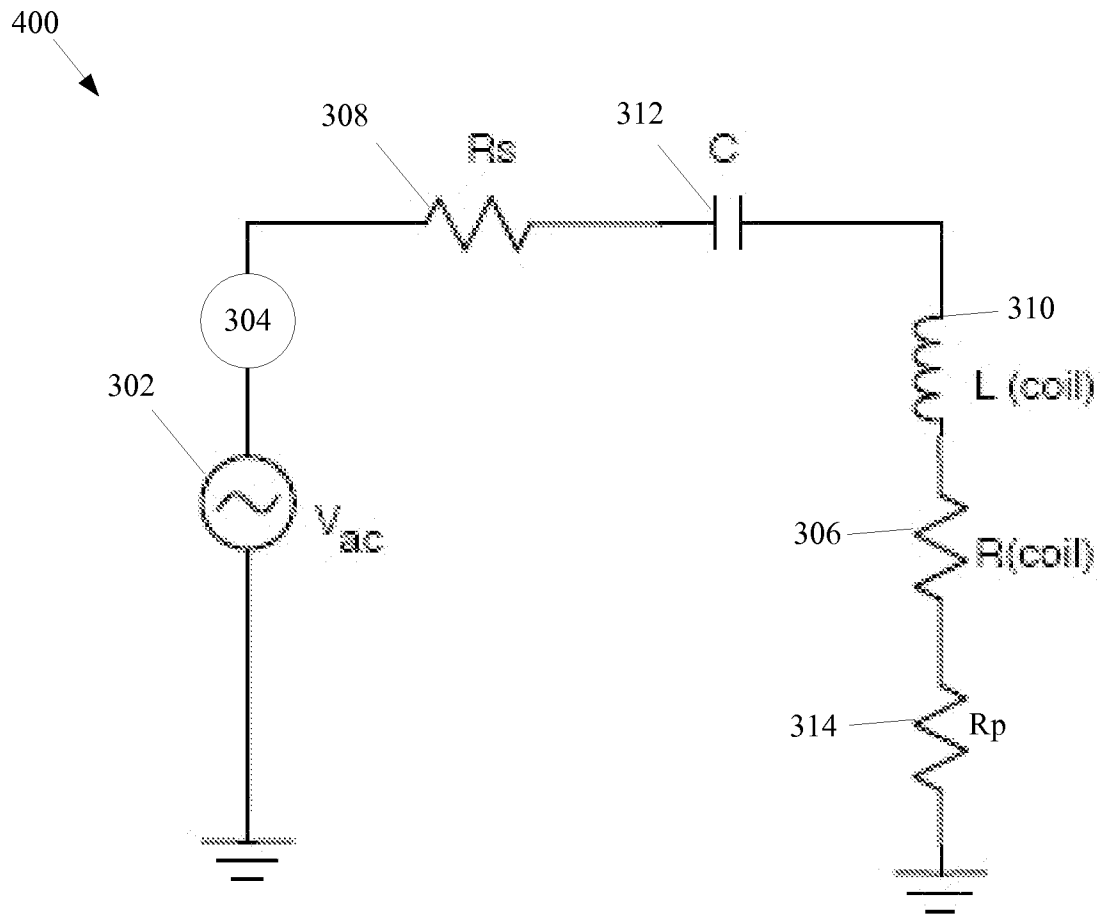


FIG. 4

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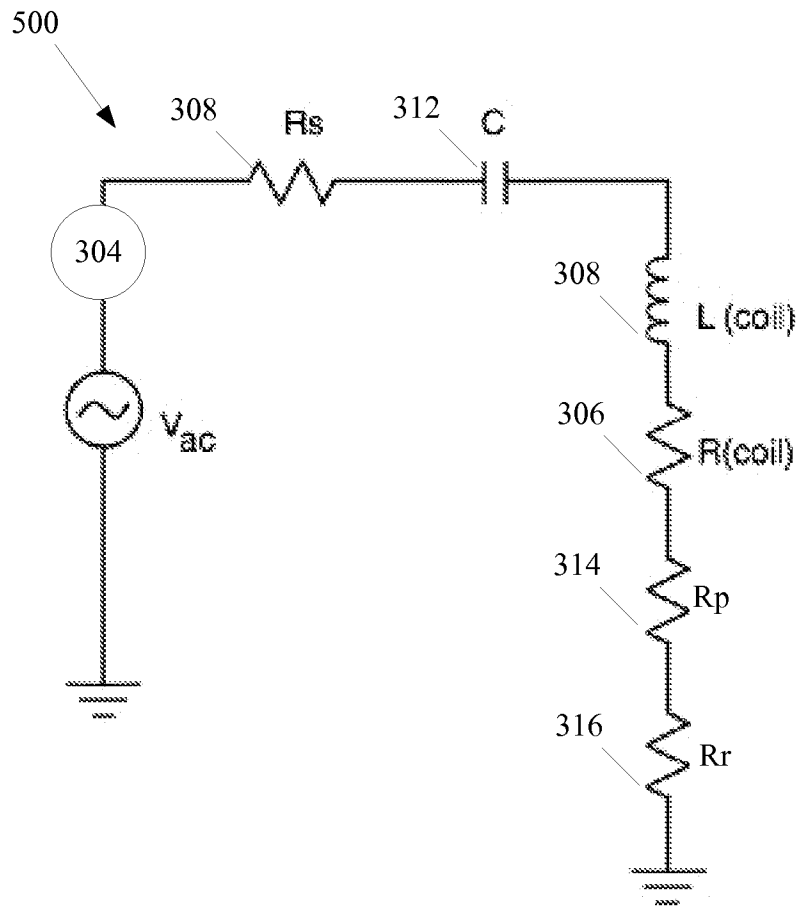


