

INDUCTIVE POWER TRANSFER SYSTEM

[0001] This invention relates to an inductive power transfer system. In general terms this application relates to maximising DC to load efficiency for inductive power transfer.

BACKGROUND

5 [0002] Inductive power transfer (IPT) without a magnetic core was first proposed by Nikola Tesla to supply wireless mains power over long distances around 100 years ago [1]. Since then, low-power, closely-coupled wireless charging methods have been used to power medical implants [2], while the wireless powering of portable devices through charging mats is now available via commercial products [3]. Nonetheless, there has been
10 recent interest in wireless power transfer (WPT) for medium range (*i.e.* 10s of cm) applications, such as electric vehicle charging through resonant inductive coupling [4]–[7].

[0003] For many industrial and commercial applications, IPT systems must be capable of achieving a high end-to-end efficiency η_{ee} , while transferring hundreds of watts at sub-metre distances, otherwise they will not be adopted. Several approaches for achieving
15 good link efficiencies have been developed by several research groups. The first is to work at relatively low frequencies (tens of kHz), where efficient driver circuits can be easily realised and by increasing the coupling factor k of the system, using field-shaping techniques; for example, by employing metamaterials [8] and ferrite cores [6]. In [6], 2 kW of power was transferred at a distance of 10 cm using Litz wire coils at 20 kHz. The
20 operating frequency was defined by the power handling capabilities of the coil driver, limiting the maximum coil unloaded Q-factor to 290. Field-shaping techniques normally occupy useful volume, require heavy materials, employ expensive fabrication techniques and need a precise coil alignment. These solutions make the field-shaping approach unsuitable for many applications, where the size, weight and cost of the system are limiting
25 factors.

[0004] The second approach relies on transferring energy at the optimum frequency for maximum power transfer given a particular coil size, where the unloaded Q is maximised and compensates for the low coupling factor. In the past, this approach was not considered efficient, since low driver efficiency (due to semiconductor losses) dramatically
30 reduced the end-to-end efficiency of the IPT system. An example of this was described by Kurs *et al.* [9], where the use of a 9.9 MHz Colpitts oscillator driver achieved an end-to-end efficiency of only 15%, when the transfer efficiency was 50%.

[0005] Other attempts at this approach have been successful, with the use of commercially-off-the-shelf (COTS) equipment to drive and impedance match the TX coils

at frequencies above 3 MHz and with $\eta_{\text{transfer}} = 95\%$, while also reducing the coil losses by using a surface spiral [10].

[0006] The highest η_{ee} have been demonstrated by the commercial IPT systems currently available on the market. High efficiencies of $\eta_{\text{ee}} = 90\%$ have been achieved at
5 distances of less than 30 cm but with relatively heavy systems (30-40 kg) that use field shaping ferromagnetic materials. In contrast, a system with frequency tracking and no ferromagnetic materials was used in [11], where an estimated $\eta_{\text{dc-load}} = 70\%$ was calculated. Here, no clear description of the driver's efficiency is given, as it is based on a COTS 50 Ω system with added TX and RX loops. Emphasis was again given to the control
10 of the link and transfer efficiency, rather than the dc-to-load efficiency. Other interesting attempts to increase the end-to-end efficiency have been presented in [12], [13], where $\eta_{\text{ee}} > 60\%$ have been achieved at close proximity.

[0007] The present invention, at least in the presently preferred embodiments, seeks to provide a high frequency, cost effective and efficient solution for mid-range IPT in the
15 absence of field-shaping techniques, allowing a light-weight system to be achieved. A system with a TX-RX coil size difference represents a more realistic system, where the receiver size is usually constrained by its application. This system should be able to achieve high efficiency for lower coupling factors, due to the smaller RX coil size. Furthermore, this system should be able to achieve high efficiencies even under situations
20 where perfect alignment is not always achievable (e.g. electric vehicle or wireless sensor charging).

BRIEF SUMMARY OF THE DISCLOSURE

[0008] In accordance with the present invention there is provided an inductive power transfer system comprising a transmitter coil and a receiver coil spaced from the
25 transmitter coil, a transmitter circuit comprising the transmitter coil and a receiver circuit comprising the receiver coil. The transmitter circuit is in the form of a Class E amplifier comprising a first inductor and a transistor in series between the terminals of a power supply. The transistor is arranged to switch at a first frequency. The transmitter circuit further comprises a first transmitter capacitor in parallel with the transistor between the first
30 inductor and a power supply terminal, a primary tank circuit in parallel with the first transmitter capacitor, the primary tank circuit comprising the transmitter coil and a second transmitter capacitor arranged in parallel or series with the transmitter coil, and a third transmitter capacitor in series with the first inductor between the first transmitter capacitor and the primary tank circuit. In the presently preferred embodiment, the capacitance of the
35 second transmitter capacitor is selected, such that the resonant frequency of the primary tank circuit is greater than the first frequency. In this way, the primary tank circuit operates

in semi-resonance so that the primary tank circuit retains some inductive impedance. The ratio of the first frequency to the resonant frequency of the primary tank circuit is typically in the range 0.5 to 0.9, preferably 0.7 to 0.9. The capacitance of the second transmitter capacitor may be selected, such that the resonant frequency of the primary tank circuit is less than the first frequency. For example, the ratio of the first frequency to the resonant frequency of the primary tank circuit may be in the range 1.1 to 1.5. In embodiments of the invention, the ratio of the first frequency to the resonant frequency of the primary tank circuit is in the range 0.5 to 1.5.

[0009] In a preferred arrangement, the receiver circuit has a resonant frequency and the transmitter circuit is configured to vary the first frequency, in order to achieve a desired impedance of the primary tank circuit.

[0010] Advantageously, the transmitter coil and/or the receiver coil has an air core. This provides a lightweight design. In embodiments of the invention the transmitter coil and/or the receiver coil has a diameter of at least 5cm, preferably at least 10cm. In embodiments of the invention the spacing between the transmitter coil and the receiver coil, in use, is at least 15cm.

[0011] The transistor is typically a MOSFET. The first frequency may be at least 80kHz, preferably at least 1 MHz. The power transferred between the transmitter coil and the receiver coil may be at least 1 watt, preferably at least 10 watts.

