A High-Power Multiphase Wireless Dynamic Charging System with low Output Power Pulsation for Electric Vehicles

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Abstract—Wireless Dynamic Charging (WDC) of Electric Vehicles (EV) is a new initiative aimed to increase EV uptake. However, pulsating output power along the driving direction is a major problem with this technology. Moreover, in order to deliver enough driving power for EVs, WDC systems must operate at high power. This paper proposes a WDC system that has the capability of providing a low pulsating and high output power to EVs. The proposed WDC utilizes multiple primary windings that guarantees a homogeneous mutual magnetic flux for the receiver along the driving direction. This results in a constant induced voltage across the receiver and hence constant output power to charge the EV battery. High output power is realized by using multiple transmitter windings which have been arranged in a novel winding method. The structure layout of the proposed WDC, including ferrite cores, windings and resonant tanks is presented. Furthermore, a theoretical analysis is conducted to determine conditions of each winding's current phase and amplitude for achieving constant output power. The crossing mutual inductances between the different transmitter's phases are compensated by adding small capacitors in series with each transmitter winding. An optimization analysis using Maxwell 3D simulation for both, transmitter and receiver is carried out to achieve the highest coupling factor by using minimum ferrite material. The effectiveness of the proposed system is analytically demonstrated and experimentally verified using a 3-kW laboratory prototype. Finally, the performance of the proposed method is compared with similar systems presented in the literature. The comparison shows that the proposed system has the advantages of using simple control, it also eliminates communications between the primary and secondary side and delivers a normalized power of 2.25 which is 125% higher compared to conventional single-phase systems.

Index Terms—Dynamic Wireless Charging Inductive Power Transfer, Electric Vehicle Battery Charger, Constant Output Power.

I. INTRODUCTION

Electric vehicles (EV) have recently received a great deal of attention because of their clean, efficient and environment-friendly features [1]. The high penetration of EVs still limited at present time due to many technological barriers, including battery capacity. An expensive and large battery with long charging time is normally equipped to achieve satisfactory EV driving range. To mitigate this problem, the concept of WDC has been proposed in [2] to overcome the limitation of current battery technology. WDC enables EV to be charged wirelessly while in motion. As a result, the battery size of EVs can be greatly reduced and the driving range limitation can be

completely alleviated. WDC is commonly developed based on inductive power transfer (IPT) technology, where a time varying magnetic field is generated by transmitter coils; which are installed underneath road surface, to wirelessly power receiver coils, which charging the EV's battery continuously [3]-[5].

A WDC are categorized into short-individual transmitters [6-8] or long-track transmitters [9-11] according to the length of transmitter coils (i.e. Fig. 1a and Fig. 1b respectively). In the short-individual transmitters, the transmitter dimension is the same as the receiver, which is typical within 1m². Furthermore, multi-transmitter coils are arranged in the primary side to alternatively transfer power to secondary side. Additionally, each transmitter is connected to an independent compensation circuit, which can be switched ON or OFF depending on the EV's position. When the EV's receiver is in align with a specific transmitter, then the power switch (or contactor) connects the associated inverter. In order to switch a specific transmitter, position sensors and associated electronic circuit are required. Therefore, the effect of the electromagnetic field to surrounding environment is minimized and total conduction losses on transmitters are reduced. Moreover, the high efficiency can be achieved due to high values of coupling coefficient between transmitter and receiver coil. However, the short-individual transmitters requires more inverters, position sensors and other control circuitries that make the system more complicated. More importantly, when the EV moves between two adjacent transmitters, the output power becomes pulsating and significantly reduced.

On the other hand, long-track transmitter [9-11] is much longer than the EV's length and only requires a single power inverter with one compensation tank. This brings benefits in terms of cost and simplicity for implementation of such a system. However, the low coupling coefficient is indicated as the main drawback of this system as the receiver coil covers only a small portion of the long transmitter coil. Consequently, the efficiency is reduced and electromagnetic interference (EMI) is increased. Several studies are conducted to address EMI problem and also to simplify the structure of the system [12-14]. In [12] and [13], I-type and S-type transmitters are proposed with a width of 10 cm and 4 cm, respectively. Additional, by adopting the alternating magnetic polarity for adjacent poles, the EMI around the transmitter coils can be substantially reduced through electromagnetic field cancelation between two adjacent magnetic poles of opposite polarity. The N-type is proposed in [14] to further reduce the volume of ferrite core while maintaining the advantages of I-type structure. However, the studies also pointed out that the magnetic flux density generated by I-type, S-type or N-type power supply on a single receiver varies in a nearly sinusoidal function depending on receiver's position along the driving direction.

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According to [12], magnetic flux density reaches to its maximum at the centre of each ferrite pole and reduces to nearly zero at the centre between two adjacent poles. As a consequence, the induced voltage and received power are significantly varied depending on EV's position. The received output power even reaches to nearly zero at some specific position of EV while moving along the transmitter.

As discussed in the above literature, the output power variation is one of the most serious problems for WDC, in both short-track and long-track systems. Therefore, a number of studies attempted to realize constant output power. For example, in short-track system, the stable output power is achieved by proper design of the compensation network to compensate for misalignment [15-16]. However, this can only be attained for a narrow range of misalignment. Further, in [17], the variation of the output power is reduced by arranging the transmitter segments closer to each other. However, this can be only achieved at a certain dimensions of the transmitter and receiver. In [18], the low output power variation is attained by adopting three half-bridge inverters connected in parallel as well as optimizing the receiver length. Constant power is only achieved if the ratio of receiver's length over transmitter's length is 1.5, which may not be suitable in real practical applications where receivers with different dimensions are required. Other methods are proposed in [19-20] by adjusting the transmitter current under misalignment. Transmitter current is controlled by a primary inverter and/or a DC/DC converter to increase its amplitude when coupling between two sides is reduced. However, overall efficiency is decreased as the increase in losses on Inverter and Transmitter winding. Moreover, mutual inductance measurement and communication between two sides are required while EV is in motion.

Similarly, several attempts have been reported in the literature aiming at reducing the output pulsation of long-track systems. For instance, authors in [21] added additional receiver windings to create a multiphase receiver system with a single primary transmitter. The coils dimensions in both sides are optimized to achieve constant power at the output side. Another solution was suggested by [22], where two primary transmitter coils with two independently controlled inverters are arranged. The constant output power can be attained by controlling the output amplitude and phase of the two inverters. However, this requires information from the secondary side such as load resistance and mutual inductances, which results in a complicated measurement and communication especially when the EV is in motion. Furthermore, the control method presented in [22] becomes even more complicated when multiple loads (EVs) are considered, simultaneously.

Besides output power pulsation problems, the transferred power of a single-phase WDC system is also limited by the rating of semiconductor devices, transmitter coil and compensation circuit. In real applications, several EVs may need to be charged simultaneously, thus the high power demand becomes a necessity. Moreover, the power level required for public transport such as electric bus or train systems may go up to hundreds of kilowatts or even megawatts scale [23]. In these applications, single-phase system may not be able to produce such a requirement. Therefore, multi-phase systems are seen as a solution to transfer higher output power [24-25].

