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Estimation of the Transferred Power in LCC Compensated Wireless Power Transmitters with the use of PWM-Synchronized Sampling Technique

Abstract—In this paper, a new technique of synchronized sampling to estimate the transferred power in wireless power transfer (WPT) systems is proposed. Defining demand factor as a criterion linking unknown receiver parameters to the transferred power, it can be estimated by measuring the front-end variables of the transmitter LCC compensating network. Moreover, with the use of this technique, one does not need to know how the mutual inductance or load varies. As the demand factor is directly linked to the transferred power, the estimation of the demand factor can be considered as an increasingly effective way to control the flow of power in wireless power transfer systems. To prove the validity of this approach, a WPT system equipped with an LCC network to compensate the transmitter, and a series capacitor to compensate the receiver, is experimentally built and tested, and the results show a close consistency between the measured demand factor and the transferred power.

Index Terms—LCC Compensator, Load Estimation, Power Estimation, Resonant Converters, Soft-Switching, Wireless Power Transfer Systems.

I. INTRODUCTION

WIRELESS POWER TRANSFER (WPT) systems have drawn attention for many years. Attractive features, such as safety of user, ease of use, absence of mechanical connectors, immunity to harsh environments (e.g. dirt, chemicals, snow), flexibility in the receiver position, and dynamic power transmission, make industries and academia increasingly explore a vast number of remarkable ideas [1]–[4]. In general, WPT research can be categorized into magnetic design of the WPT pads, transmitter and receiver compensation techniques, power electronic converters, and feedback and control techniques.

In addition to the aforementioned system level studies, identification of the receiver parameters and controlling the flow of power between them can be considered as one of the most important key concepts. For instance, in dynamic WPT where electric vehicles are charged while they move, the utility needs to meter the transferred power (energy) to charge the user according to the available energy tariff [5], [6]. Fast techniques of transferred power estimation is essential to control the transferred power to the moving receiver and to enhance the dynamics of the receiver circuitry [7], [8]. For this purpose, transmitter senses and controls the power so that the received power is smooth and levelled when the moving receiver interacts with the transmitter pad. Hence, the proposed PWM-synchronized sampling (PWMS2) power estimation technique could provide an efficient way of transferred power estimation necessary for the effective control of the power transfer process.

Fig. 1. Different techniques in wireless identification of WPT coils; (a) indirect, (b) direct transmissive, and (c) direct reflective approaches.

As power in WPT systems is transferred wirelessly, the information flow between such coils must be carried out without any physical connection. Therefore, wireless sensing of variables is introduced as a measure of identification of unknown parameters. The wireless sensing techniques can be categorized into two basic types, namely indirect and direct methods, as shown in Fig. 1. In the indirect approach, WPT sides communicate through additional wireless sensing equipment (for example, using techniques such as image processing [9], global positioning system (GPS) [10], [11], infrared signal [12], ultrasonic signal [13], or wireless power signal [14]), as shown in Fig. 1(a). Although some of these sensing techniques are extensively used in different applications such as position detection, they cannot be directly used to sense

the WPT variables such as the profile of transferred power for controlling purposes. To sense the transferred power profile, the sensing needs to be done **through the same or additional WPT pads (sensing pads)**, and the sensing pads need to create an inductance profile (mutual inductance and self-inductance) similar to the main WPT pads. However, sensing and power coils may not create identical inductance profiles and that can lead to an inaccurate estimation. The other problem that may occur as a result of using indirect identification is the delay in response. **As an example, Miller *et al.*, in [15], proposed a strategy that controls the flow of power using an indirect sensing technique (additional communication link to transfer the receiver-side information to the transmitter side). Although a novel control strategy is proposed and implemented in a comprehensive WPT system, due to an additional communication link, this method increases the complexity and delay in control.**

On the other hand, in direct approach, WPT coil measurements are directly used for identification. This can be done transmissively, by sensing the power transferred from one side and received from the other side, as shown in Fig.1(b), or reflectively, **by sending power from the transmitting side and measuring its reflection at the same side**, as shown in Fig.1(c). Identification in the reflective direct techniques can be carried out by observing the amplitude of the influenced (reflected) signal [16], [17] or its frequency content [18]–[20]. **In reflective measurement techniques, the presence of the receiver which is effectively coupled with the transmitter influences the characteristic of the system which can also be observed through the transmitter pads. In other words, the presence of the receiver provides a signal flow loop between the transmitter and receiver through which the reflected induced voltage of a current driven coil (or reflected induced current in a voltage drive coil) can be used to sense the receiver [21], [22].**

Reflective measurements are particularly attractive for primary-side (transmitter-side) control and stabilization as it does not need any additional communication link to read the information from the receiver. To stabilize and control the flow of power through a reflective measurement and without any additional communication link between transmitter and receiver, Huanjie *et al.*, [23], estimated the transmitter power by taking root-mean-square of the inverter output voltage and transmitter current in a series compensated transmitter. Although this technique is simple, it completely relies on the output voltage of inverter and the transmitter current phasors to have the same phase angle, which does not necessarily happen in reality due to the uncertainties in tuning the compensation circuitry.

The sensing can be carried out **intermittently** or continuously. In the first approach of the interrupted cycles of sensing, the WPT process is to be halted for a short period of time to identify the receiver parameters. In this direct approach, although the same WPT transmitter is used to sense the receiver, sensing interruptions can decrease the overall transferred power and it can create some undesirable transients in the transmitter-side [18]–[20]. Therefore, continuous techniques are proposed to observe the receiver through the same WPT transmitter with a minimum influence on the WPT

Fig. 2. Circuit diagram of a WPT system equipped with an LCC and series capacitor in transmitter and receiver coils respectively.