[0012] In one arrangement, the receiver circuit comprises a Class E rectifier. The receiver circuit may comprise a first receiver capacitor arranged in parallel with a load, in use, and a secondary tank circuit in parallel with the first receiver capacitor. The secondary tank circuit may comprise the receiver coil and a second receiver capacitor arranged in parallel or series with the receiver coil. A first diode may be provided between the secondary tank circuit and the first receiver capacitor. The capacitance of the second receiver capacitor may be selected such that the resonant frequency of the secondary tank circuit differs from the first frequency, whereby the secondary tank circuit operates in semi-resonance and maintains some reactive impedance. In this way, the necessary reactance for the Class E operation of the rectifier may be provided by the receiver coil alone. The ratio of the first frequency to the resonant frequency of the secondary tank circuit is typically in the range 0.2 to 3. For example, the ratio of the first frequency to the resonant frequency of the secondary tank circuit may be in the range 0.2 to 0.9 or 1.1 to 3. The only inductor in the receiver circuit may be the receiver coil.

[0013] The receiver circuit may comprise at least a second diode in parallel with the first diode. In this way, the junction capacitance of the diodes may provide the necessary capacitance for Class E operation without the limitation of the operating voltage of

additional capacitors. Thus, the only capacitance in parallel with the diode(s) may be provided by the junction capacitance of the diode(s). The diode(s) may be silicon carbide diodes, gallium nitride diodes or other wide band gap material.

[0014] The invention extends to the receiver circuit of the inductive power transfer system. The invention extends to the transmitter circuit of the inductive power transfer system.

BRIEF DESCRIPTION OF THE DRAWINGS

[0015] Embodiments of the invention are further described hereinafter with reference to the accompanying drawings, in which:

- 10 Figure 1 shows an inductive power transfer system architecture;
- Figure 2 shows an experimental set-up for distance and transverse offset measurements;
- Figure 3 shows an experimental set-up for angular misalignment measurements;
- Figure 4 illustrates measurements of coupling factor against coil separation
- 15 distance in air, with perfectly aligned coils;
- Figure 5 illustrates measurements of coupling factor against coil transverse offset in air, at a distance of 30 cm;
- Figure 6 illustrates measurements of coupling factor against transmitter coil angular misalignment, at a distance of 30 cm;
- 20 Figure 7 illustrates measurements of coupling factor against receiver coil angular misalignment, at a distance of 30 cm;
- Figure 8 shows a semi-resonant Class-E topology, with $\omega_d < \omega_{oTX}$ according to an embodiment of the invention;
- Figure 9 shows simulated C_{par} values against ω_d / ω_{oTX} for Class-E MOSFET
- 25 selection with a drain-source voltage of 230 V;
- Figure 10 shows the simulated drain-source voltage (PSpice) against time t for the semi-resonant Class-E driver;
- Figure 11 shows a perfectly aligned IPT test rig with $D = 30$ cm;
- Figure 12 shows the drain to source voltage against time for the IPT system with a
- 30 30 cm separation distance and $P_{dc} = 90$ W;

Figure 13 illustrates measurement of the dc-to-load efficiency against separation distance with fixed clock frequency tuning to aligned 30 cm separation distance scenario;

5 Figure 14 illustrates measurement of the dc-to-load efficiency against separation distance with clock frequency tuning

Figure 15 illustrates measurement of the dc-to-load efficiency against coil transverse offset with fixed clock frequency tuning to aligned 30 cm separation distance scenario;

10 Figure 16 illustrates measurement of the dc-to-load efficiency against coil transverse offset with only clock frequency tuning;

Figure 17 illustrates measurement of the dc-to-load efficiency against TX coil angle with fixed clock frequency tuning to aligned 30 cm separation distance scenario;

Figure 18 illustrates measurement of the dc-to-load efficiency against TX coil angular misalignment with clock frequency tuning;

15 Figure 19 illustrates measurement of the dc-to-load efficiency against RX coil angular misalignment with fixed clock frequency impedance tuning to aligned 30 cm separation distance scenario;

Figure 20 illustrates measurement of the dc-to-load efficiency against RX coil angular misalignment with clock frequency tuning;

20 Figure 21 is a schematic representation of Class E receiver circuit for use in an embodiment of the invention;

Figure 22 is a schematic representation of Class E receiver circuit according to a further embodiment of the invention;

25 Figure 23 is a schematic representation of Class E receiver circuit according to a yet further embodiment of the invention.

DETAILED DESCRIPTION

[0016] Inductive Power Transfer (IPT) systems for transmitting tens to hundreds of watts have been reported for almost a decade. Most of the work has concentrated on the optimization of the link efficiency and has not taken into account the efficiency of the driver. Class-E amplifiers have been identified as ideal drivers for IPT applications, but their power handling capability at tens of MHz has been a crucial limiting factor, since the load and inductor characteristics are set by the requirements of the resonant inductive system. The frequency limitation of the driver restricts the unloaded Q factor of the coils and thus

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the link efficiency. With a suitable driver, copper coil unloaded Q factors of over 1,000 can be achieved in the low MHz region, enabling a cost-effective high Q coil assembly. The system described herein alleviates the use of heavy and expensive field-shaping techniques by presenting an efficient IPT system capable of transmitting energy with a dc-to-load efficiency above 77% at 6 MHz across a distance of 30 cm. This is believed to be the highest dc-to-load efficiency achieved for an IPT system without introducing restrictive coupling factor enhancement techniques.

[0017] A basic IPT system architecture comprises several modules, as illustrated in Figure 1. The architecture includes DC power supply units (PSUs), coil driver (*i.e.* clock generator and power amplifier (PA) having an impedance matching network), transmitting (TX) coil with separation distance D from a receiving (RX) coil (measured from the centre-to-centre of the coils), an optional rectifier/regulator and a load. To fully characterize the complete system, the end-to end efficiency η_{ee} of all the building blocks, from the AC source to the load, can be considered as $\eta_{ee} = \eta_{dc-PSU} \eta_{dc-load}$, where the efficiency terms are shown in Figure 1. In Figure 1, $\eta_{dc-load} = \eta_{driver} \eta_{transfer}$, $\eta_{driver} = \eta_{clock} \eta_{amp}$, η_{dc-PSU} is the combined efficiency of the dc power supplies, $\eta_{dc-load}$ is the dc-to-load efficiency, η_{driver} is the efficiency of the driver, η_{link} is the link efficiency, $\eta_{transfer}$ is the transfer efficiency, η_{clock} is the efficiency of the driver clock, η_{amp} is the efficiency of the power amplifier, $\eta_{rectifier}$ is the rectifier efficiency and $\eta_{regulator}$ is the efficiency of the regulator. The present invention focuses on optimizing $\eta_{dc-load}$ for an IPT system without a rectifier or regulator, *i.e.* maximising: $\eta_{dc-load} = P_{load}/P_{dc}$ where P_{dc} is the total DC input power to the system (*i.e.* into the clock and power amplifier) and P_{load} is the real power dissipated in the load.