The work in this paper achieves for the first time constant and high output power through the utilization of multiphase

transmitter system. The transmitter system is constructed using multiple magnetic poles with different polarities and multiple windings. It should be noted that the proposed system possesses the same features of [12 -14], however it delivers higher and constant power. A multiphase inverter is adopted to produce the required currents (i.e. amplitude and phase) for each transmitter's winding. Modified primary LCC tank is proposed to compensate for the crossing coupling between transmitter windings. The paper presents a generalized multiphase system, however the number of phases is selected based on the required output power. Maxwell 3D simulations is carried out to design and optimize the magnetic components for attaining the highest value of coupling factor with a minimum ferrite core under misalignment conditions. The proposed concept is validated with an illustrative example based on a three-phase system. A scale down laboratory prototype of 3-kW is developed to experimentally verify the feasibility and effectiveness of the proposed system. Furthermore, in order to show the superiority of the proposed system, a comparison with similar systems is carried out [12-14], [21-22]. The following are the main benefits:

- (1) Achieving constant output power with a simple control requirement.
- (2) Less number of inverters compared to [22].
- (3) Elimination of measurement of mutual inductance and load impedance as well as communication between the grid and vehicle side while the EV is in motion.
- (4) Attaining higher output power (i.e. 2.25 higher for the three-phase case) for the same system parameters, e.g. mutual inductance, transmitter's current, switching frequency and receiver's dimension, compared to the single-phase system.
- (5) Unlike the previous systems, the proposed approach can be easily extended to multiple EV charging without any changes to the primary side control. This will be more suitable for real scenarios.

The paper is organized as follows. The proposed circuit configuration and multiphase transmitter layout are presented in Section II. An illustrative design example with three-phase system is mentioned in Section III. In Section IV, an optimized design of transmitter and receiver is conducted to achieve the highest of coupling factor. A 3-kW prototype is built and tested to demonstrate the validity of the proposed method and to evaluate the performance of the proposed WDC system. The system comparison with other similar systems [12, 13, 14, 21, 22] is presented in Section V. Finally, some conclusions are given in Section VI.

Fig. 1. Two types of WDC. (a) Long track transmitter; (b) Short-individual transmitters

II. ANALYSIS OF THE PROPOSED MULTIPHASE WDC

A. The circuit configuration

Fig. 2a illustrates the proposed multiple transmitters WDC aiming to achieve constant output power for an EV's dynamic charging system. Primary side consists of n transmitters L_i ($i = 1, 2, \dots, n$) energizing one receiver coil, L_s . As these transmitters are physically placed closed to each other, crossing couplings exists between them as depicted in Fig. 2.a. The mutual inductances between transmitter L_i and receiver is denoted by M_{is} ($i = 1, 2, \dots, n$). In the experiment a $0.2\text{-}\Omega$ resistance was measured for each transmitter coil. At 3kW output power an RMS current of 14 A is required and this produces a total loss of 120 W . As such a small percentage is caused by the coils' resistances, then their values are neglected in the analysis for simplicity. Each transmitter is driven by a constant current source regardless of load and coupling coefficient conditions. Detailed structures of transmitter and receiver are described in the next Section. A multi-phase DC/AC inverter is adopted at the primary side to produce a high frequency AC voltage for each transmitter. The Inverter consists of n legs, namely leg 1, leg 2,..., leg i ,... leg n . Each leg has two switches that operate complementally to avoid short circuiting the dc supply. The operating waveforms of the multi-phase inverter is depicted in Fig. 2b, which includes the gate signals of active switches and the associated inverter's output voltages. The phase-shift between two adjacent legs' output voltages (i.e. $V_{(i-1)i}$ and $V_{i(i+1)}$) is set at $2\pi/n$. The switching frequency of the inverter is chosen to be 85 kHz , in accordance with SAE J2954 standard [26] and the period of gate signal is $T = 1/f$.

A modified LCC resonant circuit based on [27], which consists of L_{ia} , C_{ia} and C_{ib} ($i = 1, 2, \dots, n$), is used in primary side. In order to produce constant transmitter currents I_i ($i = 1, 2, \dots, n$), the resonance between L_{ia} and C_{ia} is set at the switching frequency of the inverter. The series capacitors C_{ib} ($i = 1, 2, \dots, n$) are used here for two purposes: (1) to resonate with the transmitter's inductance L_i and (2) to suppress the induced voltages caused by the cross-couplings from other transmitters. Therefore, values of C_{ib} ($i = 1, 2, \dots, n$) in this work are different from previous papers on LCC resonant network [4], [27]. Please note that the phase of the transmitter current I_i is always leading 90 degrees as compared to the phase of Inverter's output voltage $V_{i(i+1)}$. At the secondary side, another LCC resonant tank is connected with a full-bridge rectifier to provide DC power to the output load. There should be another DC/DC regulator after the rectifier for battery charging purposes. However, for simplicity, only resistive load is used to investigate the proposed method. The proposed layout of primary transmitters and conditions of currents I_i ($i = 1, 2, \dots, n$) are further detailed in the following subsections. It is worth noting that the configuration in Fig. 2(a) is only applicable with a minimum number of phases (n) of three. However, when n equals to 1 or 2, the primary inverter connection needs to be modified as shown in Fig. 2(c) and (d).

Fig. 2 (a) The proposed WDC circuit with multiple transmitter coils ($n \geq 3$), (b) The operating waveforms of the multiphase Inverter, Modification of inverter connection with (c) $n = 1$, (d) $n = 2$

B. The layout of the proposed multiple transmitters

The derivation of the proposed system is inspired by the limitations of a single transmitter system. Fig. 3(a) illustrates WDC that consists of multiple alternating magnetic I-pole core, where the receiver is covering one magnetic pole. When receiver is perfectly aligned with center of each magnetic pole, i.e. completely covers the winding loop, then mutual inductance will reach its maximum value. On the other hand, the mutual inductance drops to zero when the receiver centers at the middle point between each adjacent poles, where the magnetic fluxes cancel each other (i.e. same magnitude but opposite directions), resulting in no induced voltage. Therefore, as the receiver (EV) moves along the transmitter, spatial periodic induced voltage and power (P_o) are obtained, which both are proportional with mutual inductance value. Consequently, as the vehicle travels along, the battery will not receive any power at the points of zero-mutual-inductance (i.e. *pulsating power*). The spatial distribution of mutual inductance between a single transmitter and receiver $M_{1s}(x)$ can be approximately expressed by Eq. (1) where M_o and l_o are the maximum mutual inductance and the length of two adjacent poles respectively. The measured and simulated mutual inductance for 2.4m and 4-pole single transmitter system are portrayed in Fig 3(b) (detailed parameters can be found in Section IV).

$$M_{1s}(x) = M_o \sin\left(\frac{2\pi x}{l_o}\right) \quad (1)$$

Fig. 3 Single transmitter system (a) Side view when receiver positions in lowest mutual inductance point (b) Spatial distribution of mutual inductances