Fig. 3. Symmetric T-Bridge.

process [16], [17]. As a result, direct approaches are more reasonable to sense the flow of the transferred power. Different combinations of all these techniques can be used to extract more information from the wirelessly coupled coils. These techniques can effectively estimate the reflected impedance of the receiver coil seen from the transmitter coil. Although in some articles, the *estimation of mutual inductance* is the topic of concern, without knowing the load information of the far-end coil (*i.e receiver impedance*), the estimation of mutual inductance is not possible [24]–[26]. Therefore, another communication link or predetermined load information is needed for the estimation of mutual inductance.

In this article, by sampling of the LCC parallel capacitor voltage in the WPT system shown in Fig. 2, the transferred power to the receiver is estimated. A novel approach of PWM2 is used to estimate the transferred power. The studies are carried out by defining a new criterion termed as demand factor. To elaborate this technique, Section II describes how the transferred power can be estimated from a transmitter equipped with an LCC network. Section III is dedicated to analyze the influence of the detuned reactances on the operation of the proposed system. In Section IV, the influence of stray series resistances on the estimation is analyzed. Section V compares the numerical and experimental results to prove the validity of the proposed approach. Finally, in Section VI, a summary of the proposed technique of the power estimation is provided.

II. OPERATION PRINCIPLES OF THE SYSTEM

In this section, how the variation of mutual inductance influences the transmitter LCC compensator will be investigated. For this purpose, the reflection of the load impedance in a symmetric T-Bridge is explained. Then, using this concept, it is shown how the receiver load will be seen from the transmitter LCC input terminals.

Fig. 4. Equivalent circuit of the LCC-S compensated WPT system indicating impedances at different stages.

A. Symmetric T-Bridge

Symmetric T-Bridge is considered as a passive T-shape LC network, as shown in Fig. 3. It has identical magnitude of reactance in all its branches with the front-end and back-end impedances being complex conjugate of the mid-branch impedance. This configuration can be repeatedly seen in different places of WPT systems, such as coupled inductors and LCC compensators. Therefore, characterizing its operation will make the analysis of WPT systems easier. For this bridge, the load impedance Z_L seen from the input terminals of the bridge Z'_L is given in (1).

$$Z'_L = |Z_{12}|^2 / Z_L = X_{C/L}^2 / Z_L \quad (1)$$

B. Measuring the Demand Factor in an LCC Compensator

The proposed WPT system shown in Fig. 2 can be rearranged to form a cascaded symmetric T-bridge network as shown in Fig. 4. In this circuit, transmitter self-inductance L_{TT} is divided into two parts $L'_{TT} = L_S$ and $L''_{TT} = L_{TT} - L_S$ to make the symmetric T-bridges visible in the equivalent circuit. The component values, including the front-end inductance, the coil inductances, and the compensation capacitors, are chosen to comply with the following relation at the operating angular frequency of ω .

$$\omega = \frac{1}{\sqrt{L_S C_P}} = \frac{1}{\sqrt{L'_{TT} C_P}} = \frac{1}{\sqrt{L_{RR} C_R}} \quad (2)$$

To achieve a desirable level of soft-switching,

$$(K_{SSF} - 1)\omega L_S = \omega L''_{TT} - \frac{1}{\omega C_T} \quad (3)$$

where ω is the fundamental angular frequency of the source and K_{SSF} is the Soft-Switching Factor of the LCC Network. This factor is unity for zero phase switching (when the fundamental of v_{in} has the same phase angle with i_{in}) and is equal to $\pi^2/8$ for zero current switching [27] (when i_{in} is zero at the zero crossing moments of v_{in}). Next, in order to simplify the analysis, parasitic resistances of all the components, such as equivalent series resistances (ESRs) of L_S and power supply (R_S) and transmitter (R_T) are assumed to be negligibly small and are approximated to zero for the subsequent analyses. The input impedance seen from the input of the LCC network Z_{in} at the operating angular frequency ω can be derived from (4).

$$Z_{in} = \frac{L_S^2 \omega^2}{\frac{L_{TR}^2 \omega^2}{R_L} + j(K_{SSF} - 1)L_S \omega} \quad (4)$$

Fig. 5. V_{LS} and V_{CP} phasors of the LCC front-end signals.

where R_L is the load resistance. From (4) it can be seen that Z_{in} is capacitive for $K_{SSF} > 1$, and it is inductive for $K_{SSF} < 1$. It is also clear that Z_{in} is resistive for $K_{SSF} = 1$, and it makes i_{in} have the same phase angle with the fundamental of v_{in} . However, as v_{in} is a rectangular waveform, some capacitive Z_{in} is needed to maintain the zero current switching. The optimized capacitive behavior for zero current switching when the system is fed by a rectangular waveform is $K_{SSF} = \pi^2/8$ [27]. The injected apparent power to the LCC compensator at the fundamental angular frequency ω can be obtained from (5).

$$S_{in} = \frac{|V_{in}|^2}{Z_{in}} = \underbrace{\left(\frac{L_{TR}^2 \omega^2}{R_L} \right) \frac{1}{L_S^2 \omega^2} |V_{in}|^2}_{\text{Active Power}} - j \underbrace{\frac{K_{SSF} - 1}{L_S \omega} |V_{in}|^2}_{\text{Reactive Power}} \quad (5)$$

which includes active power and reactive power components. On the other hand, as the primary (front-end) loop of the LCC compensator is fully compensated, it acts as a current source for the transmitter current I_T , which is equal to $-jC_P \omega V_{in}$. Therefore, receiver current $I_R = [L_{TR}/(R_L L_S)] V_{in}$, and the transferred power (output power) (P_{out}) can be derived from the following equation.

$$P_{out} = R_L |I_R|^2 = \underbrace{\left(\frac{L_{TR}^2 \omega^2}{R_L} \right) \frac{1}{L_S^2 \omega^2} |V_{in}|^2}_{\xi} = \frac{\xi}{L_S^2 \omega^2} |V_{in}|^2 \quad (6)$$