[0018] It is important that consistent, well-defined figures of merit, such as $\eta_{dc-load}$ and η_{ee} , are used to evaluate IPT systems to allow a straightforward comparison of the different emerging technologies in this field. From the IPT systems architecture in Figure 1, the transfer efficiency just describes part of the system's efficiency and does not take into account the driver.

[0019] The table below shows comparisons of the state of the art in IPT systems. In the table $\eta_{transfer}$, $\eta_{dc-load}$ and η_{ee} have been separated out, where possible, to highlight that dc-to-load efficiency can be substantially lower than the transfer efficiency.

D [cm]	f ₀ [kHz]	Driver Technology	Coil Technology	Magnetic Material	P _{load} [W]	η _{source-free} [%]	η _{dc-load} [%]	η _{tot} [%]	Ref.
6	134	Class-E	Litz wire	No	295	-	-	75.7	[12]
6	240	Class-E	Litz wire	No	3.7	71	-	66	[13]
10	20	H-bridge	Litz wire	Yes	2,000	85	-	-	[6]
10	-	-	-	Yes	3,300	-	-	90	[14]
15	6,700	HF transceiver	Loop + pancake coil	No	-	93	-	-	[15]
15	-	H-bridge	Litz wire	Yes	2,000	95	-	-	[16]
18	145	-	Litz wire	Yes	300-3,000	-	-	90	[17]-[19]
20	4,000	Class-E	Copper wire coil	No	2	-	50	-	[20]
20	20	H-bridge	Power lines	Yes	60,000	80	-	-	[21]
20	20	H-bridge	Power rail	Yes	27,000	-	-	74	[22]
30	3,700	HF transceiver	Surface spiral	No	220	95	-	-	[10]
30	6,000	Class-E	Copper pipe coils	No	95	-	77	-	[our work]
18-30, 40**	20	H-bridge	-	Yes	3,000	-	-	> 85	[23], [24]
70	7,650	Signal generator	Loop + pancake coil	No	30	75	-	-	[25]
50	13,560	Class-E	Loop + rectangle coil	No	70	85	70***	-	[11]
50	27,000	HF transceiver	Loop + spiral coil	No	40	47	-	-	[8]
100	500.5	Class-D	Litz wire	No	5-35	76	-	-	[26]
200	9,900	Colpitts oscillator	Litz wire	No	60	50	-	15	[9]

*Maximum power transfer distance stated in the cited website [24]

**Calculated based on estimated value of η_{source-free} [12]

[0020] In the following there is provided an overview of IPT theory, outlining systems

architecture and key component selection that define the system's end-to-end and dc-to-

load efficiencies. Cost efficient coil design, simulations and measurements to achieve dc-

to-load efficiencies above 70% for sub-metre distances are described. A framework for

driver modelling, component selection and layout considerations to achieve a low loss,

high frequency DC-RF conversion, capable of delivering more than 100 W at a distance of

30 cm is described. A full system characterization under different misalignment scenarios

is also described.

[0021] With the typical IPT system architecture shown in Figure 1, the driver provides

high frequency power to the TX coil, having an unloaded quality factor Q_{TX}, which couples

as defined by the coupling factor (or coefficient) *k* to the RX coil, having an unloaded

quality factor Q_{RX}. It is known that by using receiver (or secondary) resonance and

optimising the load impedance, the link efficiency can be maximized to give:

$$\eta_{link} = \frac{k^2 Q_{TX} Q_{RX}}{(1 + \sqrt{1 + k^2 Q_{TX} Q_{RX}})^2}$$

[0022] As can be seen from the above formula, the key to achieving high efficiency is to

maximise $k^2 Q_{TX} Q_{RX}$. The coil Q factor can be maximized by choosing the correct operating

frequency [27]. Analysis on the interactions of these key variables, using both closed form

mathematical expressions and more detailed numerical modelling in Matlab, has yielded

the following underlying principles for optimisation [28]:

- The loop radii should be maximized, in order to maximize the coupling factor;
- For a given constraint on loop dimensions, there is an optimal frequency, which is approximately the point at which the radiation resistance begins to be significant compared to the skin-effect resistance;

- The wire radius and the number of coil turns should be as large as possible (bearing in mind that the coils should remain electrically small, to limit the electric field and hence radiation); and
- In the case where the loops are not of equal size, the maximum operating frequency will be mainly determined by the larger of the two coils, also this dictates the lowest self-resonance frequency.

[0023] As presented in [12], four different configurations have been widely used for IPT systems. A series resonance can only be used if the parasitic shunt capacitance of the inductor is assumed to be negligible. In contrast, this assumption is not needed for the parallel case, since the parallel resonator capacitor can absorb the parasitic capacitance of the coil.

[0024] Furthermore, the coupled RX coil is always assumed to be operating at resonance; this way the equivalent optimal load on the transmitter, reflected from the receiver, will only be resistive, affecting only the damping of the transmitter tank [2].

[0025] To increase the efficiency of an IPT system, capable of transmitting tens to hundreds of watts at a distance of 30 cm, with perfectly aligned coils, simulations as described in [28] and measurements as described in [29], were undertaken for the TX and RX coils. With this technique, the Q-factor was measured through transmission coefficient measurements using two loosely inductive coupled coils as probes. For electromagnetic design reasons, the distance D from the centre-to-centre of the coils is used. However, it is important to note that the minimum distance between coils is $(D-7)$ cm in these results. As mentioned previously, a different sized TX and RX coil was used in the setup; this was thought to be more realistic for most scenarios. The coils were fabricated with copper piping having a 1 cm diameter and 1 mm wall thickness.

[0026] After characterizing the coils, the highest Q for both TX and RX coils is found close to 6 MHz, where skin depth is only 27 μm . The maximum unloaded Q value for the 5-turn, 20 cm diameter RX coil was $Q_{\text{RX}} = 1,100$ and $Q_{\text{TX}} = 1,270$ for the 3-turn, 30 cm TX coil; these matched simulation results when using the following standard expression for the unloaded Q-factor of a coil:

$$Q = \frac{\omega_d L}{R_{rad}(\omega_d) + R_{skin}(\omega_d)}$$

where

$$R_{rad}(\omega_d) = N^2 \eta_o \left(\frac{\pi}{6} \right) (\beta_o(\omega_d) r)^4$$

$$R_{skin}(\omega_d) \approx \frac{Nr}{2a} \sqrt{\frac{\omega_d \mu_o}{2\sigma_o}}$$

where ω_d is the driven angular frequency of operation, L is the self inductance of the coil, $R_{rad}(\omega_d)$ is the radiation resistance [30], N is the number of turns of the coil, η_o is the impedance of free space, r is the radius of the coil, $\beta_o(\omega_d) = 2\pi/\lambda_d$, λ_d is the free space wavelength at the driving clock frequency, a is radius of the copper pipe, σ_o is the low frequency conductivity of copper and μ_o is the permeability of free space. $R_{skin}(\omega_d)$ is an approximation of the skin-effect resistance but was calculated in simulations using Butterworth's numerical model [31], which also takes into account proximity effects.