In order to solve abovementioned problems, additional transmitters can be added into the original system to nullify the zero mutual inductance points. The number of additional transmitters is determined by the output power requirement volt-ampere capability of the system components such as power switches, transmitter coils and resonant capacitors. However, increasing number of transmitters results in higher implementing costs, therefore the selection of number of transmitter is a trade-off between cost and reliability. To attain a constant output power, it is essential to ensure that all mutual inductances $M_{is}(x)$ ($i = 2, 3, \dots, n$) are evenly distributed along the transmitter length to avoid any zero mutual inductance points. It refers that between any two adjacent null points of $M_{1s}(x)$, there are new $(n - 1)$ null

points from added $M_{is}(x)$ ($i = 2, 3, \dots, n$). In other words, the phase difference between any two adjacent mutual inductances $M_{is}(x)$ and $M_{(i+1)s}(x)$ should be equal as illustrated in Fig. 4. It is also avoided to add new null points of all additional $M_{is}(x)$ ($i = 2, 3, \dots, n$) into the same positions with the existing one of $M_{1s}(x)$, otherwise, added $M_{is}(x)$ does not contribute to compensate the null-point of $M_{1s}(x)$. The phase-shift difference between any two adjacent mutual inductance curves $M_{is}(x)$ and $M_{(i+1)s}(x)$ is set at $2\pi/n$. Equation (2) expresses all (n) mutual inductances (i.e. $M_{is}(x)$ ($i = 1, 2, \dots, n$)) including the original $M_{1s}(x)$. It is worth mentioning that $M_{(i+1)s}(x)$ can be either leading or lagging with a phase angle of $2\pi/n$ with respect to $M_{is}(x)$. Although in this work, $M_{(i+1)s}(x)$ is lagging $M_{is}(x)$, either way (lagging or leading) will result in compensating the zero mutual inductance points. Equation (2) satisfies with any number of phases. There is only one exception if $n = 2$ where the phase-shift between two adjacent phases is $\pi/2$ instead of π . This is to guarantee the null points of $M_{1s}(x)$ and $M_{2s}(x)$ do not overlap.

$$\begin{cases} M_{1s}(x) = M_o \sin\left(\frac{2\pi x}{l_o}\right) \\ M_{2s}(x) = M_o \sin\left(\frac{2\pi x}{l_o} - \frac{2\pi}{n}\right) \\ \vdots \\ M_{is}(x) = M_o \sin\left[\frac{2\pi x}{l_o} - (i-1)\frac{2\pi}{n}\right] \\ \vdots \\ M_{ns}(x) = M_o \sin\left[\frac{2\pi x}{l_o} - (n-1)\frac{2\pi}{n}\right] \end{cases} \quad (2)$$

Fig. 4 Expected spatial distribution of mutual inductances for the proposed multiple transmitters

In order to obtain evenly distributed $M_{is}(x)$ as shown in (2), the proposed n transmitters is structured as depicted in Fig. 5. The proposed transmitter is based on I-type ferrite shape as presented in [12]. As discussed above, $(n - 1)$ additional mutual inductances are evenly distributed over the length of l_o . Furthermore, in the proposed system, each original core of the conventional single transmitter (i.e. Fig. 5a) is divided into n small cores (Fig. 5b). These small cores dimensions are adjustable and determined by the application requirements. Please note that if the number of phases is an even number, then the original core can be divided into either $0.5 * n$ or n small cores. It is worth noting that two adjacent maximum mutual inductance points occur between a distance of $l_o/2n$ with the amplitude value of M_o (either sign: positive or negative).

In addition, as the multiple transmitter windings of the proposed system are placed close to each other, there will be crossing mutual inductance between them. These cause higher currents stresses through the inverter switches as well as resonant components of L_{ia} and C_{ia} and have destructive effects on the inverter switches [28]. To alleviate this effect, a small capacitor is connected in series with each transmitter to cancel the induced voltages caused by the crossing coupling (as shown in Fig. 9).

Fig. 5 Layout configuration of (a) Conventional single transmitter [12] (b) Top view and (c) Side view of the proposed multiple transmitters

C. Different winding methods to reduce copper wire length

In order to reduce the cost of the system implementation as well as the associated losses, the transmitter can be configured in different layouts to reduce the required copper. As an example, this Section presents two different winding methods for the proposed multiphase transmitters. The first winding method shown in Fig. 6a is to wind the copper wire from the first to the end poles while the second method is to wind the copper wire around individual poles, which is presented in Fig. 6b. Although the proposed transmitter presented in Section II.C uses three windings, however, for simplicity only one winding is illustrated in Fig. 6(a) and (b) in this section. The pole's length in this Section is of $(\frac{l_o}{2} - d_p)$ which consists of three smaller poles.

For the first and second winding method, the total lengths of copper wire l_1 and l_2 are given in (3) and (4) respectively:

$$l_1 \approx (l_o - 2d_p)N_{tu}N_{po} + 2N_{tu}(N_{po} - 1) \sqrt{d_p^2 + \frac{w_p^2}{4}} \quad (3)$$

$$l_2 \approx (l_o - 2d_p)N_{tu}N_{po} + t_p(N_{tu} - 1)N_{po} + 2(N_{po} - 1) \sqrt{d_p^2 + \frac{w_p^2}{4}} \quad (4)$$

Where N_{po} is number of poles and N_{tu} is number of turns.

Therefore, the required copper for the proposed transmitter is calculated using (3) and (4) and presented in Fig. 6(c) and (d). Even though the difference is not considerable if N_{po} and N_{tu} are of small values, however the second method saves up to 10% as compared to the first arrangement when $N_{po} = N_{tu} = 8$. Therefore, second method has been chosen to implement the proposed transmitter.

Fig. 6 Copper winding methods for the proposed multiphase transmitters (a) First method, (b) Second method. Comparison of the total copper wire lengths for three-phase system (c) First method (d) Second method.

D. Conditions of achieving constant output power

The open-circuit induced voltage V_s on the receiver can be expressed by:

$$V_s = j\omega \sum_{i=1}^n M_{is}(x)I_i \quad (5)$$

Where I_i is the current through transmitter i and $M_{is}(x)$ is the mutual inductance between transmitter i and the receiver.

Assuming the inverter produces equal voltages (i.e. with phase shift of $\frac{2\pi}{n}$) across each resonant tank, the currents through transmitter windings can be expressed by:

$$\begin{cases} I_1 = I_o e^{j0} \\ I_2 = I_o e^{j\frac{-2\pi}{n}} \\ \vdots \\ I_i = I_o e^{j(i-1)\frac{-2\pi}{n}} \\ \vdots \\ I_n = I_o e^{j(n-1)\frac{-2\pi}{n}} \end{cases} \quad (6)$$

Consequently, V_s can be further expressed as follows:

$$|V_s| = \omega \left| \sum_{i=1}^n M_o \sin \left(\frac{2\pi x}{l_o} - (i-1) \frac{2\pi}{n} \right) I_o e^{j(i-1) \left(\frac{-2\pi}{n} \right)} \right| \quad (7)$$

$$|V_s| = \omega M_o I_o *$$

$$\left| \sum_{i=1}^n \left\{ \frac{e^{j \left[\frac{2\pi x}{l_o} - (i-1) \frac{2\pi}{n} \right]} - e^{-j \left[\frac{2\pi x}{l_o} - (i-1) \frac{2\pi}{n} \right]}}{2j} \right\} e^{j(i-1) \left(\frac{-2\pi}{n} \right)} \right|$$

$$|V_s| = \frac{\omega M_o I_o}{2} \left| \sum_{i=1}^n \left[e^{j \frac{2\pi x}{l_o} - j(i-1) \frac{4\pi}{n}} - e^{-j \frac{2\pi x}{l_o}} \right] \right| \quad (8)$$

$$|V_s| = \frac{\omega M_o I_o}{2} \left| e^{j \frac{2\pi x}{l_o}} \sum_{i=1}^n e^{-j(i-1) \frac{4\pi}{n}} - e^{-j \frac{2\pi x}{l_o}} \right|$$