It is interesting to observe from (5) and (6) that under the assumption of lossless components, the active power component in S_{in} is equal to the output active power consumed by the load, and the reactive power component of S_{in} does not depend on the transmitter-receiver mutual inductance and receiver parameters. However, the active power is also dependent on the receiver load R_L and the transmitter-receiver mutual inductance L_{TR} through $L_{TR}^2 \omega^2 / R_L$. Therefore, the term $L_{TR}^2 \omega^2 / R_L$, which makes a direct relationship between the input and output powers and depends on the receiver parameters, is defined as the demand factor ξ . In LCC networks, this factor is equal to the reflected impedance Z'_L ; however, it is different for other types of compensators. ξ can be estimated by directly measuring the active power from the LCC front-end terminals. In fact, with the assumption of loss-less ideal components, this is the expected phenomenon as input active power is completely consumed by the receiver load. Nevertheless, this principle can still be exploited even in a practical scenario (when the ESRs are present) to estimate and control the output

Fig. 6. v_{in} , v_{CP} , and v_{LS} LCC front-end wave-forms of voltage, and variation of v_{CP} PWM-synchronized samples with the changes in ξ .

power as long as the quality factor of the transmitter circuitry is kept high.

Therefore, in order to estimate the output power requirement, one does not need to estimate both mutual inductance and the load impedance. Instead, the transferred power can be directly calculated by estimating ξ . For this purpose, ξ can be identified by measuring the fundamental voltage component of the LCC front-end elements of L_S or C_P , V_{LS} or V_{CP} , which are given in (7) and (8), respectively. According to (7) and (8), the quadrature component of these voltages is proportional to the demand factor, which makes V_{LS} and V_{CP} excellent measurement candidates for the estimation of the demand factor. In case of zero phase switching, $K_{SSF} = 1$, V_{LS} is completely proportional to the demand factor. Nevertheless, zero phase switching may not be preferable due to the switching concerns [27]. In that case, the quadrature component of the measured V_{LS} or V_{CP} with reference to V_{in} needs to be obtained, as illustrated in Fig. 5.

$$V_{LS} = (1 - K_{SSF})V_{in} + j \left(\frac{\xi}{L_S \omega} \right) V_{in} \quad (7)$$

$$V_{CP} = V_{in} - V_{LS} = (K_{SSF})V_{in} - j \left(\frac{\xi}{L_S \omega} \right) V_{in} \quad (8)$$

Fig. 5 shows that the quadrature components of V_{LS} and V_{CP} have the same amplitude, but they are 180° out of phase. However, this assumption is only valid when there are no other harmonics except the fundamental component. Therefore the following subsection, analyses a more realistic implementation by considering the effects of higher-order harmonics.

C. Harmonic Analysis

Practically, the WPT system is driven by a rectangular waveform of voltage as shown in Fig. 6. Using the Fourier analysis, the simplified equivalent circuit for the n^{th} harmonic is shown in Fig. 7, and neglecting the effect of LCC back-end branch for harmonics, the voltage across L_S and C_P can be determined as given in (9) and (10) [27].

$$\bar{v}_{LS} = \frac{4V_{DC}}{\pi} \sum_{n=3,5,\dots}^{\infty} \frac{\hat{V}_{LS,n}}{n^2 - 1} \sin(n\omega t) \quad (9)$$

Fig. 7. Simplified circuit of the transmitter circuitry for harmonic contents.

$$\bar{v}_{CP} = -\frac{4V_{DC}}{\pi} \sum_{n=3,5,\dots}^{\infty} \frac{\hat{V}_{CP,n}}{n(n^2 - 1)} \sin(n\omega t) \quad (10)$$

where n is the harmonic order of the rectangular waveform of the input voltage V_{in} shown in Fig. 6, and $\bar{v}_{LS}$ and $\bar{v}_{CP}$ represent the harmonic content of the corresponding voltages. Comparing (9) and (10), it is apparent that the influence of harmonics on $\hat{V}_{LS,n}$ is n^2 times that of $\hat{V}_{CP,n}$, which makes V_{CP} less vulnerable to the harmonics. Therefore, V_{CP} becomes a more viable candidate for direct estimation of ξ . This is especially important when the direct measurements are to be done to estimate the transferred power. To explain the PWMS2, fundamental components of the input voltage and voltage $v_{CP}(t)$ can be written as in (11) and (12).

$$v_{in} = |V_{in}| \sin(\omega t) \quad (11)$$

$$v_{CP} = \overbrace{|V_{CP,d}| \cos(\psi) \sin(\omega t)}^{V_{CP,d}} + \overbrace{|V_{CP,q}| \sin(\psi) \cos(\omega t)}^{V_{CP,q}} \quad (12)$$

where ψ is the phase angle between V_{CP} and V_{in} and $k \in \mathbb{Z}$ is the sampling index. Therefore, the quadrature component of the LCC parallel capacitor $V_{CP,q}$, can be achieved as follows.

$$|V_{CP,q}(k)| = |v_{CP}(\omega t)|; \quad \omega t = 2k\pi \quad (13)$$

Equation (13) explains how by taking samples from v_{CP} at the moments of input voltage v_{in} zero crossing, the quadrature component of the front-end (V_{CP}) is achievable. $V_{CP,q}$ samples for different demand factors are shown as the circled markers at the input voltage zero crossing sampling instant shown in Fig.6. Therefore, from equations (8) and (13), the transferred power P_{out} for negligible R_S and R_T can be estimated by only taking PWM-Synchronized samples from v_{CP} and calculating the input power (P_{in}) as elaborated in (14).

$$P_{out}(k) \approx P_{in}(k) = \frac{2}{\pi} \frac{V_{DC}}{L_S \omega} |V_{CP,q}(k)| \quad (14)$$

where V_{DC} is the DC-link voltage of the resonant converter shown in Fig. 2.

III. INFLUENCE OF THE DETUNED REACTANCES

In this section, the effect of detuned parameters, which can happen in a practical situation is analyzed. To this end, it is assumed that the front-end loop of the LCC network, back-end loop of the LCC network, and series compensated

Fig. 8. Equivalent circuit of the proposed WPT system with detuned reactances.

receiver are not properly tuned, and they are detuned by ΔX_S , ΔX_T , and ΔX_R respectively. ESRs of the network are also considered negligible to simplify the calculations. Therefore, having the detuned reactances, the equivalent circuit diagram of the system is obtained as shown in Fig. 8.

