

Voltage-fed Full Bridge Inverter Topology for Inductive Wireless Power Transfer Application

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Abstract

This paper presents wireless power transfer topology for consumer electronics and electric vehicle (EV) charging application. The main focus is given on high-frequency DCAC and AC-DC wireless power transfer stage where the inversion stage is Voltage-fed topology. The required resonance in transmitter side is (C) (LC) type and in the receiver-side series (LC) type. The proposed topology network reduces the inverter switch stress to half of the conventional topology. Capacitor in dc-link provides constant input voltage. The required hightransmitter coil (TC) current for air core coupled inductor circulates through parallel resonating capacitor which reduces inverter switch current stress. Resonant converter provides soft switching at device turn ON which reduces switching loss. Also, soft recovery at bridge rectifier eliminates diode reverse recovery loss. The converter is analysed and simulated using PSIM 9.03.

1 INTRODUCTION

Wireless power transfer Research has increased significantly since its practicality is proven by many research works. WPT technology finds its application in electric vehicle, electronic gadgets, chemical plants, lighting, and underwater vehicle applications. This technology has huge potential to expand especially in electric vehicle (EV) [1] [2] market since it reduces the energy storage/distance limitations of EVs and the charging process is very convenient to the users. Because of these advantages, WPT technology can shift the gasoline-dependent vehicles to EVs with faster rate, hence less dependent on petroleum and less environmental impact. Of the two nonradiation based WPT technologies, inductive power transfer (IPT) is suitable for both low and high-power applications, whereas capacitive power transfer (CPT) is suitable for lower power applications [3]. Since, in EV, the charging power level is in kilowatt level, IPT is preferred for EV charging applications. Fig. 1 shows a general power train of EV charging with IPT technology [4]. The crucial element in IPT system is coupled inductor where transmitter coil (TC) and receiver coil (RC) has an air gap of 150-300 mm. Effective power transfer is not possible by injecting normal power frequency voltage to TC since the coupling coefficient between TC and RC is very low. However, efficient power transfer is possible, if TC receives a voltage with its resonance frequency and RC resonance frequency is tuned to same frequency [5] [6]. Because of the presence of large air gap, the overall efficiency in this system is slightly lower compared with conductive charging [7] [8]. Several IPT topologies based on different types of transmitter and RC resonances are discussed in [2], [9][16]. Conventionally, the inversion stage is chosen as voltage source inverter (VSI) topology or matrix converter topology [17]. In VSI topology, the transmitter-side series resonance is achieved by placing the capacitor in series. Since the required reactive component of TC current flows through the VSI switches, TC resonating network is modified as LCL network [2], [9].

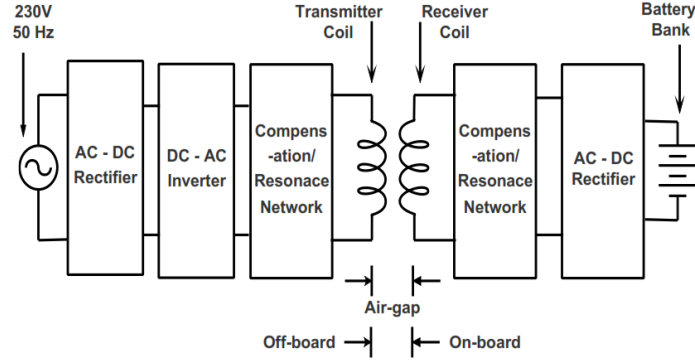


Fig.1. Powertrain of wireless Power Transfer

voltage-fed inverter topology is also a viable solution for WIPT and this is the main focus of this paper. It reduces the current stress through the devices which lead to reduced conduction losses of the switches. In addition, the capacitor is located on transmitter side, which is stationary. Therefore, for some applications, increase in weight or slight increase in volume for one stationary component might not be a major issue. In-build short-circuit protection is an added benefit for this application. Inverter operates at very high frequency; hence, the inductor size is small. The cost of inductor is generally higher. For such reasons, voltage-fed topology is also a possible solution. In voltage-fed topology, parallel resonance in the TC is achieved by placing capacitor across the coil. Since the reactive current of a voltage-fed resonant converter circulates inside the parallel resonant tank without owing through the switching network, the current rating of the switching devices is smaller and the conduction loss is reduced for a given power level. Also, TC current in this topology is almost pure sinusoidal since parallel capacitor provides much lower impedance than the TC. In spite of these advantages, the implementation of Voltage-fed topology in high-power IPT application did not get much attention [16]. The major limitation of this topology is that the voltage stress on inverter switches are high, especially when the load power is high and it becomes worse when coefficient of coil coupling is lower. In this paper, a new IPT topology with inversion stage as voltage-fed topology is presented. The required resonance network in transmitter side is (C) (CL) type and the receiver side resonance network is

series type. Along with all other advantages of voltage-fed converter in IPT application, lower voltage stress across the inverter switches is also achieved with the capacitor in series with the TC.