[0027] A re-configurable test fixture was fabricated to hold the coils and allow for reproducible and easily adjustable operating scenarios, as will be described below. Perspex was used for both the stands and the coil spacers, to avoid the generation of eddy currents that could result in measurement errors. The coil spacers helped to maintain a fixed distance of 2 cm between windings, measured from the centres of the pipe, to reduce the proximity effect between turns. The test fixture allows the variation of D , transverse offset and angular misalignment between both coils.

[0028] Coupling factor measurements were undertaken to characterize the coil coupling in an array of different scenarios. Measurements were undertaken with different separation distances or against transverse coil offsets h , as illustrated in Figure 2; or transmitter or receiver coil angular misalignment, θ_{TX} or θ_{RX} , respectively, as illustrated in Figure 3. For experiments involving transverse offset or angular misalignment, the centre-to-centre distance was fixed at $D = 30$ cm, (the minimum distance between coils is 23 cm). Data from these measurements was also used to predict the operating characteristics, as well as the expected efficiency of the IPT system. The k measurements and calculations were performed with a well-known voltage transfer technique, as described in detail in [2].

[0029] Figures 4 and 5 show the coupling factor and link efficiency for different distance and transverse offset measurements.

[0030] Configurations which give rise to the same coupling coefficient are expected to achieve the same efficiency, i.e. perfectly aligned at a separation of $D = 40$ cm should achieve the same efficiency as an offset of $h = 21$ cm at $D = 30$ cm.

[0031] Figures 6 and 7 show k measurement results for TX and RX angular

5 misalignment, respectively. In Figure 6, while the TX coil angle increases, the distance between both coils reduces enough to compensate for angular misalignment. With the RX coil angle variation, a trough is seen at $\theta_{RX} = 75$ degrees, since this is the point at which the distance between the coils is not enough to compensate for the angular misalignment. These conclusions are based on a clear correlation between the coupling factor
10 measurements and dc-to-load efficiency.

[0032] Based on the above coupling factor analysis and measurements of the coils to be used in the later experiments, a high frequency, high power driver is required. In a typical IPT system, this is achieved by driving the coils with a 50Ω loop that is impedance
15 matched to a high frequency COTS RF transmitter with an output power amplifier. In this typical RF scenario, the maximum power transfer is achieved but not the desired maximum efficiency. In accordance with the invention, the number of stages can be reduced if the 50Ω impedance is avoided, by integrating the power amplifier and impedance matching circuits into one driver sub-system. This is achieved by carefully designing a high efficiency power amplifier capable of high frequency operation. The Class-E amplifier is an ideal
20 solution, since zero voltage and zero current switching can be achieved with the appropriate choice of components.

[0033] Class-E amplifiers have been designed and used extensively since Sokal and Sokal demonstrated the operational characteristics of their zero-switching power amplifier [32]. It is important to note, however, that even though this power amplifier topology is
25 widely known, designing high power amplifiers capable of working at 100 W and switching at a few MHz is not a trivial task. This is mainly due to the high power rating and fast switching capabilities that only a suitable power RF MOSFET can achieve, as well as the need to employ high Q capacitors. Furthermore, since an atypical non- 50Ω power amplifier is needed, to avoid additional impedance matching network components and their
30 associated losses, the resonant Class-E topology needs to be modified to suit the coils' characteristics.

[0034] To achieve a good efficiency, a semi-resonant Class-E topology was selected as a suitable solution [2]. Figure 8 shows the circuit of a semi-resonant Class-E amplifier for the transmitter resonant tank, where the apparent load (represented by the TX coil series
35 resistance R_{ps} and the effective receiver impedance R_{seq}) and the apparent inductor (represented by the primary coil's inductance L_p), appear to be larger, thus helping to

increase both driver and link efficiencies. This is achieved by tuning the primary resonant tank at a higher resonant frequency ω_{0TX} . This frequency is higher than the receiver's resonant tank driven resonant frequency $\omega_0 = \omega_{0RX}$, at which the MOSFET gate driver switches at an operating frequency ω_d , where $\omega_{0TX} > \omega_{0RX} \equiv \omega_d$. This semi-resonant operation also avoids the losses associated with an extra inductor, typically added in series with the TX coil to increase the driver efficiency [12]. Furthermore, as will be demonstrated below, the use of semi-resonant operation allows a simple but effective tuning mechanism; by modifying the frequency ratio ω_d / ω_{0TX} , the effective equivalent resistance and inductance of the primary tank can change for different operating scenarios.

[0035] Thus, in accordance with an embodiment of the invention, the inductive power transfer system illustrated in Figure 8 comprises an air core transmitter coil TX and an air core receiver coil RX spaced from the transmitter coil. The transmitter coil and/or the receiver coil has a diameter of at least 10cm. In the illustrated embodiment the diameter of the transmitter coil is 30cm and the diameter of the receiver coil is 20cm. The spacing between the transmitter coil and the receiver coil, in use, is at least 15cm. In the examples herein the spacing D between the transmitter coil and the receiver coil ranges between 30cm and 60cm.

[0036] The system further comprises a transmitter circuit comprising the transmitter coil and a receiver circuit comprising the receiver coil. The transmitter circuit is in the form of a Class E amplifier comprising a first inductor L_{choke} and a transistor (MOSFET) in series between the terminals of a power supply V_{DD} . The transistor is arranged to switch at a first frequency ω_d by means of a gate drive. A first capacitor C_{par} is arranged in parallel with the transistor between the first inductor L_{choke} and a power supply terminal. A primary resonant tank circuit is provided in parallel with the first capacitor C_{par} . The tank circuit comprises the transmitter coil and a second capacitor C_{res} arranged in parallel with the transmitter coil. A third capacitor C_{ser} is provided in series with the first inductor L_{choke} between the first capacitor C_{par} and the tank circuit. In Figure 8, the inductance of the transmitter coil TX is represented by the inductor L_p and the resistance of the transmitter coil is represented by the resistor R_{ps} . The resistor R_{seq} in Figure 8 represents the resistive effect of the receiver coil RX on the primary resonant tank circuit. The receiver circuit comprises a Class E rectifier.