Where:

$$\sum_{i=1}^n e^{-j(i-1) \frac{4\pi}{n}} = e^{0j} + e^{-j \frac{4\pi}{n}} + e^{-j \frac{8\pi}{n}} + e^{-j \frac{12\pi}{n}} + \dots + e^{-j \frac{(n-1)4\pi}{n}} \quad (9)$$

$$= 1 + Z_1 + Z_2 + Z_3 + \dots + Z_{n-1}$$

$$\text{In which: } \begin{cases} Z_1 = e^{-j \frac{4\pi}{n}} \\ Z_2 = e^{-j \frac{8\pi}{n}} = Z_1^2 \\ Z_3 = e^{-j \frac{12\pi}{n}} = Z_1^3 \\ \vdots \\ Z_{n-1} = e^{-j \frac{(n-1)4\pi}{n}} = Z_1^{n-1} \end{cases}$$

Therefore, the sum in (8) is rewritten as (10) with attenuation that $Z_1^n = e^{-j \frac{4\pi}{n} n} = \cos(-4\pi) + j \sin(-4\pi) = 1$

$$\sum_{i=1}^n e^{-j(i-1) \frac{4\pi}{n}} = 1 + Z_1 + Z_2 + Z_3 + \dots + Z_{n-1} \quad (10)$$

$$= 1 + Z_1 + Z_1^2 + Z_1^3 + \dots + Z_1^{n-1}$$

$$= \frac{1 - Z_1^n}{1 - Z_1} = 0$$

Considering $\left| e^{-j \frac{2\pi x}{l_o}} \right| = 1$, the final expression of $|V_s|$ is given as:

$$|V_s| = \frac{\omega M_o I_o}{2} \left| 0 - n e^{-j \frac{2\pi x}{l_o}} \right| = \frac{n \omega M_o I_o}{2} \left| e^{-j \frac{2\pi x}{l_o}} \right| \quad (11)$$

$$= \frac{n \omega M_o I_o}{2}$$

At this point, the proposed system consisting of multiple transmitter windings can be simplified as only one transmitter combining with receiver as showed in Fig. 7. In the simplified model, M can be seen as the equivalent mutual inductance between primary and secondary side which is defined as a half of multiplication between the number of transmitters n and amplitude of each mutual inductances M_o . We also

assume that self-inductance of each transmitter is the same so that the equivalent transmitter has its inductance value of L_p , in which $L_p = L_1 = L_2 = \dots = L_n$. Finally, the equivalent transmitter current amplitude I_p is equal to I_o so that V_s can be expressed as Eq. (11). It is also noted that the phase of I_p does not matter here and constant $|V_s|$ can be attained with any phase value. The equivalent model helps to simplify required calculation and design procedure. According to (11), the induced voltage is constant regardless of receiver's position if adopting the transmitter's layout with expression of mutual inductances in (2) and condition for transmitter's current is shown in (6). Higher value of $|V_s|$ and corresponding power are realized with increase in the number of phase n .

Fig. 7 (a) Model of the proposed multiple transmitters (only coupled-coil part), (b) the simplified form of model in Fig. 7(a).

III. ILLUSTRATIVE DESIGN EXAMPLE OF THREE-PHASE WDC SYSTEM

The proposed multi-phase system is validated with an illustrative example based on a three-phase system in this Section. The circuit configuration is shown in Fig. 8 in which a conventional three-phase inverter is adopted to the primary side. The proposed layout in Fig. 5 is applied with $n = 3$ and the three transmitters are shown in Fig. 12. Mutual inductance distributions are depicted in Fig. 9 and phase difference between two consecutive transmitter is set to $2\pi/3$.

Fig. 8. The circuit configuration of three-phase dynamic charging system

Fig. 9. Spatial distribution of mutual inductances of three-phase transmitters system in Fig. 12

Therefore, mutual inductances equations are shown in (12).

$$\begin{cases} M_{1s}(x) = M_o \sin\left(\frac{2\pi x}{l_o}\right) \\ M_{2s}(x) = M_o \sin\left(\frac{2\pi x}{l_o} - \frac{2\pi}{3}\right) \\ M_{3s}(x) = M_o \sin\left(\frac{2\pi x}{l_o} - \frac{4\pi}{3}\right) \end{cases} \quad (12)$$

Firstly the output voltage of three-phase inverter can be depicted as (13):

$$\begin{cases} V_{12} = \frac{2\sqrt{3}}{\pi} V_{in} e^{j0} \\ V_{23} = \frac{2\sqrt{3}}{\pi} V_{in} e^{j(-\frac{2\pi}{3})} \\ V_{31} = \frac{2\sqrt{3}}{\pi} V_{in} e^{j(-\frac{4\pi}{3})} \end{cases} \quad (13)$$

To simplify control requirements of transmitter currents I_i ($i = \overline{1,3}$), resonance of L_{ia} and C_{ia} ($i = \overline{1,3}$) is adopted at ω_o as shown in (14) to keep a constant amplitude of the transmitter's current. Therefore, I_i are expressed in (15).

$$\omega_o^2 = \frac{1}{L_{ia} C_{ia}} \quad (i = \overline{1,3}) \quad (14)$$

$$\begin{cases} I_1 = V_{12} \omega_o C_{1a} e^{j(-\frac{\pi}{2})} = \frac{2\sqrt{3}}{\pi} V_{in} \omega_o C_{1a} e^{j(-\frac{\pi}{2})} = I_o e^{j(-\frac{\pi}{2})} \\ I_2 = V_{23} \omega_o C_{2a} e^{j(-\frac{\pi}{2})} = \frac{2\sqrt{3}}{\pi} V_{in} \omega_o C_{2a} e^{j(-\frac{7\pi}{6})} = I_o e^{j(-\frac{7\pi}{6})} \\ I_3 = V_{31} \omega_o C_{3a} e^{j(-\frac{\pi}{2})} = \frac{2\sqrt{3}}{\pi} V_{in} \omega_o C_{3a} e^{j(-\frac{11\pi}{6})} = I_o e^{j(-\frac{11\pi}{6})} \end{cases} \quad (15)$$

Fig. 10 presents the primary LCC tank considering magnetic interferences between three transmitters. The total induced voltages within the transmitters 1, 2 and 3 namely V_{T1} , V_{T2} and V_{T3} which are caused by the other transmitters are given by (16). It is noted that $M_{12} = M_{21}$, $M_{13} = M_{31}$ and $M_{23} = M_{32}$. These crossing induced voltages do not affect the receiver's induced voltage V_s and output power P_o but cause higher current stresses on inverter and the resonant components L_{ia} and C_{ia} .

$$\begin{cases} V_{T1} = j\omega_o M_{12} I_2 + j\omega_o M_{13} I_3 \\ V_{T2} = j\omega_o M_{21} I_1 + j\omega_o M_{23} I_3 \\ V_{T3} = j\omega_o M_{31} I_1 + j\omega_o M_{32} I_2 \end{cases} \quad (16)$$