In the first step, it is assumed that only receiver is detuned by ΔX_R . This assumption changes equation (5) to (15).

$$S_{in} = \underbrace{\frac{\xi_R}{L_S^2 \omega^2} |V_{in}|^2}_{\text{Active Power}} + j \underbrace{\left(-\frac{\xi_I}{L_S^2 \omega^2} + \frac{K_{SSF} - 1}{L_S \omega} \right) |V_{in}|^2}_{\text{Reactive Power}} \quad (15)$$

where $\xi = \xi_R - j\xi_I$ becomes a complex number due to the presence of ΔX_R . In (15), $\xi_R = L_{TR}^2 \omega^2 \left(\frac{R_L}{R_L^2 + \Delta X_R^2} \right)$ and $\xi_I = L_{TR}^2 \omega^2 \left(\frac{\Delta X_R}{R_L^2 + \Delta X_R^2} \right)$. Therefore, V_{CP} is changed to the following equation.

$$V_{CP} = \underbrace{\left(K_{SSF} - \frac{\xi_I}{L_S \omega} \right) V_{in}}_{V_{CP,d}} + j \underbrace{\left(\frac{-\xi_R}{L_S \omega} \right) V_{in}}_{V_{CP,q}} \quad (16)$$

From (16), it is clear that for the WPT system with a detuned receiver, one still can estimate the transferred active power by performing PWMS2 from the quadrature term of the LCC parallel capacitor voltage ($V_{CP,q}$). However, some reactive power is also demanded by the receiver. Unlike the transmitter fixed reactive power $j \frac{K_{SSF}-1}{L_S \omega}$, the reactive power corresponding to the detuned receiver is variable, and if necessary, by performing PWMS2 from the direct term of V_{CP} ($V_{CP,d}$), it can be estimated.

In the next step, the influence of the LCC back-end detuned reactance (ΔX_T) is taken into account. Following the same analogy for S_{in} , the detuned reactance of ΔX_T adds another constant reactive power component to the input apparent power S_{in} as obtained in (17).

$$S_{in} = \underbrace{\frac{\xi_R}{L_S^2 \omega^2} |V_{in}|^2}_{\text{Active Power}} + j \underbrace{\left(-\frac{\xi_I}{L_S^2 \omega^2} + \frac{K_{SSF} - 1}{L_S \omega} + \frac{\Delta X_T}{L_S^2 \omega^2} \right) |V_{in}|^2}_{\text{Reactive Power}} \quad (17)$$

It is worthy to mention that the reactive power caused by ΔX_T occurs due to the lack of tuning in the transmitter-side, and the receiver is not responsible for it. Therefore, if the utility charges the for the reactive power, ΔX_T needs to be properly compensated or the resulting reactive power should be removed from the price calculations. Adding the detuned reactance of ΔX_T to the back-end LCC compensator causes V_{CP} to have another constant direct term of $\frac{\Delta X_T}{L_S \omega} V_{in}$ as follows.

$$V_{CP} = \underbrace{\left(K_{SSF} - \frac{\xi_I}{L_S \omega} + \frac{\Delta X_T}{L_S \omega} \right) V_{in}}_{V_{CP,d}} + j \underbrace{\left(\frac{-\xi_R}{L_S \omega} \right) V_{in}}_{V_{CP,q}} \quad (18)$$

Therefore, still PWMS2 can be effectively used to estimate the active and reactive power components when ΔX_R and ΔX_T are present in the system. However, this is not the case when the detuned reactance of the front-end LCC network ΔX_S comes into effect. This reactance prevents the LCC T-bridge to be directly fed by the constant voltage of V_{in} . Therefore, it drastically influences the LCC network and causes it not to operate according to its ideal operational goal of the current source. Hence, the input apparent power S_{in} and the LCC parallel capacitor voltage V_{in} are changed to (19) and (20) respectively.

$$S_{in} = \underbrace{|K_S|^2 \frac{\xi_R}{L_S^2 \omega^2} |V_{in}|^2}_{\text{Active Power}} + j \underbrace{|K_S|^2 \left(-\frac{\xi_I}{L_S^2 \omega^2} + \frac{K_{SSF} - 1}{L_S \omega} + \frac{\Delta X_T}{L_S^2 \omega^2} \right) |V_{in}|^2}_{\text{Reactive Power}} \quad (19)$$

$$V_{CP} = K_S \left(K_{SSF} - \frac{\xi_I}{L_S \omega} + \frac{\Delta X_T}{L_S \omega} \right) V_{in} + j K_S \left(\frac{-\xi_R}{L_S \omega} \right) V_{in} \quad (20)$$

where K_S can be obtained from (21).

$$K_S = \frac{L_S^2 \omega^2}{j \left(-(K_{SSF} - 1) L_S \omega - \Delta X_T + j \frac{L_{TR}^2 \omega^2}{R_L + j \Delta X_R} \right) \Delta X_S + L_S^2 \omega^2} \quad (21)$$

It is clear from (19) and (20), the lack of tuning in the front-end loop of the LCC network (i.e., $\Delta X_S \neq 0$) causes the other detuned parameters (ΔX_T and ΔX_R) to influence the power estimation in a non-linear way. Therefore, amongst ΔX_S , ΔX_T , and ΔX_R , proper tuning of ΔX_S is crucially important to achieve an acceptable result from the proposed technique of transferred power estimation.