[0037] The capacitance of the second capacitor C_{res} is selected, such that the resonant frequency ω_{0TX} of the tank circuit is greater than the first frequency ω_d . Typically, the ratio of the first frequency ω_d to the resonant frequency ω_{0TX} of the tank circuit is within the range 0.7 to 1. The receiver circuit has a resonant frequency ω_0 and the transmitter circuit

is configured to vary the first frequency ω_d , in order to achieve a desired impedance of the tank circuit. The first frequency ω_d is at least 80kHz, typically at least 1 MHz. The power transferred between the transmitter coil and the receiver coil is at least 10 milliwatts. In an alternative arrangement, the second capacitor C_{res} is arranged in series, rather than
 5 parallel, with the transmitter coil TX. In this case, the capacitance of the second capacitor C_{res} is selected, such that the resonant frequency ω_{oTX} of the tank circuit is less than the first frequency ω_d . Typically, the ratio of the first frequency ω_d to the resonant frequency ω_{oTX} of the tank circuit is within the range 1 to 1.5.

10 **[0038]** Using the results provided above, for an operating scenario with $D = 30$ cm and a perfect coil alignment, PSpice simulations were performed to validate the design equations and design guidelines presented in [2], [33], [34] but modified to account for semi-resonant operation.

15 **[0039]** The IXYSRF IXZ421DF12N100 module, which includes a DE375-102N12A power MOSFET and integrated gate driver, was selected as the best available MOSFET because of its high power handling and nanosecond switching capabilities. This module was also selected due to its relatively low output capacitance C_{oss} at drain-source voltage $V_{DS} = 230V$, required for 100W operation. It is important to note that C_{oss} is effectively absorbed by C_{par} and thus is a limiting factor for selecting the maximum ω_d / ω_{oTX} required for high efficiencies. Figure 9 shows this dependency, where a maximum $\omega_d / \omega_{oTX} = 0.82$, for the
 20 set of coils described in the previous section, can be achieved using the selected MOSFET.

25 **[0040]** Working past this threshold would result in a detuned Class-E amplifier, incapable of achieving zero-voltage, zero-current at the time of switching. At this optimal point, for the same power, V_{DS} will increase and I_{DS} will decrease, resulting in a greater Class-E efficiency.

[0041] During simulations, parasitic inductances and capacitances were added to the model as well as the effective series resistance (ESR) of the capacitors to account, as accurately as possible, for all the losses during operation. Variations of less than 5% in the Class-E capacitor values, compared to those used in the PSpice simulation, were required
 30 to achieve a zero voltage, zero current crossing and account for the high loaded Q of the resonant circuit due to a low coupling factor.

[0042] As shown in Figure 10, a smooth landing of V_{DS} was possible without any negative ringing and an almost ideal Class-E operation was achieved with a simulated dc-to-load efficiency of 80%.

[0043] Several key layout considerations had to be taken into account, in order to avoid ground bouncing and ensure good operation. The integrated driver/MOSFET module's input dc bus and gate signal were kept as short as possible; this also applied to ground paths for the module, PSU and load. To achieve this, ground planes were placed around all components in both layers of the standard low-cost FR-4 substrate, leaving arcing clearances around the tracks and components that were located close to the coil, where voltages as high as 1 kV are present during operation. Similarly, as with the tracks between the driver and MOSFET, all grounding tracks were kept as short as possible to decrease their resistance and inductance, but wide enough to avoid track lifting due to overheating.

[0044] In addition to layout considerations, component selection was crucial to enable high frequency operation. A combination of Dielectric Laboratories C40AH capacitor values were employed for C_{par} , C_{ser} and both resonator capacitors as they have very high Q and low ESR. Finally, the choice of the choke inductor that ensures only dc current from the PSU flows through the MOSFET was particularly challenging to design, due to the high current and high frequency characteristics of the system. A ferrite core was not suitable, due to its poor high frequency performance. For this reason, an iron powder core was selected, due to its low permeability and stability for high power applications, as well as high self-resonance frequency.

[0045] To fully characterize the practical IPT demonstrator system, shown in Figure 11, a thorough experimental analysis was performed. The main goal of these experiments was to investigate its behaviour in different scenarios, by varying distance, transverse offset and angular misalignment; this allows a comparison against perfect alignment, to which the system was initially tuned (based on simulations). Furthermore, the results from these experiments were compared against results from frequency tuning the system for each different scenario.

[0046] In all experiments, the input voltage V_{DD} was kept constant at 60V. This allowed sensible values of drain-source voltage to be obtained during operation, to which the output capacitance of the MOSFET is dependent, and helped to avoid replacing C_{par} in each experiment. A constant input DC power for all experiments could not be achieved. This is because, in several scenarios, driver operation was far away from Class-E operation, resulting in high losses during switching that could have degraded or destroyed the MOSFET module.

[0047] Due to the fact that the optimal load is very large (e.g. 21 k Ω for the simulated scenario), a non-inductive resistor was used. Special considerations were taken to achieve the desired load with a resistive network, because a non-inductive resistor capable

of handling more than 50W was not commercially available. Metal film resistors were chosen, given their low inductance behaviour at high frequencies and capacity to handle a few watts (enough to withstand up to 100W, once the load network was made). A major limitation for this type of resistor is that as its resistance and operating frequency increase the parasitic shunt capacitance also increases and as the resistor temperature increases its resistance varies. The parasitic capacitance for these resistors was calculated to be 2.8 pF at 6 MHz. This was taken into account when designing the load network and also the selection of the receiver's tuning capacitor. The total capacitance from the load resistor was absorbed by the calculated tuning capacitance, to ensure receiver resonance, thus avoiding the reflection of capacitive reactance to the transmitter side.

[0048] The dc-to-load efficiency of the system was initially measured using Agilent N2783A current probes, but, after several measurements, it was noted that the results were not reflecting the true operation of the circuit. Also, the current probes are not capable of measuring current accurately in the presence of significant electromagnetic noise [35]. The voltage across the load could not be measured with the oscilloscope probe, because the probe's capacitance is 15 pF, enough to detune the receiver coil from resonance. For these reasons, and the fact that the resistor's precise temperature dependence is unknown, an indirect method of measuring the dc-to-load efficiency was implemented. Power was inferred from accurate steady-state heat-sink temperature measurements, since both the driver and the load (including the tuning capacitors) were placed over separated, isolated heat sinks without forced-air cooling. The input DC power was also measured accurately and used together with the RX thermal measurements to calculate the dc-to-load efficiency using the following:

$$\eta_{dc-load} = \frac{T_{ssRX} - T_{amb}}{R_{thRX}(T)P_{dc}}$$

where T_{amb} is the ambient temperature, T_{ssRX} is the heat sink steady-state temperature of the receiving coils and $R_{thRX}(T)$ is the lumped thermal resistance of the RX load. The temperature measurements were calibrated by applying a known DC power to the RX load until all temperatures reached steady state. Measurements under the same thermal experimental conditions as when the IPT system was tested were performed.