To eliminate the crossing induced voltages on each transmitters, compensated capacitors C_{com1} , C_{com2} and C_{com3} are added into each transmitter producing a reversed voltage against V_{T1} , V_{T2} and V_{T3} . From experimental

measurement, all crossing mutual inductances have the same negative value because of the layout's structure. Therefore, we can set $M_{12} = M_{21} = M_{13} = M_{31} = M_{23} = M_{32} = M_{cros}$ in which M_{cros} is a negative number. Further derivations have been made with (16) and V_{T1} , V_{T2} and V_{T3} can be expressed depending on M_{cros} as follows.

$$\begin{cases} V_{T1} = j\omega_o M_{cros} I_1 e^{j\frac{2\pi}{3}} + j\omega_o M_{cros} I_1 e^{j\frac{4\pi}{3}} \\ \quad = -j\omega_o M_{cros} I_1 = j\omega_o |M_{cros}| I_1 \\ V_{T2} = j\omega_o M_{cros} I_2 e^{j\frac{-2\pi}{3}} + j\omega_o M_{cros} I_2 e^{j\frac{2\pi}{3}} \\ \quad = -j\omega_o M_{cros} I_2 = j\omega_o |M_{cros}| I_2 \\ V_{T3} = j\omega_o M_{cros} I_3 e^{j\frac{-4\pi}{3}} + j\omega_o M_{cros} I_3 e^{j\frac{-2\pi}{3}} \\ \quad = -j\omega_o M_{cros} I_3 = j\omega_o |M_{cros}| I_3 \end{cases} \quad (17)$$

Voltages on compensated capacitors (i.e. V_{Ccom1} , V_{Ccom2} , V_{Ccom3}) should be equal in magnitude and reversed in phase with (V_{T1}, V_{T2}, V_{T3}) to effectively cancel these above crossing induced voltages.

$$\begin{cases} V_{Ccom1} = \frac{1}{j\omega_o C_{com1}} I_1 = -V_{T1} \\ V_{Ccom2} = \frac{1}{j\omega_o C_{com2}} I_2 = -V_{T2} \\ V_{Ccom3} = \frac{1}{j\omega_o C_{com3}} I_3 = -V_{T3} \end{cases} \quad (18)$$

Therefore, three compensated capacitors can be calculated based on (19). These capacitors have the same value and depend only on crossing mutual inductance and operating frequency.

$$\begin{cases} C_{com1} = \frac{1}{\omega_o^2 |M_{cros}|} \\ C_{com2} = \frac{1}{\omega_o^2 |M_{cros}|} \\ C_{com3} = \frac{1}{\omega_o^2 |M_{cros}|} \end{cases} \quad (19)$$

Fig. 10. Primary LCC tanks considering magnetic interferences

Finally, C_{com1} , C_{com2} and C_{com3} can be integrated into C_{1b} , C_{2b} and C_{3b} to become a single serial capacitor for each

phase. These serial capacitors are determined by (20). In Section V, experimental results are presented to compare between including and not including these compensated capacitors.

$$\begin{cases} C_{1b} = \frac{1}{\omega_o^2} \left(\frac{1}{L_1 - L_{1a}} + \frac{1}{|M_{cros}|} \right) \\ C_{2b} = \frac{1}{\omega_o^2} \left(\frac{1}{L_2 - L_{2a}} + \frac{1}{|M_{cros}|} \right) \\ C_{3b} = \frac{1}{\omega_o^2} \left(\frac{1}{L_3 - L_{3a}} + \frac{1}{|M_{cros}|} \right) \end{cases} \quad (20)$$

In an dynamic charging system, constant current charge is preferable than constant voltage charge because EV battery's state of charge (SoC) is rarely reached to fully charged [4]. Therefore another LCC tank is constructed at secondary side tank to produce output current source as shown in Fig. 11. Capacitors C_{1s} is connected in series with L_s while C_{2s} and L_{1s} are additional parallel capacitor and inductor for each secondary tank. According to equivalent circuit on Fig. 11, DC output current is depicted as (21) if the resonant tank at secondary side is designed as (22) and (23). In (21), M_{is} and I_i ($i = \overline{1,3}$) can be found in (12) and (15) respectively.

$$I_{bat} = \frac{2}{\pi} \omega_o^2 C_{2s} \left| \sum_{i=1}^3 M_{is} I_i \right| = \frac{3 \omega_o^2 C_{2s} M_o I_o}{\pi} \quad (21)$$

$$= \frac{6\sqrt{3}}{\pi^2} \omega_o^3 C_{ia} C_{2s} M_o V_{in} \quad (22)$$

$$j\omega_o L_s + \frac{1}{j\omega_o C_{1s}} + \frac{1}{j\omega_o C_{2s}} = 0 \quad (22)$$

$$j\omega_o L_{1s} + \frac{1}{j\omega_o C_{2s}} = 0 \quad (23)$$

Fig. 11. AC Equivalent circuit of secondary circuit

Combining (14), (20), (22) and (23), the equation of the resonant frequency ω_o can be determined as (24), in which ω_o is regardless of both load condition (R_o) and mutual inductances between transmitters and receiver (M_{1s} , M_{2s} and M_{3s}).

$$\omega_o^2 = \frac{1}{L_{ia} C_{ia}} = \frac{1}{C_{ib}} \left(\frac{1}{L_i - L_{ia}} + \frac{1}{|M_{cros}|} \right) = \frac{1}{L_{1s} C_{2s}} \quad (24)$$

$$= \frac{1}{C_{1s}} \left(\frac{1}{L_s - L_{1s}} \right) \quad (i = \overline{1,3})$$

The DC output voltage and power are presented in (25) and (26) adopting R_o which is equivalent battery's impedance. It needs to emphasize that if switching frequency ω_o and secondary side capacitor C_{2s} are assumed to be constant, then each load output power depends only on the multiplication $M_o I_o$. Obviously, it can be seen that P_o is constant and regardless to receiver's position.

$$V_{bat} = I_{bat} R_o = \frac{6\sqrt{3}}{\pi^2} \omega_o^3 C_{ia} C_{2s} M_o V_{in} R_o \quad (25)$$

$$P_o = I_{bat}^2 R_o = \left(\frac{6\sqrt{3}}{\pi^2} \omega_o^3 C_{ia} C_{2s} M_o V_{in} \right)^2 R_o \quad (26)$$

IV. OPTIMIZED DESIGN OF MAGNETIC COUPLER

A. Design of Three-phase Transmitters

This section presents the transmitters magnetic design aiming for maximum output power of 3 kW (scale down value). Several simulations are conducted using MAXWELL 3D software to select the optimal parameters for transmitters, which are presented in Fig. 12. They consists of the ferrite pole sizes (thickness t_p , distance d_p , height h_p , length l_p and plate width w_p), ferrite bottom sizes (width w_b and thickness t_b) as well as number of turns for three transmitter's windings N_1 , N_2 and N_3 . As three transmitter are identical, we consider only one transmitter in simulation model which is shown in Fig. 13a for simplicity. The purpose of simulation is to determine the optimal parameters so that the maximum of magnetic flux density B can be attained at the center of every three poles (as shown in Fig. 12). With an appropriate receiver design, maximum of B also means maximum coupling coefficient and the highest power delivery as well as high efficiency. Simultaneously, we try to minimize the ferrite core's volume and cable's length but the system still needs to satisfy the power delivery requirement and avoid the core's saturation. In order to produce a homogeneous magnetic flux along driving direction, number of turns should be selected equally for three windings and all three transmitter's currents are set as maximum of 20 A. The maximum air-gap between transmitter and receiver are selected as 15 cm.