FIG. 5

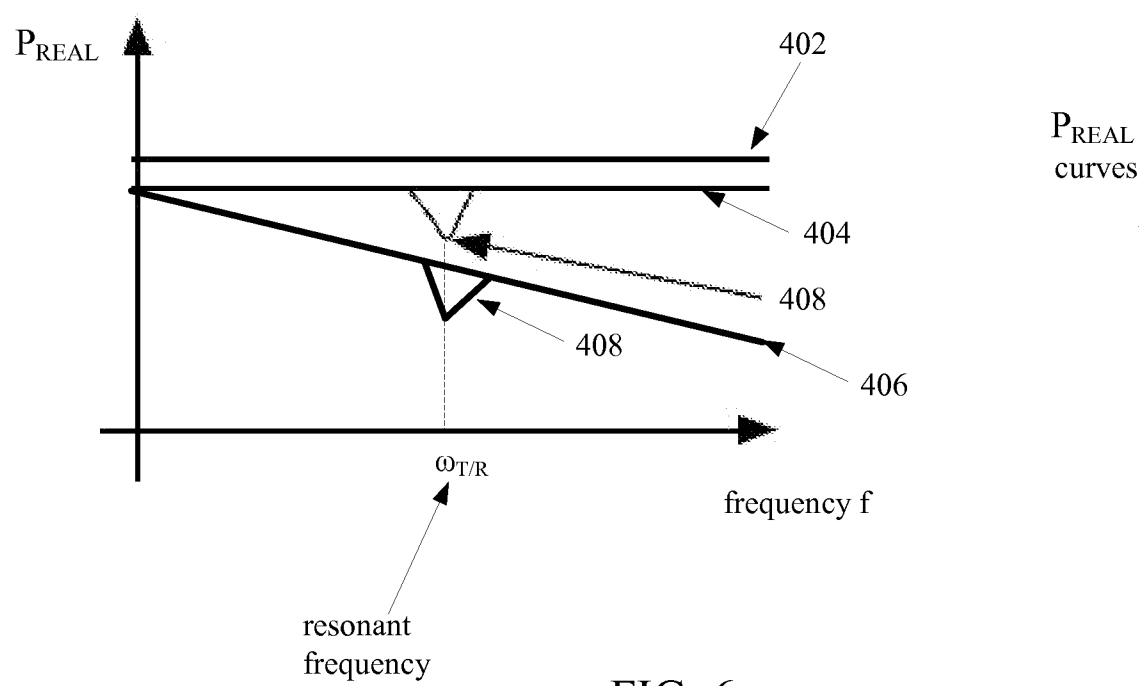


FIG. 6

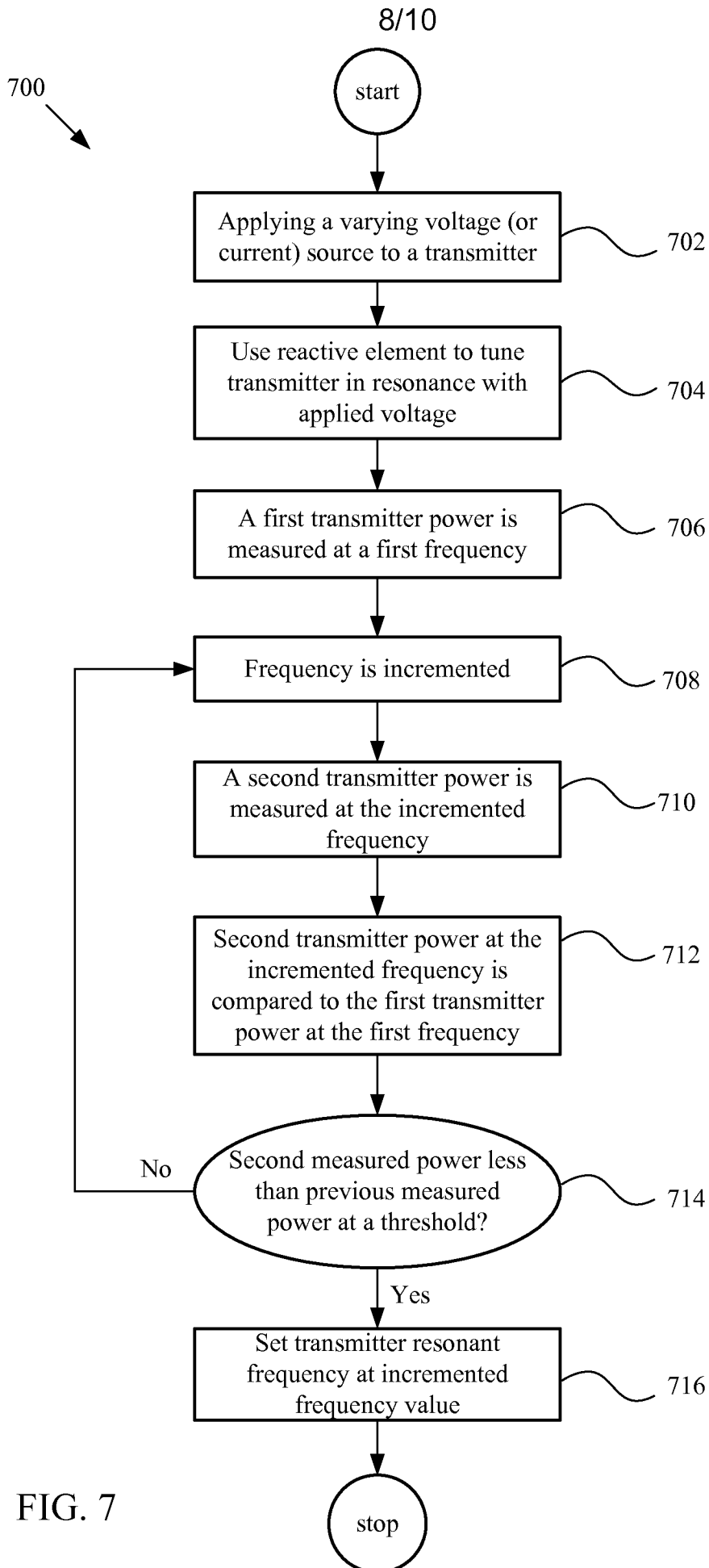


FIG. 7

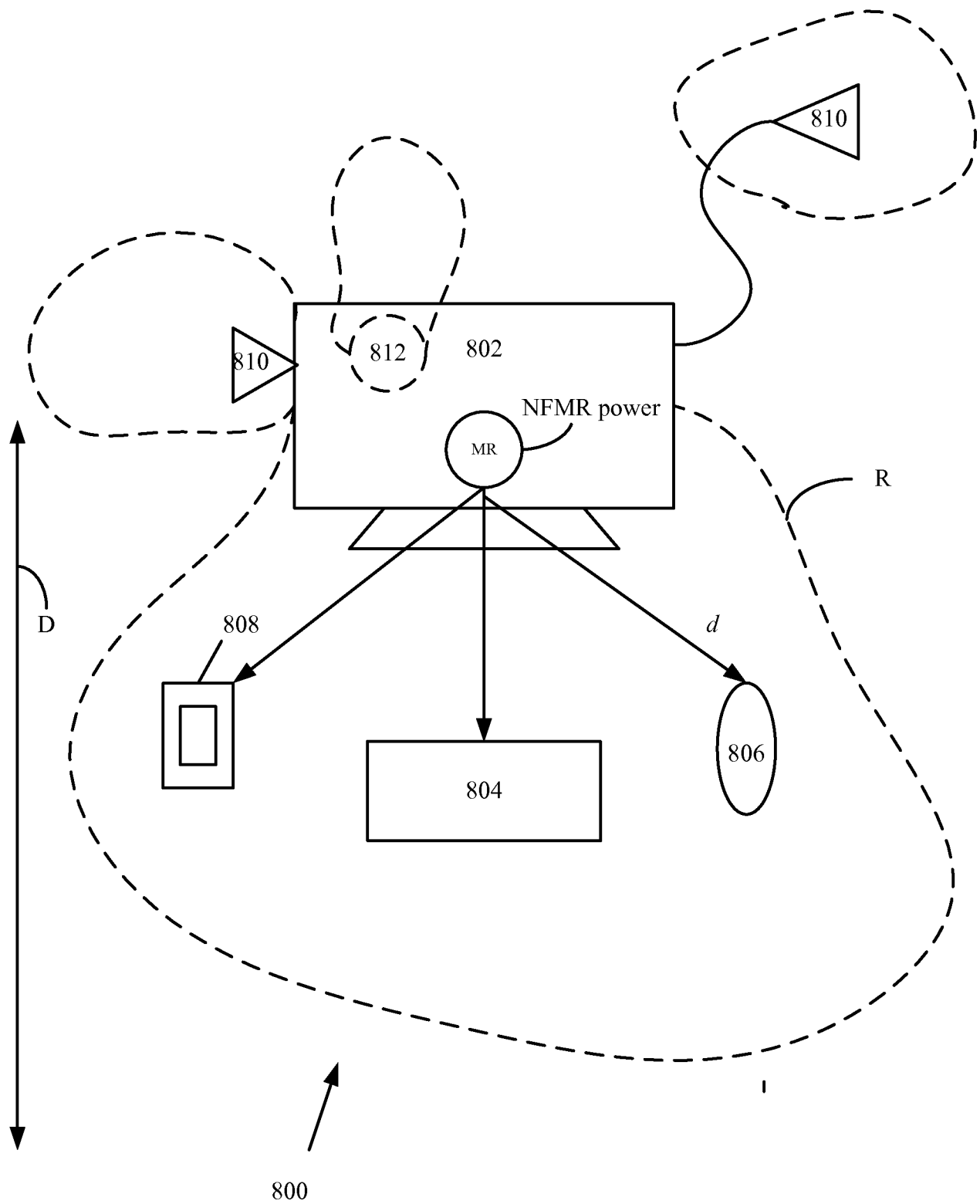


FIG. 8

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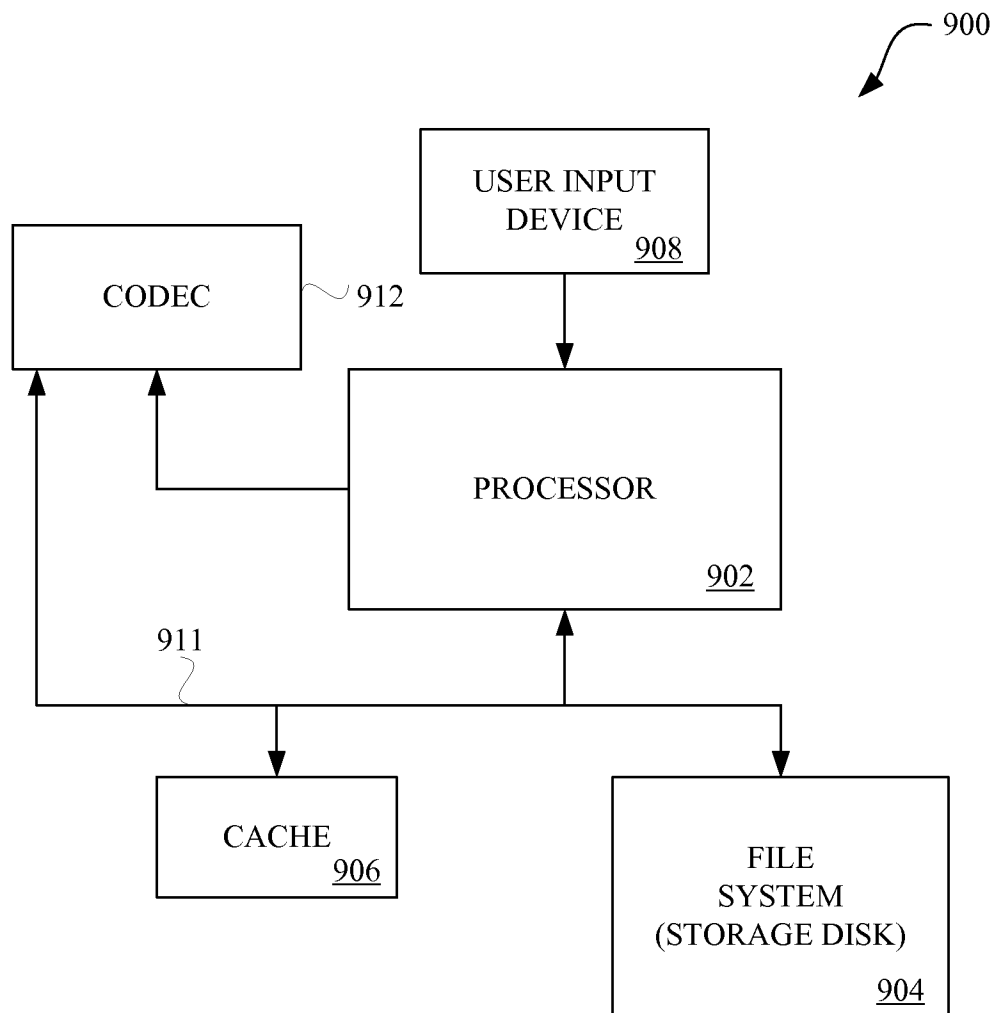


FIG. 9

A. CLASSIFICATION OF SUBJECT MATTER***H02J 17/00(2006.01)i, H02J 7/00(2006.01)i, H04B 5/02(2006.01)i***

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

H02J 17/00; H02M 3/335; G06F 1/26; G08C 19/00

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Korean utility models and applications for utility models

Japanese utility models and applications for utility models

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

eKOMPASS(KIPO internal) & Keywords: resonator, de-coupling, transmitter, NFMR, coupling, energy transfer, receiver

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 2009-0271048 A1 (WAKAMATSU, M.) 29 October 2009 See abstract, claims 1-12, figures 1-17.	1-52
A	US 2009-0134712 A1 (COOK, N. P. et al.) 28 May 2009 See abstract, claims 1-53, figures 1-12.	1-52
A	US 2009-0058189 A1 (COOK, N. P. et al.) 05 March 2009 See abstract, claims 1-30, figures 1-3.	1-52
A	US 2003-0214821 A1 (GIANNOPOULOS, D. et al.) 20 November 2003 See abstract, claims 1-24, figures 1-5.	1-52
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Further documents are listed in the continuation of Box C.



See patent family annex.

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Date of the actual completion of the international search

30 DECEMBER 2011 (30.12.2011)

Date of mailing of the international search report

02 JANUARY 2012 (02.01.2012)

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Information on patent family members

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