[0049] Due to the RX load spatial distribution over the heat sink and the fact that the heat sink was positioned with the fins facing downwards on the bench, with a 333 K temperature gradient $R_{thRX} = 208$ K/W compared well to the manufacturer's 203 K/W. Furthermore, by characterizing the load arrangement, the non-linear behaviour of the heat

sink was accounted for, which can be as high as 25% to 50% of the dissipated heat, according to [36].

[0050] It is important to note that this is a conservative dc-to-load efficiency calculation, since T_{amb} will increase as T_{ssRX} increases; giving a lower $\eta_{dc-load}$ when compared to the scenario where T_{amb} could be kept constant until the steady state of the system is reached. Even more important is the fact that as the temperature of the resistors increases the value of the load resistance will start to drift away from its optimal value, drifting away from maximum efficiency.

[0051] To achieve a semi-resonant Class-E operation similar to that observed by simulations, an iterative tuning process was performed. This establishes the appropriate values for the driver and the coil capacitors (taking into account their fabrication tolerances) and the coil Q variations (due to metallic objects, such as bench supports, being in close physical proximity to the experiments).

[0052] First, the receiver resonator's capacitor had to be decreased, to account for the load resistor's equivalent shunt capacitance for correct receiver resonance. With an untuned receiver, the value of the transmitter resonator's capacitor would need to change, to account for the reflected reactance from the receiver onto the transmitter and ensure that semi-resonance operation is still present. This changes the ratio of ω_d / ω_{oTX} , which creates the need for retuning C_{par} and C_{ser} . The major limitation of this scenario is that if the ratio ω_d / ω_{oTX} starts to increase, there is a point at which the required C_{par} needed to tune the driver is lower than C_{oss} of the MOSFET, which makes the MOSFET unsuitable for Class-E operation.

[0053] Once receiver resonance is obtained, a similar procedure can be followed to achieve zero switching operation, as described in [33]. Since C_{par} is implemented by an external physical capacitor and C_{oss} , which is dependent on V_{DS} , extra iterations are needed to achieve good operation. As seen from simulations, V_{DS} is a very useful guide to Class-E operation, therefore tuning the peak-to-peak voltage is as important as achieving zero voltage zero current switching to increase the efficiency [34]. Based on [32], if $V_{DS} > 3.56 V_{DD}$, C_{par} needs to be increased in steps of 5 pF and if $V_{DS} < 3.56 V_{DD}$, it should be decreased by the same amount until the correct V_{DS} is achieved. While doing this, C_{ser} may need a slight adjustment to bring back the driver to zero switching.

[0054] Since the load resistance value varies with temperature, several iterations were performed by increasing or decreasing the load resistance by $\pm 5\%$, until a maximum efficiency of 66% was achieved. For each iteration, receiver resonance was achieved and tuning of C_{par} and C_{ser} was performed as previously described. Figure 12 shows the drain-

source voltage (simulated and measured) for the tuned IPT system for an aligned set of coils at a separation distance of 30 cm; the input dc power was 90 W.

[0055] Whilst the waveform generated from simulations has a smooth landing, a negative ripple of less than 10V can be seen when the MOSFET is ON. This negative ripple is generated by a small voltage still present at the drain, when the MOSFET is turned ON, due to a higher than expected loaded Q for the transmitter resonant circuit. This was the best possible switching achieved with the discrete capacitors available. The higher measured V_{DS} results could be decreased by adding more capacitance to C_{par} , but this change was not reflected in a measurable efficiency improvement. Therefore, to decrease the losses in the capacitor (due to ESR), no extra parallel capacitor was added and V_{DS} was left higher than expected in the simulation.

[0056] Once the IPT system was optimally tuned to achieve a high efficiency, while being perfectly aligned, measurements with different separation distances were performed to observe the operating capabilities without performing extra tuning. As can be seen in Figure 13, the distance between the coils was varied between 30 and 60 cm. As D increases the efficiency decreases as the coupling factor decreases, detuning the driver and creating the need for different C_{par} and C_{ser} values to re-establish zero switching operation.

[0057] An easier tuning alternative is to change the operating frequency of the clock, thus relying on the semi-resonant operation of the driver. When this was performed, the receiver was no longer in resonance and the transmitter sees a reflected reactance. This extra reactance, in addition to the transmitter's reactance, was enough to improve the tuning of the semi-resonant Class-E driver, modifying ω_d / ω_{oTX} and shifting the driver's waveforms closer to zero-switching operation. As seen in Figure 14, as the clock frequency was altered for each different measurement, the efficiency increased considerably over untuned operation. The dc-to-load efficiency from the clock-frequency tuned version at a 50 cm separation distance was 25%, compared to 20%, as seen in Figure 13. It is important to note that to achieve this increase in efficiency, a clock frequency change of less than 1% was required.

[0058] To analyse the efficiency of the IPT system, for a set of scenarios with different offsets, as shown in Figure 2, measurements for both the perfectly aligned 30 cm impedance tuning with fixed clock frequency case and with clock frequency tuning were performed. Figure 15 shows the results for IPT with different coil offsets. In this case, dc-to-load efficiency decreases; following a similar trend as the corresponding measurements above. It can be seen that even with an offset of 10 cm, and no additional tuning, the IPT system performed with a dc-to-load efficiency above 58%.

[0059] With clock frequency tuning, as shown in Figure 16, the dc-to-load efficiency was above 50%, with $h < 14$ cm, and an efficiency increase of 5% was achieved with an offset of 20 cm.

[0060] Although higher dc-to-load efficiencies could be achieved with tuning the optimal load for each offset, the results presented in this figure demonstrate that efficiencies above 50% can be achieved even at highly misaligned scenarios without the need for load tuning or complex and heavy coupling factor enhancement techniques.

[0061] To characterise how the system operates while varying coil misalignment angle θ , in both transmitter and receiver, measurements at a fixed distance of $D = 30$ cm were performed, as shown in Figure 3. From Figures 17 and 18, measurements for a varying TX coil angle θ_{TX} were performed. As predicted, by the coupling factor measurements, a constant high efficiency was achieved for angles below 75 degrees when clock frequency tuning was performed. In contrast, a decrease in efficiency, not clearly linked with k was seen with the fixed clock frequency scenario. This is due to the fact that with the fixed clock frequency tuning case the presence of a larger reflecting load from the receiver influenced efficiency more than with the relatively large and constant coupling factor in the tuning of the IPT system. With the clock frequency tuning scenario, the frequency variation was enough to tune the Class-E and exploit the benefit of almost constant k . DC-to-load efficiencies above 60% were achieved for almost all $\theta_{TX} < 72$ degrees with clock frequency tuning, showing the capabilities of the system to perform in a wide range of transmission angles with a fractional frequency variation of less than 6%.