Fig. 12 Design parameters of three-phase primary transmitters

Thickness and width of pole and bottom plates are selected to prevent saturation problems. Thereout, thickness values have very limited effect again magnetic field level. Fig. 13b presents distribution of magnetic flux on ferrite core with bottom thickness (t_b) = pole plate thickness (t_{pp}) = 0.5 cm and pole thickness (t_p) = 1.5 cm. According to that the maximum magnetic flux density of 0.152 T occurs around the crossing winding areas on the core bottom. In this work, the ferrite core of PC47 by TDK Electronics is selected and their saturated magnetic flux density B_s is founded as 0.54 T at 25 °C [29]. Therefore, we can guarantee of no saturation for the primary core with output power level of 3 kW and transmitter current of 20 A. Note that it is possible to implement smaller thickness values, however, values below 0.5 cm are difficult to achieve practically due to the physical cutting process of ferrite. The width is selected as small as possible to reduce the width of the rail, then the constructed cost and pedestrian

EMI contamination can be reduced. Here, the widths of both bottom and pole plate are selected as 12 cm.

The remaining three design parameters are : pole distance (d_p), pole length (l_p) and number of turns (N). The relationship between d_p and l_p can be found in (27) with l_o is fixed based on system's requirements, which is 2.4 meters in this work. Therefore, only d_p needs to be found and l_p will follow according to (27). Fig. 13c presents the relationship between normalized magnetic coupling against pole distance d_p . It is noted that the measurement point are located right above the center of each group of 3 ferrite poles as shown in Fig. 12. When d_p varies in a range from 5 to 22.5 cm, then normalized value of magnetic flux density reduces accordingly. It can be seen that the maximum magnetic flux density value can be achieved at pole distances from 5 to 7.5 cm. This fact can be explained by deriving the magnetic flux equations depending winding's width and length [30]. Value d_p of 7.5 cm is selected in this work because 5-cm value creates a quite small space which may not enough incooperate a large number of winding turns.

$$d_p + l_p = \frac{l_o}{6} \quad (27)$$

Fig. 13d demonstrates how magnetic flux density B_l varies according to number of turns for one single winding N_l . Theoretically, B_l becomes linear with N_l and while number of turns increases from 2 to 6, B_l rises from 31 to 93 μT as a result. Higher number of turns result in higher power delivery capability, however, it also increases the total copper wire's length, transmitter inductance and pedestrian EMI contamination. A number of 4 turns is finally selected and Litz wire (600 strands-AWG 38) is adopted to reduce conduction losses on windings. After several trial and errors in both simulations and experiments final parameters are selected and depicted in TABLE I to meet with system requirements.

TABLE I
PARAMETER VALUES OF PRIMARY TRANSMITTERS

t_b	Bottom thickness	0.5 cm
w_b	Bottom width	12 cm
t_p	Pole thickness	1.5 cm
d_p	Pole distance	7.5 cm
h_p	Pole height	4 cm
l_p	Pole length	12 cm
w_p	Pole plate width	12 cm
t_{pp}	Pole plate thickness	0.5 cm
d	Air gap	15 cm
	Number of ferrite poles (individual pole)	12
	Number of winding	3
N_1, N_2, N_3	Number of turns for each winding	4
$2l_o$	Total transmitter's length	2.4 meters
I_1, I_2, I_3	Transmitter's currents (maximum)	20A

Fig.13 (a) Simulation model of the proposed transmitters with bottom thickness $t_b = 0.5$ cm, bottom width $w_b = 12$ cm, pole thickness $t_p = 1.5$ cm, pole height $h_p = 4$ cm, pole length $l_p = 12$ cm, pole plate width $w_p = 12$ cm, pole plate thickness $t_{pp} = 0.5$ cm. Pole distance d_p and number of turns N_l are varied in simulation (b) Contribution of magnetic flux density on transmitter's ferrite core with $N_l I_l = 80$ A (c) Normalized magnetic flux density depending on the pole distance with $N_l I_l = 80$ A (d) Magnetic flux density depending on number of turns with $d_p = 7.5$ cm.

B. Design of receiver

After obtaining transmitter's parameters, this section provides the design for receiver's parameters, which include receiver's length l_r , width w_r and number of turns N_s . The most sensitive parameter as regards the coupling coefficient k is the receiver's length l_r . Fig. 14a shows result of how coupling coefficient against l_r with transmitter parameters on Table I. As shown in Fig. 14a, normalized coupling value increases from 0.05 at $l_r = 30$ cm and reaches to the highest value when $l_r = 60$ cm (equal to $l_o/2$). If l_r goes higher than 60 cm, then k drops significantly to nearly zero at 100 cm. It can be explained by the fact that when $l_r < 60$ cm, then receiver's winding cannot cover entirely the align transmitter winding, therefore, k is always lower than the highest value. For $l_r > 60$ cm, then receiver is also received magnetic fluxes from other two adjacent transmitter windings which have reversed current directions and create opposite magnetic fluxes as compared to the align winding. Totally, there is a flux reduction on receiver when $l_r > 60$ cm as compared to the highest case. More seriously, the total flux on receiver can be

cancelled completely if the total flux coming in equals to the flux coming out receiver. It results in no coupling between two sides and no induced voltage in the receiver winding.

Fig. 14b and 14c provides simulation results of coupling coefficient under different receiver's width and lateral misalignments values. The coupling value k needs to be high as much as possible at different lateral misalignment conditions. Fig. 14b provides coupling values regarding different receiver's widths w_r under no misalignment. It shows that k increases from 0.062 to nearly 0.1 when w_r rises from 10 to 30 cm. However, if w_r continues to increase up to 70 cm then k will slightly reduce. Under lateral misalignment conditions, it can be investigated in Fig. 14c that how reduction of coupling regarding different receiver's width values. If misalignment level equals to 30 cm, then k will reduce to nearly 0%, 30% and 60% associating with $w_r = 50$, 60 and 70 cm, respectively. Fundamentally, higher w_r values enhances coupling and output power under lateral misalignment, however, coupling coefficient is also reduced when increasing w_r at perfect align condition. Moreover, w_r is also limited by installation space at bottom of vehicle. Considering all above factors, w_r is selected as 60 cm in this work so that output power level can remain at 30% of maximum power under 30 cm of lateral misalignment. After selecting receiver's dimensions of $l_r = w_r = 60$ cm, then number of turns N_s can be estimated as 8 turns based on the required output voltage (i.e. 450 V) and operating frequency (i.e. 85 kHz).

Fig. 14. Simulation on Maxwell 3D of normalized coupling coefficient with transmitter's parameters from Table I against (a) different receiver's length (b) different receiver's width at no misalignment (c) different lateral misalignment conditions.