IV. INFLUENCE OF THE STRAY SERIES RESISTANCES

After analyzing the influence of the detuned reactances, in this section, the effect of the dominant ESRs in a properly tuned system is studied. In particular, coil resistances R_S and R_T need to be carefully studied to justify the efficacy of the proposed approach (ESR of capacitors that are used in WPT applications, such as Polypropylene capacitors, are negligibly small). Although the receiver coil ESR can also be relatively high, its effect is integrated into the receiver load as R_L . Therefore, the influence of R_S and R_T on the estimated power can be explained by (22).

$$\xi_{est} = \frac{L_S^2 \omega^2 (\xi^* + R_T)}{R_S (\xi^* + R_T) + L_S^2 \omega^2} \quad (22)$$

where ξ^* is the demand factor when ESRs are not considered in the system, and ξ_{est} is the demand factor when R_S and R_T influence the operation of the system. Similar to what is explained for the detuned front-end LCC loop (i.e. ΔX_S), the presence of R_S can also result in a nonlinear relationship between ξ and ξ^* , and it has to keep as small as possible. Having defined ξ^* and ξ_{est} , the error of estimation can be described as follows.

$$e_{\text{est}} = \frac{\xi_{\text{est}} - \xi^*}{\xi_{\text{est}}} = 1 - \frac{R_S}{L_S^2 \omega^2} \xi^* - \frac{1}{\xi^* + R_T} \xi^* \quad (23)$$

Considering R_S and R_T to represent the power loss in the LCC network and transmitter, the efficiency of the transmitter system can be obtained as follows:

$$\eta = \frac{\xi^*}{\xi^* + \frac{(\xi^* + R_T)^2}{L_S^2 \omega^2} R_S + R_T} \quad (24)$$

Defining $Q_{ST} = L_S \omega / \sqrt{R_S R_T}$ as the quality factor of the LCC network and transmitter, e_{est} and η can be written as

$$e_{\text{est}} = 1 - \frac{\left(\frac{\xi^*}{R_T}\right)}{Q_{ST}^2} - \frac{\left(\frac{\xi^*}{R_T}\right)}{\left(\frac{\xi^*}{R_T}\right) + 1} \quad (25)$$

$$\eta = \frac{\left(\frac{\xi^*}{R_T}\right)}{\left(\frac{\left(\frac{\xi^*}{R_T}\right) + 1}{Q_{ST}^2}\right) + \left(\frac{\xi^*}{R_T}\right) + 1} \quad (26)$$

Therefore, to achieve the maximum efficiency,

$$\left(\frac{\xi^* (\eta_{\text{max}})}{R_T}\right) = \sqrt{Q_{ST}^2 + 1} \quad (27)$$

Figs. 9(a) and (b) illustrate the effect of ξ^* and Q_{ST} variations on e_{est} and η respectively. In these figures, red solid line shows the places at which the efficiency is the highest (η_{max}).

As it can be seen in (22), the presence of R_S makes a nonlinear relationship between ξ_{est} and ξ^* , and R_S has an undesirable effect on the estimation of ξ^* . As shown in Fig. 9, to avoid the occurrence of such a phenomenon, the quality factor of the LCC front-end loop (Q_S) should be as high as possible. In the next section, along with the experimental verification of the proposed concept, the influence of Q_{ST} and load variation (R_L) on ξ_{est} will be numerically investigated and compared with the experimental results.

V. EXPERIMENTAL RESULTS

In this section, the experimental results obtained from the proposed system, shown in Fig. 2, are analyzed and compared with the expected numerical results. Fig. 10 shows the laboratory prototype of the WPT system, which has been built based on the specifications given in Table I. The WPT system is driven by a resonant converter with the DC-link voltage of $V_{\text{DC}} = 180$ V and the frequency of 80 kHz. The experiments are carried out to study the Steady State Operation, Dynamic Estimation of the Transferred Power, and The Influence of Sampling Jitter Noise in the proposed system.

Fig. 9. Variation of (a) e_{est} and (b) η_{max} with respect to Q_{ST} and ξ^* .

In these studies, to achieve the actual input power (P_{in}) and output power P_{out} , the input voltage (v_{in}) and current (i_{in}) and the load voltage (v_L) and current (i_R) are measured, respectively. Then the voltages and currents are captured for one cycle of the operating angular frequency ($T = 1/(2\pi\omega)$), and the average of their multiplication in one cycle (i.e. $T^{-1} \int_0^T v i dt$) is used to measure the average of the active power in one cycle of operation. To obtain the actual efficiency (η), the actual output power is divided by the actual input power.

To estimate the demand factor, the mutual inductance of the WPT system (L_{TR}) is experimentally measured as shown in Fig 11 and together with the known values of R_L and angular frequency (ω) are used in the equation of the estimated demand factor ($\xi = (L_{TR}\omega)^2/R_L$). The output power is estimated with the use of PWMS2 technique explained in Section II. According to (25), the quadrature component of V_{CP} phasor ($V_{\text{CP},q}$) has a direct relationship to the output power. To find the amplitude of $V_{\text{CP},q}$, PWM-synchronized samples are taken when the fundamental component of the input voltage crosses zero. For this purpose, the sampling interrupts can be easily synchronized with the reference of the PWM signals generated for the resonant converter.

Fig. 10. The experimental setup; (a) the whole setup, (b) WPT pads, and (c) transmitter driving system.

TABLE I
SIMULATION AND EXPERIMENTAL SPECIFICATIONS
(DIMENSIONS ARE LABELED IN FIG. 10(B)).

Parameter	Value
Transmitter LCC series inductance (L_S)	69 μ H
Transmitter LCC parallel capacitance (C_P)	68 nF
Transmitter LCC series capacitance (C_T)	68 nF
Transmitter self-inductance (L_{TT})	132 μ H
Receiver self-inductance (L_{RR})	73 μ H
Receiver series capacitance (C_R)	74 nF
Ferrite blocks type	N97
Ferrite blocks dimensions ($L_F \times H_F \times W_F$) (mm)	102 $\times$ 7 $\times$ 38
Litz wire	435/AWG 40
Number of turns in WPT pads [N_{Tx}, N_{Rx}]	[16, 12]
WPT pads dimensions ($L_P \times H_P \times W_P$) (mm)	200 $\times$ 20 $\times$ 200

Fig. 11. Mutual inductance profiles of the WPT system for different vertical components of the air gap (G) and the symmetric range under study for the minimum to maximum level of the transferred power.