[0062] Finally, measurements with a varying θ_{RX} were performed, as shown in Figure 19. The efficiency was almost constant and above 50% up to $\theta_{RX} = 52$ degrees. Above this angle, the efficiency decreased dramatically, as predicted by the coupling factor measurements. A noticeable difference could be appreciated in Figure 20, with $\theta_{RX} = 45$ degrees, where the dc-to-load efficiency was 56% in the clock frequency tuned case and only 40% in the fixed clock frequency impedance tuned version.

[0063] Finally, the efficiency of the system was increased until the power dissipated by the load resistors caused them to overheat and fail. The highest dc-to-load efficiency achieved with the current prototype design was $\eta_{dc-load} = 77\%$ for an aligned set of coils at a distance of 30 cm with $P_{load} = 105$ W.

[0064] The calculated link efficiency based on the unloaded Q and k measurements was 95%. The dc-to-load efficiency of the system was increased thanks to a higher V_{DD} , which allowed $C_{oss} = C_{par}$. This avoided the use of an external capacitor and allowed for a higher ω_d / ω_{oTX} , which increased the apparent driver inductance of the system. This is believed

to be the highest dc-to-load efficiency ever presented for an IPT system, without k enhancement techniques.

[0065] A comparison of state-of-the-art IPT systems has been given above, a clear efficiency analysis is suggested for a meaningful comparison between competing solutions and key differences between link and dc-to-load efficiencies have been highlighted.

[0066] An indirect thermal method for measuring P_{load} has been presented for the first time with an IPT system, to avoid measurement inaccuracies due to load resistance variations and high external electromagnetic fields in the current measurements.

[0067] This method was compared against the above coupling factor formula and well-known coupling factor measurements and a clear correlation can be seen, demonstrating the robustness of the efficiency measurement procedure. Low cost, high Q coils and a complete design and operational analysis of a semi-resonant Class-E driver for this IPT system has been described. The driver topology and component selection enabled high frequency, medium power, wireless power transfer for different transmitter and receiver coil sizes. A detailed transverse offset and angular misalignment characterization demonstrated efficiencies above 50% for transverse offsets up to 14 cm and $\theta_{\text{RX}} = 52$ degrees.

[0068] Finally, dc-to-load efficiencies of 77% were demonstrated in a perfectly aligned scenario for $D = 30$ cm, having a link efficiency of 95%.

[0069] Figure 21 is a schematic representation of a typical Class E receiver circuit according to an embodiment of the invention. The receiver circuit comprises the air core receiver coil RX of the inductive power transfer system described above. In Figure 21, the inductance of the receiver coil RX is represented by the inductor L_{RX} and the resistance of the receiver coil is represented by the resistor R_{RXS} . The induced signal in the receiver coil is represented by the signal generator $k\eta V_{\text{TX}}$. The resistor R_L in Figure 21 represents the load powered by the receiver circuit.

[0070] The receiver circuit comprises a first capacitor C_L arranged in parallel with the load R_L . A secondary resonant tank circuit is provided in parallel with the first capacitor C_L . The tank circuit comprises the receiver coil and a second capacitor C_{res} arranged in parallel with the receiver coil. In an alternative embodiment, the second capacitor C_{res} is arranged in series, rather than parallel, with the receiver coil. A first inductor L_r is arranged in series with a first diode D_r between the tank circuit and the first capacitor C_L . The junction capacitance of the first diode D_r is represented in Figure 21 by the capacitor C_j in parallel with the first diode D_r . A third capacitor C_r is arranged in parallel with the first diode D_r .

[0071] In traditional operation, the capacitance of the second capacitor C_{res} is selected, such that the resonant frequency ω_{oRX} of the secondary tank circuit is equal to the first frequency ω_d , i.e. the gate drive frequency of the transmitter circuit.

[0072] Figure 22 is a schematic representation of Class E receiver circuit according to a further embodiment of the invention. This receiver circuit differs from the circuit of Figure 21 in that the first inductor L_r is not present. In addition, the capacitance of the second capacitor C_{res} is selected, such that the resonant frequency ω_{oRX} of the secondary tank circuit differs from the first frequency ω_d , the gate drive frequency of the transmitter circuit. The secondary tank circuit therefore operates in semi-resonance and maintains some inductive impedance which performs the role of the first inductor L_r in the receiver circuit of Figure 21. In this way, an inductor in addition to the receiver coil is not required. In the receiver circuit of Figure 21, the first inductor L_r generates a significant amount of heat because of the large AC current it carries. This potential source of power loss is therefore avoided in the receiver circuit of Figure 22.

[0073] Figure 23 is a schematic representation of Class E receiver circuit according to a yet further embodiment of the invention. This receiver circuit differs from the circuit of Figure 22 in that the third capacitor C_r is not present. Instead, a second diode is provided in parallel with the first diode. The junction capacitances C_{j1} , C_{j2} of the two diodes provide the equivalent capacitance to the third capacitor C_r and the junction capacitance C_j in the circuit of Figures 21 and 22. In this way, the operation of the receiver circuit is not constrained by the maximum voltage of the third capacitor C_r . Silicon carbide (SiC) diodes provide the appropriate junction capacitance at the required operating voltages.

[0074] In the embodiment of Figures 22 and 23, the ratio ω_d / ω_{oRX} of the first frequency ω_d to the resonant frequency ω_{oRX} of the receiver tank is in the range 0.2 to 3. This range allows the tuning of the rectifier circuit for a desired output voltage and efficiency across different input voltages and receiver coil inductances. The inductance L_{RX} of the receiver coil is typically in the range 0.5 to 8 μ H. The capacitance of the first capacitor C_L is typically in the range 0.02 to 100 μ F. The capacitance of the second capacitor C_{res} is typically in the range 8 pF to 1.5 nF.

[0075] The transmitter circuit of Figure 8 may be used in combination with the receiver circuits of any of Figures 21 to 23, but the circuit of Figure 23 is presently preferred.