V. EXPERIMENTAL VALIDATION AND DISCUSSIONS

In order to validate the proposed WDC, a scaled down prototype of 3-kW system is developed in lab as portrayed in Fig. 15. The parameters of the transmitter and receiver are tabulated in Table I. As shown in Fig. 15(b), the transmitter windings are arranged in three layer, which is isolated from each other using plastic sheets. Furthermore, Kapton-tape-shield Litz wire is used to further improve the isolation windings different loops of the same phase's winding. The receiver shown in Fig. 15(c) is constructed using 5-long-bar ferrite core and 8-turn Litz wires. As highlighted in in Section IV, the dimension of the receiver is 60cm x 60cm while the maximum air-gap between the transmitter and receiver is tested at 15 cm. The three-phase inverter is built using SiC MOSFET, C2M0025120D from Cree while SiC-Schottky-diodes, HFA50PA60 are selected for the rectifier at the output side. Polypropylene film capacitors are chosen for the resonant tank owing to their low equivalent series resistance and high current carrying capability. The specifications of the proposed WDC are summarized in Table II along with the parameters of resonant circuit. Please note that the compensated capacitors C_{com1} , C_{com2} , C_{com3} are adjusted from 76.22 nF in simulation to 66.5 nF in experiment for achieving Zero-Voltage-Switching (ZVS) of all Inverter's switches.

Fig. 15. (a) Experimental test-rig of the proposed WDC system (b) Three-phase transmitter (c) Receiver

TABLE II
SPECIFICATION AND PARAMETERS OF THE PROPOSED WDC

Symbol	Parameter	Simulated values	Practical values
V_{in}	Input voltage	400 V	400 V
I_o	Rated output current	6.5 A	6.45 A
P_o	Rated output power	3 kW	2.91 kW
L_1, L_2, L_3	Self-inductance of transmitters	182 μ H	178.6 μ H, 180.2 μ H, 179.6 μ H
L_s	Self-inductance of receiver	118 μ H	120.5 μ H
M_o	Maximum mutual inductance	12.6 μ H	12.48 μ H
M_{12}, M_{13}, M_{23}	Crossing mutual inductances	-46 μ H	-44.5 μ H, -45.8 μ H, -47.1 μ H
L_{1a}, L_{2a}, L_{3a}	Primary additional inductors	37 μ H	36.2 μ H, 37.3 μ H, 37.5 μ H
C_{1a}, C_{2a}, C_{3a}	Primary parallel capacitors	95 nF	95.2 nF, 95.1 nF, 95.2 nF
C_{1b}, C_{2b}, C_{3b}	Primary series capacitors (without compensated capacitors)	24.2 nF	24.2 nF
$C_{com1}, C_{com2}, C_{com3}$	Primary compensated capacitors	76.22 nF	66.5 nF
L_{1s}	Secondary additional inductors	35 μ H	36.2 μ H
C_{2s}	Secondary parallel capacitors	100 nF	100.6 nF
C_{1s}	Secondary series capacitors	42.2 nF	42.5 nF
f	Switching frequency	85 kHz	85.05 kHz

Fig. 16 illustrates key operating waveforms at the primary side circuit. Specifically, Fig. 16(a) depicts the Gate PWM signals, while the output voltages V_{12} , V_{23} and V_{31} can be seen in Fig. 16(b). The corresponding transmitter currents I_1 , I_2 and I_3 are shown in Fig. 16(c), where they have the same amplitude and a phase-shift of 120 degrees apart from each other. Fig. 16(d) depicts the measured voltage V_{ds} and current I_{ds} of the MOSFET switches (i.e. S_1 , S_3 and S_5) at the rated power. As it can be noticed that turn-on Zero Voltage Switching (ZVS) is achieved for all switches, which leads to a significant reduction in switching losses of the inverter.

Fig. 16. Key waveforms at primary side: (a) Gate signal of three-phase inverter, (b) Inverter's voltage (c) Transmitter's currents (d) Inverter switches' voltage V_{ds} and current I_{ds} under the rated load condition (with the receiver's displacement of 60 cm)

Fig. 17 further demonstrates the operation of the proposed system under different loading conditions (i.e. 17 Ω , 40 Ω and 70 Ω) while receiver is in the moving process. Moreover, the

harvested output power at different positions along the driving direction x as portrayed in Fig. 18. It should be noted that x starts only from 15cm and ends at 225cm as only cases when the whole receiver is positioned on the transmitter windings are considered.

The power fluctuation factor is defined as $\Delta P = \left(\frac{P_{max} - P_{min}}{P_{max}} \right) * 100\%$. For this particular experiment, ΔP are found to be 4.9 %, 5.6 % and 5.1 % with the load of 17, 40 and 70 Ω , respectively.

Fig. 17. Experimental waveforms with different load while receiver moving along driving direction (a) $R_o = 17 \Omega$, (b) $R_o = 40 \Omega$ and (c) $R_o = 70 \Omega$

Fig. 18. Measured output power with different load conditions and different receiver positions along the driving direction

Fig. 19 compares the efficiency performance between conventional single-phase system (one transmitter) and the proposed three-phase one. The DC-to-DC efficiency of the WDC system was measured by a YOKOGAWA WT1600 Power Analyzer. During both tests, receiver only moves along the driving direction from displacement x of 60 cm to 120 cm (be equivalent to $l_o/2$) with the reference point at the beginning of transmitter. Resistive load is adjusted so that output powers are 1 and 3 kW for single and three-phase system, respectively. The results indicate that efficiency drops significantly to nearly zero in the single-phase system at the points of 60 and 120 cm (i.e. null-mutual-inductance points). Efficiency then increases when receiver moves either from 70 to 90 cm or in reverse direction from 110 to 90 cm and reaches to maximum of 89.6% at 90 cm. Differentiating from single-phase system, the proposed three-phase offers a nearly constant efficiency along the driving direction as output power are kept unchanged when receiver moving. At the rated power of 3 kW then efficiency achieve around 87.8%.

Fig. 19. Efficiency comparison between single-phase and three-phase system

The performance of proposed WDC is further investigated under lateral misalignment with different air-gap, h of 5, 10 and 15 cm. As demonstrated in Fig. 20(a), the output power reduces with higher lateral misalignment between the receiver and the transmitter. The output power also drops as the air-gap between the receiver and transmitter increases. Under testing conditions of 15 cm air-gap and 30-cm misalignment, then the output power reaches to 29% of the maximum power of 3 kW. Furthermore, Fig. 20(b), illustrates the system efficiency across a wide range of misalignment as well as for different air gaps. As expected, higher efficiency attained with smaller misalignment as can be seen from the same figure.

(a)

Fig. 20 System performance under different lateral misalignments and air-gap value at rated power ($R_o = 70 \Omega$) (a) Normalized output power, (b) Measured output efficiency.

To realize the effect of the crossing couplings between the multiple windings on Inverter's switches and resonant components, we conducted two different tests with and without compensated capacitors C_{com1} , C_{com2} and C_{com3} . In both tests, the temperature of the semiconductor devices of the primary inverter is measured using FLIR E5 thermal imaging camera. As it can be observed from Fig. 21, before adding the compensating capacitors, the temperature on the power MOSFET reaches 123°C with only 0.2kW output power and 300V input DC. However, once the compensating capacitors are added, the temperature does not exceeds 105°C at full load (i.e. 3kW). Hence, the crossings mutual inductances in multiphase system must be carefully compensated in order to attain the maximum power transfer without exceeding the thermal limits.