Fig. 12. v_{in} , v_{CP} , and v_{CP} PWM-synchronized samples at five different receiver positions for $R_L = 0.75 \Omega$ and $G = 130$ mm.

A. Steady State Operation

To analyze the steady state operation of the WPT system, the receiver is located at different positions and the actual input power, actual output power, actual efficiency, estimated demand factor, estimated output power, and estimation error are measured when the system reaches its steady state behavior at each position. The investigations are carried out for different values of L_{TR} , R_L , R_S , and R_T .

To study the effect of L_{TR} on the transferred power estimation, the receiver pad moves horizontally along the x axis. The horizontal displacement along x axis is carried out for three different vertical components of the transmitter-receiver air gap (G), which are $G = 180$ mm, $G = 130$ mm, and $G = 80$ mm. For clarity, $G = 130$ mm is considered as the benchmark of the investigation for L_{TR} variations. As the mutual inductance profile of the WPT system is symmetric along $x = 0$ cm, for the sake of simplicity, the measurements are carried out for sixteen evenly distributed positions of the receiver (starting from $x = 0$ cm (aligned position) to $x = 15$ cm where the null of coupling (the null of power) occurs for $G = 130$ mm) as shown in Fig. 11. This range also covers the null of coupling for $G = 80$ mm; therefore, the whole possible range for the transferred power for $G = 130$ mm and $G = 80$ mm is covered within $x = 0$ cm to $x = 15$ cm.

Fig. 13. Experimental results for the estimated (ξ_{est}) and actual (ξ^*) demand factor profiles ($x = 0$ cm to $x = 15$ cm) for (a) different load values (R_L) when $G = 130$ mm and (b) different vertical components of the air gap (G) when $R_L = 3 \Omega$.

To study the effect of different loads on the estimation of the transferred power, three different loads ($R_L = 0.75 \Omega$, $R_L = 1.5 \Omega$, and $R_L = 3 \Omega$) are connected to the receiver terminals, among which $R_L = 3 \Omega$ is considered as the benchmark load resistance for the investigations. Moreover, the influence of stray resistors, R_S and R_T , are studied by adding different artificial resistors to the front-end loop and back-end loop of the LCC compensator.

Having the measured transmitter-receiver mutual inductance (L_{TR}), load at the receiver (R_L), LCC series inductance (L_S) and with the use of (6), the actual demand factor (ξ^*) is calculated, and the results are shown in Fig. 13. In this figure, the experimental result of the estimated demand factor (ξ_{est}) obtained from PWMS2 is also shown for the comparison. Fig. 13(a) shows the demand factor profiles for different loads ($R_L = 0.75 \Omega$, $R_L = 1.5 \Omega$, and $R_L = 3 \Omega$) when the vertical component of the air gap is $G = 130$ mm, and Fig. 13(b) shows how different vertical components of the air gap ($G = 80$ mm, $G = 130$ mm, and $G = 180$ mm) influence the demand factor profiles when the load resistance is $R_L = 3 \Omega$. As expected from (22) and (23), R_S and R_T can impact the accuracy of the estimation and cause ξ_{est} to deviate from ξ^* . This phenomenon is also shown in Fig. 9(a) which represents how R_T and Q_{ST} influence the estimation error of the demand factor (ξ_{est}) compared to its actual value (ξ^*).

Fig. 14. Experimental results for the input power (P_{in}), output power (P_{out}), and estimated power (P_{est}) profiles for (a) different load values (R_L) when $G = 130$ mm and (b) different vertical components of the air gap (G) when $R_L = 3 \Omega$.

In a similar way to the measurement of the demand factor, the input power (P_{in}), output power (P_{out}), and the estimated power (P_{est}) are experimentally measured at different receiver positions, as shown in Fig. 14. Fig. 14(a) shows the power profiles for different loads ($R_L = 0.75 \Omega$, $R_L = 1.5 \Omega$, and $R_L = 3 \Omega$) and when $G = 130$ mm, and Fig. 14(b) shows the power profiles for different vertical components of the air gap ($G = 80$ mm, $G = 130$ mm, and $G = 180$ mm) and when $R_L = 3 \Omega$.

The efficiency profiles of the system for different load values and different vertical components of the air gap are shown in Figs. 15(a) and (b), respectively.

Similar to other estimation techniques, estimation error is an unavoidable part of the proposed technique. This error can be clearly observed by comparing the estimated and actual values shown in Figs. 13 and 14, particularly, where the mutual inductance between transmitter and receiver is very small.

As P_{out} is obtained by measuring the voltage across and the current through the resistive load, power loss caused by the receiver coil ESR (R_R) is not considered in the obtained output power. Regardless of the power estimation technique, R_R makes an error of estimation in the transferred power, and when R_L get closer to R_R the error of estimation increases. This fact is partly shown in Fig. 16 for the experimental results obtained for the estimation error (e_{est}) versus the output power

Fig. 15. Experimental results for the efficiency (η) profiles for (a) different load values (R_L) when $G = 130$ mm and (b) different vertical components of the air gap (G) when $R_L = 3 \Omega$.

(P_{out}) when the receiver moves from $x = 0$ cm to $x = 15$ cm and for different load values (R_L). As shown in Fig. 16(a), when R_L (at $G = 130$ mm) decreases (and becomes closer to R_R) the estimation error increases. However, Fig. 16(b) implies that for different vertical components of the air gap and when $R_L = 3 \Omega$, the obtained estimation errors follow a same pattern.