[0076] In summary, an inductive power transfer system comprises a transmitter coil TX and a receiver coil RX spaced from the transmitter coil. A transmitter circuit comprises the transmitter coil and is in the form of a Class E amplifier with a first inductor L_{choke} and a transistor in series between the terminals of a power supply, a first transmitter capacitor C_{par} in parallel with the transistor between the first inductor and a power supply terminal, a

primary tank circuit in parallel with the first transmitter capacitor, the primary tank circuit comprising the transmitter coil and a second transmitter capacitor C_{res} arranged in parallel or series with the transmitter coil, and a third transmitter capacitor C_{ser} in series with the first inductor between the first transmitter capacitor and the primary tank circuit. The

5 transistor is arranged to switch at a first frequency ω_d and the capacitance of the second transmitter capacitor is selected such that the resonant frequency ω_{oTX} of the primary tank circuit is greater than the first frequency. The receiver circuit comprises a Class E rectifier having a first receiver capacitor C_L arranged in parallel with a load R_L and a secondary tank circuit in parallel with the first receiver capacitor. The secondary tank circuit

10 comprises the receiver coil and a second receiver capacitor C_{res} arranged in parallel or series with the receiver coil. A first diode D_{r2} is provided between the secondary tank circuit and the first receiver capacitor. The capacitance of the second receiver capacitor is selected such that the resonant frequency ω_{oRX} of the secondary tank circuit differs from the first frequency, so that the secondary tank circuit operates in semi-resonance and

15 maintains some reactive impedance. The transmitter circuit is configured to vary the first frequency, in order to achieve a desired impedance of the primary tank circuit.

[0077] Throughout the description and claims of this specification, the words “comprise” and “contain” and variations of them mean “including but not limited to”, and they are not intended to (and do not) exclude other components, integers or steps. Throughout the

20 description and claims of this specification, the singular encompasses the plural unless the context otherwise requires. In particular, where the indefinite article is used, the specification is to be understood as contemplating plurality as well as singularity, unless the context requires otherwise.

[0078] Features, integers, characteristics or groups described in conjunction with a

25 particular aspect, embodiment or example of the invention are to be understood to be applicable to any other aspect, embodiment or example described herein unless incompatible therewith. All of the features disclosed in this specification (including any accompanying claims, abstract and drawings), and/or all of the steps of any method or process so disclosed, may be combined in any combination, except combinations where at

30 least some of such features and/or steps are mutually exclusive. The invention is not restricted to the details of any foregoing embodiments. The invention extends to any novel one, or any novel combination, of the features disclosed in this specification (including any accompanying claims, abstract and drawings), or to any novel one, or any novel combination, of the steps of any method or process so disclosed.

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CLAIMS

1. An inductive power transfer system comprising:
 - a transmitter coil and a receiver coil spaced from the transmitter coil;
 - a transmitter circuit comprising the transmitter coil; and
 - 5 a receiver circuit comprising the receiver coil,wherein the transmitter circuit is in the form of a Class E amplifier comprising:
 - a first inductor and a transistor in series between the terminals of a power supply, the transistor being arranged to switch at a first frequency;
 - a first transmitter capacitor in parallel with the transistor between the first inductor
 - 10 and a power supply terminal;
 - a primary tank circuit in parallel with the first transmitter capacitor, the primary tank circuit comprising the transmitter coil and a second transmitter capacitor arranged in parallel or series with the transmitter coil;
 - a third transmitter capacitor in series with the first inductor between the first
 - 15 transmitter capacitor and the primary tank circuit.
2. An inductive power transfer system as claimed in claim 1, wherein the capacitance of the second transmitter capacitor is selected, such that the resonant frequency of the primary tank circuit is greater than the first frequency.
3. An inductive power transfer system as claimed in claim 1, wherein the capacitance
- 20 of the second transmitter capacitor is selected, such that the resonant frequency of the primary tank circuit is less than the first frequency.
4. An inductive power transfer system as claimed in claim 2 or 3, wherein the ratio of the first frequency to the resonant frequency of the primary tank circuit is in the range 0.5 to 1.5.
- 25 5. An inductive power transfer system as claimed in claim 2, wherein the ratio of the first frequency to the resonant frequency of the primary tank circuit is in the range 0.7 to 0.9.
6. An inductive power transfer system as claimed in any preceding claim, wherein the receiver circuit has a resonant frequency and the transmitter circuit is configured to vary
- 30 the first frequency, in order to achieve a desired effective impedance of the primary tank circuit.

7. An inductive power transfer system as claimed in any preceding claim, wherein the transmitter coil and/or the receiver coil has an air core.
8. An inductive power transfer system as claimed in any preceding claim, wherein the transmitter coil and/or the receiver coil has a diameter of at least 5cm, preferably at least 10cm.
9. An inductive power transfer system as claimed in any preceding claim, wherein the spacing between the transmitter coil and the receiver coil, in use, is at least 15cm.
10. An inductive power transfer system as claimed in any preceding claim, wherein the transistor is a MOSFET.
- 10 11. An inductive power transfer system as claimed in any preceding claim, wherein the first frequency is at least 80kHz, preferably at least 1 MHz.
12. An inductive power transfer system as claimed in any preceding claim, wherein the power transferred between the transmitter coil and the receiver coil is at least 1 watt, preferably at least 10 watts.
- 15 13. An inductive power transfer system as claimed in any preceding claim, wherein the receiver circuit comprises a Class E rectifier.
14. An inductive power transfer system as claimed in claim 13, wherein the receiver circuit comprises a first receiver capacitor arranged in parallel with a load, in use, and a secondary tank circuit in parallel with the first receiver capacitor, wherein the secondary tank circuit comprises the receiver coil and a second receiver capacitor arranged in parallel with the receiver coil, and a first diode is provided between the secondary tank circuit and the first receiver capacitor.
- 20 15. An inductive power transfer system as claimed in claim 14, wherein the capacitance of the second receiver capacitor is selected such that the resonant frequency of the secondary tank circuit differs from the first frequency, whereby the secondary tank circuit operates in semi-resonance and maintains some reactive impedance.
- 25 16. An inductive power transfer system as claimed in claim 15, wherein the ratio of the first frequency to the resonant frequency of the secondary tank circuit is in the range 0.2 to 3.
- 30 17. An inductive power transfer system as claimed in any of claims 13 to 16, wherein the only inductor in the receiver circuit is the receiver coil.
18. An inductive power transfer system as claimed in any of claims 14 to 17, wherein the receiver circuit comprises at least a second diode in parallel with the first diode.

19. An inductive power transfer system as claimed in any of claims 13 to 18, wherein the only capacitance in parallel with the diode(s) is provided by the junction capacitance of the diode(s).
20. An inductive power transfer system as claimed in any of claims 13 to 19, wherein
5 the diode(s) are silicon carbide diodes, gallium nitride diodes or are formed of other wide band gap material.
21. The receiver circuit of the inductive power transfer system as claimed in any of claims 14 to 20.
22. The transmitter circuit of the inductive power transfer system as claimed in any
10 preceding claim.