Fig. 21 Temperature comparison of MOSFET inverter under different test conditions (a) without compensated capacitors when $P_o = 200 \text{ W}$, (b) with compensated capacitors when $P_o = 2960 \text{ W}$.

For completeness, the effect of the crossing mutual inductances on the primary side components is also investigated. Table III summarises the current stress on the inverter switches (S_1 , S_2 , S_3 , S_4 , S_5 , S_6), the additional inductors (L_{1a} , L_{2a} , L_{3a}) and the parallel capacitors (C_{1a} , C_{2a} , C_{3a}), with and without compensating capacitors. It should be noted that, the current through the series capacitors and the transmitters will not be effected by the crossing mutual inductances. The temperature of power MOSFETs can be very high without compensating capacitors, therefore only simulation results are provided in this case. The parameters of mutual inductances M_{1s} , M_{2s} and M_{3s} in simulation are adjusted to match with their experimental parameters, (i.e. $M_{1s} = -6.18 \mu\text{H}$, $M_{2s} = 12.5 \mu\text{H}$, $M_{3s} = -6.27 \mu\text{H}$). Load resistances in both cases are set as 45Ω . The difference of current stresses is clearly depicted in TABLE III between two cases. RMS currents on switches and components go up to very high value when compensated capacitors are not included. Oppositely, component's currents reduce significantly once including compensated capacitors which helps to suppress crossing induced voltages on each transmitter.

TABLE III
COMPARISON OF RMS CURRENT STRESSES ON PRIMARY COMPONENTS

Components	Excluding compensated capacitors (simulation results) [A]	Including compensated capacitors (experimental results) [A]
S_1	24.3	1.78
S_2	24.4	7.13
S_3	29.3	6.98
S_4	24.3	1.78
S_5	24.4	7.13
S_6	29.3	6.98
L_{1a}	22.6	3.23
L_{2a}	22.7	6.46
L_{3a}	18.4	3.56
C_{1a}	38.4	18.75
C_{2a}	38.2	18.87
C_{3a}	27.3	14.45

Finally, Electromagnetic Irradiation (EMI) surrounding the transmitter coils are measured under different distances far away from the coil's centre d_l and different displacement x along the driving direction as shown in Fig. 22. The Gauss Meter SPECTRAN NF-5030 is used as the measurement sensor. According to ICNIRP exposure limits, then the measured flux density is below $27 \mu\text{T}$ at the positions of 0.8 meters far away from the transmitter.

Fig. 22. Magnetic field measurement at different horizontal conditions

VI. SYSTEM COMPARISON

In order to demonstrate the benefits of the proposed multiphase WDC system, a comparison with other similar WDC systems [12, 13, 14, 21, 22] that have the advantage of low EMI and narrow transmitter's width is carried out. The first feature to consider is of output power capability. For the sake of a sensible comparison, the following assumptions are considered: 1) output power is normalized by having the same transmitter current, coupling factor, switching frequency and load resistance for each system. 2) the same compensation

circuit at the secondary side of each system. Please note that output power in [14], [21] depends on the number of receiver's windings (phases). The case of two parallel receiver phases is reported here. As it can be seen from Table IV, the proposed WDC system offers a constant and high output power (i.e. 2.25 times higher as compared to other systems). This is a clear advantage when high power demand is required.

Other comparisons can be found from Table IV. Apart from [22], other systems are incapable of producing constant power, however two inverters are required in [22] as opposed to only one inverter in the proposed WDC. Although the output power is different from system to another, however the proposed system achieves higher efficiency, when tested at a full load. This apparently because the system is optimized to achieve the highest magnetic coupling and also the utilization of the state of the art of SiC MOSFETs, which have the low on-state resistance.

Interestingly, the control of the secondary side is independent from the primary one, which is not the case with other reported systems. This makes the proposed system more attractive when multiple vehicle charging is needed. However, in the case of other systems, mutual inductances and load resistances information are essential, which makes infeasible with multiple EVs in motion.

On the other hand, transmitter's copper lengths between different systems are calculated and compared for one kilometre of dynamic charging road. It should be noted that the proposed system requires more copper wire at the transmitter side. However, this can be justified with the extra benefits that are gained. Nevertheless, different winding arrangement can be considered to reduce the required copper length as discussed in Section II. C. Finally, the comparison is summarised in Table IV.

TABLE IV
COMPARISON OF VARIOUS DWC SYSTEMS

	Ref. [12]	Ref. [13]	Ref. [22]	Ref. [14], [21]	The proposed one
Tested Output Power	35 kW	22 kW	250 W (Scale down)	10 kW	3 kW (can be increasable)
Operating frequency (kHz)	20	20	20	20	85
Maximum Efficiency (%)	74	71	-	-	87.6
Normalized output power versus the driving direction
Output power variation along driving direction	Not stable	Not stable	Stable	Not Stable	Stable
Number of Primary Inverter	1 (Single phase full-bridge)	1 (Single phase full-bridge)	2 (Single phase full-bridge)	-	1 (conventional three phase)
Transmitter's width (cm)	10 (I type core shape)	4 (S type core shape)	10 (dq I-type core shape)	10 (N type core shape)	12 (Multiphase I-type core)
Transmitter's copper length	8000 (1 transmitter's winding)	6000 (1 transmitter's winding)	16000 (2 transmitter's winding)	12000 (1 transmitter's winding)	20400 (3 transmitter's winding)
Receiver's dimension	80 cm x 100 cm (1 winding)	80 cm x 100 cm (1 winding)	90 cm x 100 cm (1 winding)	90 cm x 108 cm (2 windings in parallel*)	60 cm x 60cm (1 winding)
Air gap (cm)	20	20	15	20	15
Possibility of multiple vehicle charging	No	No	No	-	Yes
Load-dependent measurement	Yes	Yes	Yes	Yes	No

Note: - means no information available from the reference.

VII. CONCLUSION

This paper presented a constant and high output power WDC for EVs, utilizing multi-winding with alternative ferrite poles in primary side to effectively produce a constant coupling magnetic flux between transmitters and receiver. The detailed layout for transmitter along with the conditions of transmitter current are analysed and discussed in the paper.

An illustrative three-phase example of WDC demonstrated the feasibility of the proposed system. The effect of crossing mutual inductance is thoroughly analysed and a re-design of the series capacitors in LCC compensation tank is presented to overcome this effect. The paper also presented the magnetic design of the transmitter and the receiver that attain large coupling coefficient, and minimum loss of power under lateral misalignment with less ferrite core and copper cable.

The feasibility of the proposed system is experimentally validated using a 3 kW laboratory prototype, tested under different operating conditions. The power fluctuation factor at rated load can attain 5.6% while maximum efficiency reaches 87.8%. A detailed comparison with other similar works is conducted and the main advantages can be pointed out such as (1) achieving a constant and higher output power, (2) using less number of Inverters, (3) elimination of measurement load and (or) coupling from secondary side, (4) supporting the multiple EV charging without any complex control.

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