Therefore, according to what is already mentioned in Section IV and partly explained here, to minimize the estimation error, the quality of the transmitter (Q_{ST}) and receiver ($Q_R = L_{RR}\omega/R_R$) systems should be as high as possible. In some applications, such as the dynamic chargers used for EVs, each EV can have a different Q_R . It seems reasonable that the lack of Q_R should not be the WPT service providers' concern, and the users should be charged according to the amount of active power they receive from the transmitter (service provider) and regardless the amount of power which is dissipated as loss in their system (e.g. EVs). Therefore, in the point of view of the quality, the main concern of the WPT service providers should be the increase of the quality factor in the transmitter system, which is Q_{ST} in the proposed technique, as much as possible. In practice, low demand of power from the receiver is not considered as a working condition, and the transmitter is turned OFF for such situation. Therefore, it is expected to achieve a desirable level of accuracy in the proposed technique of power estimation when a reasonable amount of power is

Fig. 16. Experimental results for the variation of the estimation error (e_{est}) versus output power (P_{out}) for (a) different load values (R_L) when $G = 130$ mm and (b) different vertical components of the air gap (G) when $R_L = 3 \Omega$.

demanded by the receiver, during which the transmitter is ON and actively works.

The estimation error can also be used to define an effective range for the estimation of the transferred power. As an example, according to Figs. 14 and 16, to achieve an estimation error below 20% for the proposed system, the effective range of estimation for $G = 130$ cm and $G = 60$ cm should be within $[0 - 6]$ cm and $[0 - 8]$ cm, respectively. This is while for $G = 180$ cm, the estimation error is always higher than 20%, and the effective range of estimation should be defined for higher estimation errors. In fact, the higher estimation error corresponds to lower output power, which could be due to low coupling or low power demand. Such working condition is anyway not efficient, and the power transmitter is usually turned-off in practical application.

To combine the effect of transmitter-receiver mutual inductance (L_{TR}) and load variations (R_L) for different displacements (x) and loads (R_L), the variation of the experimentally estimation error (e_{est}) versus efficiency (η) is depicted and compared with the numerical results in Fig. 17. Figs. 17(a) and (b) show that the variation of the receiver load (R_L) when $G = 130$ mm and the variation of the vertical component of the air gap when ($R_L = 3 \Omega$) have a negligible effect on the distribution pattern of the operating points on $e_{est} - \eta$ plan.

The obtained experimental results are more expanded around the $Q_{ST} = 34$ contour which is dependent to the parasitic components in the transmitter pad, such as R_S and R_T .

Fig. 17(c) further verifies the dependency of the experimental point distribution to Q_{ST} . This figure also shows the nonlinear influence of the primary loop of the LCC ESR (R_S) when the additional R'_S is added to the network. Similarly, some additional ESRs are added to the transmitter pad (R'_T), and it is expected, the influence of R_T is almost linear.

From Fig. 13, it can be seen that the estimated demand factor (ξ_{est}) follows the actual demand factor (ξ^*) in a reasonable way. Moreover, Fig. 17 shows that the experimental results are completely in accordance with the numerical results, and it follows the $e_{est} - \eta$ pattern for $Q_{ST} = 34$. Fig. 17 also explains that the higher the efficiency, the lower the **estimation error**, which means this method can accurately observe the flow of power when the effect of parasitic components are negligible and transmitter and receiver are effectively coupled. On the other hand, the lack of accuracy happens when the efficiency is low, which occurs when the parasitic components are not negligible and the transmitter-receiver coupling is low. This, however, cannot be inferred as a major drawback, as the operation of WPT systems at low efficiencies is not desirable, and transmitter pads are preferred to be turned OFF at those conditions where the estimated demand factor is low (high **estimation errors**). Furthermore, by putting the transmitter in the idle mode of operation (transmission of a low amount of power only to detect the presence of the receiver), the presence of the receiver can still be detected at the low accuracy zone of this approach. This can be done owing to the meaningful variation of the estimated demand factor when the receiver approaches the transmitter, even when the coupling is low (low efficiencies). Therefore, although for low couplings the **estimation error** is high, this method can still be used to detect the effective range of the receiver displacement to activate or deactivate the transmitters.

B. Dynamic Estimation of the Transferred Power

In this subsection, the operation of the proposed PWMS2 technique when the receiver load (R_L) dynamically changes is investigated. To make the highest possible change in the demand factor (and transferred power), the transmitter and receiver pads are kept at aligned position for $G = 130$ mm, and the receiver load (R_L) is suddenly changed from 3Ω to 1.5Ω . During this change, v_{in} , i_{in} , and v_{CP} are captured to find the input power (P_{in}) and the estimated transferred power (P_{est}). To obtain P_{est} , PWMS2 is carried out at a sampling frequency of twice the operating frequency, and v_{CP} is sampled with the sampling rate of $f_{PWMS2} = 160$ kHz.

This section also compares the conventional method of power estimation with the proposed PWMS2 technique. In conventional techniques of active power estimation, the input power of the system is obtained by taking the average value of the instantaneous input power $p_{in}(t) = v_{in}(t)i_{in}(t)$ with the use of low-pass filters (LFPs), as shown in Fig. 18. In order to effectively capture the average value for $p_{in}(t)$, v_{in} and i_{in} should be sampled according to the Nyquist frequency

Fig. 17. Experimental and numerical results for the variation of the estimation error (e_{est}) versus efficiency (η) for (a) different load values (R_L) when $G = 130$ mm, (b) different vertical components of the air gap (G) when $R_L = 3 \Omega$, and (c) different additional ESRs R'_S and R'_T when $G = 130$ mm and $R_L = 3 \Omega$.

Fig. 18. Conventional approach to estimate the input power (P_{in}) of the WPT system.

criterion (i.e. sampling frequency should be at least twice the operating frequency). Therefore, the rate of sampling for v_{in}

Fig. 19. Experimental dynamic results for P_{in} and P_{est} when the receiver load changes from 3Ω to 1.5Ω at $t = 0.26$ ms. Conventional low-pass filter estimations with time constants of $T_s = 5$ us and $T_s = 100$ us are shown as LPF_1 and LPF_2 , respectively. The proposed estimation technique is also shown as $PWMS2$.

and i_{in} is considered five times more than that of the $PWMS2$ (i.e. $f_{LPF} = 5f_{PWMS2} = 800$ kHz).

The LPF used to achieve P_{in} is a first-order discrete transfer function. The equivalent continuous transfer function of the LPF is $H_{LPF}(s) = (1 + T_s s)^{-1}$, where T_s is the time constant of the filter. Increase in T_s results in increased attenuation of undesirable harmonics. However, the higher the T_s the higher the response time of the filter would be. Fig. 19 shows the obtained experimental results for P_{in} and P_{est} when the receiver load changes from 3Ω to 1.5Ω at $t = 0.26$ ms. In this figure, LPF_1 stands for the obtained P_{in} when an LPF with the time constant of $T_s = 5$ us is used. Similarly, LPF_2 represents P_{in} when the time constant of the LPF is $T_s = 100$ us. Moreover, $PWMS2$ stands for estimated transferred power (P_{est}) obtained from the PWM-Synchronized sampling technique. In the next subsection, the sensitivity of the proposed $PWMS2$ technique against sampling jitter noise is studied.

From 19, it can be seen that the dynamic changes in P_{in} and P_{est} gradually approach to their steady state values as shown in Figs. 14. Moreover, Fig. 19 shows that $PWMS2$ can effectively estimate the transferred power with a minimum rate of sampling compared to the conventional approach. The advantages of this technique compared to the conventional approach are the fast response, desirable harmonic rejection, low sampling rate, and possibility of estimation by measuring only one variable (v_{CP}).

C. The Influence of Sampling Jitter Noise

Almost all observation techniques suffer from background and sampling jitter noises. By nature, these noises are random, and their frequency spectrum is extensively wide. By definition, the noise caused by sampling jitter mainly occurs due to the random and small deviation of the sampling instances within the system. As previously mentioned, in the $PWMS2$ technique, the moment of sampling is significantly important; therefore, sampling jitter noise can be considered as one of the most influential factors on the error of estimation. In

Fig. 20. Sensitivity contours of $PWMS2$ to $\delta(t)$ and ψ .

Fig. 21. Estimation error e_{est} versus estimated demand factor ξ_{est} when the sampling instant deviated for two, one, and zero cycles of the fundamental frequency.

this subsection, the influence of sampling jitter noise on the $PWMS2$ technique is investigated. Unlike the estimation error caused by the detuned parameters and parasitic components, sampling jitter noise has a random nature and can be changed based on the quality of the microcontroller board and electromagnetic interference in the environment. Here, sampling jitter is defined as a small time deviation of $\delta(t)$ from the sampling instant. According to (23), $\delta(t)$ can be included into $V_{CP,q}$ as given in (28).

$$|V_{CP,q}^*| = |V_{CP}| \sin(\psi + \delta(t)) \quad (28)$$

where $V_{CP,q}^*$ is the quadrature term of v_{CP} when $\delta(t)$ affects the system. The sensitivity of $|V_{CP,q}^*|$ to $\delta(t)$ can be calculated as given in (29).

$$\frac{\partial |V_{CP,q}^*|}{\partial \delta(t)} = |V_{CP}| \cos(\psi + \delta(t)) \quad (29)$$

Therefore, the sensitivity of the system to the sampling jitter increases when ψ decreases. In other words, when the demand factor is small or the system is not properly loaded, the estimation error caused by sampling jitter is at its worst value. Fig.20 shows the sensitivity of $PWMS2$ to $\delta(t)$ and ψ .

To show how decrease in ψ (i.e. small demand factor) exacerbates the influence of sampling jitter ($\delta(t)$), Fig. 21 shows the variation of the estimation error e_{est} versus estimated

VII. CONCLUSION

In this article, a new approach to observe the transferred power in WPT systems is presented. Taking the PWM-synchronized samples from the front-end variables of the LCC compensator makes this technique a simple and feasible way to be used in different applications. The mathematical equations linking the transferred power (which is received by the receiver) to the front-end variables are derived, and it is shown that the transferred power can be observed by taking the PWM-synchronized samples from the LCC parallel capacitor. Then, by considering the effect of dominant equivalent resistances in the system, the accuracy of this system is characterized. Finally, the close correlation between the numerical and experimental results validates the proposed technique.

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Fig. 22. Comprehensive depiction of the proposed technique of power estimation.

demand factor ξ_{est} for sampling instant leads of 2, 1, and 0 operating frequency cycles. This figure clearly shows that for low ξ_{est} , the system is more sensitive to the deviation of the sampling instant, and it results in higher estimation error.

Therefore, the use of the PWMS2 for the given WPT system is recommended when the LCC filter is designed in such a way that the nominal ψ (or demand factor) is high enough. This way not only the steady state errors decrease, but also the system becomes more resilient against sampling jitter noises.

VI. SUMMARY

In this section, the proposed technique of transferred power estimation is summarized in the flowchart shown in Fig. 22. As previously mentioned, in the proposed technique, the transferred power is estimated without any knowledge of the receiver parameters and through the estimation of the demand factor at the transmitter-side. The main objective of the process flowchart is to estimate the demand factor, named as "Objective A", and the final objective, named as "Objective B", is to estimate the transferred power. To achieve these two main objectives, as shown in Fig. 22, a three-step flowchart should be processed. In the first step, the quadrature term of the LCC parallel capacitor voltage ($V_{CP,q}$) should be sampled. This goal is achievable by performing PWMS2 as elaborated in Section II. This step is labelled as ① in the process flowchart shown in Fig. 22. In the next step, with the knowledge of the transmitter parameters and variables, such as L_S , operating angular frequency (ω), $V_{CP,q}$, and V_{in} , the demand factor can be estimated. This step is labelled as ② in the process flowchart shown in Fig. 22. Through this step "Objective A" is accomplished. The estimated demand factor is not necessarily accurate and can be influenced by the detuned and parasitic components and the sampling jitter noise explained in Sections III, IV, and V. Finally, in the third step of the process flow chart, labelled as ③ in Fig. 22, the transferred power is estimated to fulfill "Objective B".

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