2 IPT USING VOLTAGE-FED TOPOLOGY: ANALYSIS AND DESIGN

A. IPT With Conventional Voltage-Fed Inverter

Where the inversion stage is VSI topology are discussed [18]. Since the input of the VSI is constant voltage, parallel resonance/compensation in TC coil is achieved by placing a capacitor C parallel to the TC. In the receiver side, the compensation is either parallel or series type. TC current I_t is almost pure sinusoidal even though the inverter output Voltage V_{inv} is a square wave. However, the major limitation of this circuit is the high-voltage stress on VSI switches specially when the required load power is high and the coupling between TC and RC is low. It is because the TC due to a bulk amount of reactive power when the required load power is high and coefficient of coupling is relatively low. Since this high amount of reactive power is produced by the parallel capacitor, the voltage of the capacitor increases and eventually appears across VSI switches. This limitation of the circuit is resolved by adding a capacitor in series with TC.

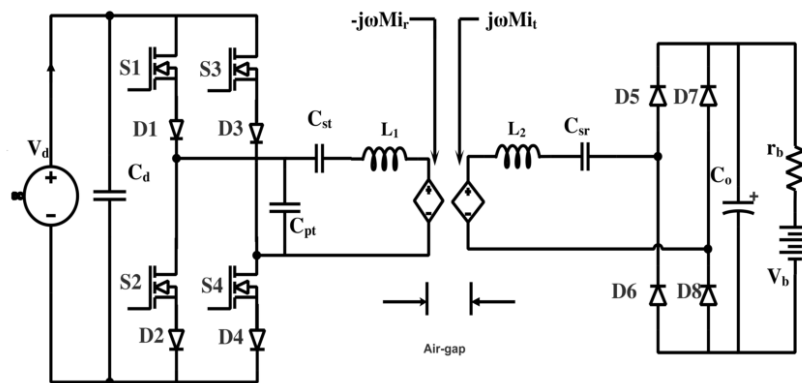


Fig.2. Circuit diagram of Voltage-fed wireless power transfer Topology

B. Proposed IPT Circuit With Lower Voltage Stress of VSI Switches

Fig. 2 shows the complete circuit diagram of proposed voltage-fed IPT converter. The input dc voltage V_d can be either a dc supply mains or the output of a power factor corrected (PFC) rectifier. To convert this dc to high-frequency ac, a voltage-fed inverter topology is used where the Capacitor C transforms stiff current I_d to stiff voltage V as shown in Fig. 2. Considering that switching frequency of VSI is sufficiently high and the Capacitor C_d is sufficiently large, the VSI injects a square wave voltage I_{inv} to next transmitter resonant network stage. Because of air core, the value of the mutual inductance M is much lower than the conventional iron core coupled inductor. Hence, to transfer a significant amount of active power through the air, the required TC current has to be high enough such that sufficient amount of voltage in the $RC(\omega MIt)$ is induced. In this topology, a capacitor C_{pt} is added in Parallel with the TC; hence, the effective impedance of the coil is reduced and parallel capacitor C_p needs to deliver lesser reactive power than the conventional circuit. To reduce the VSI switch current stress, the selection of the capacitors C_{pt} and C_s has to be such that the reactive component of the TC current circulates through C_{pt} and VSI switches supply only the active component of current. The purpose of the series diodes $[D_1, D_2, D_3, D_4]$ is to prevent short circuit of capacitor C through switches and body diodes of the switches. For example, when switches $S1$ and $S4$ are ON and if the voltage across C is negative, then there is a short-circuit path of C through $S4$ - $S2$ body diode. However, if the inverter output voltage and current are in same phase, then there is no need of these series diodes. But in practical, circuit parameters change and mutual inductances vary; hence, power factor deviates from unity. Thus, the series diodes are mandatory. Reverse voltage blocking insulated gate bipolar transistors (IGBTs) (RB-IGBT) are suitable option for this topology. However, considering the switching loss of the IGBTs at higher switching frequency and also considering the lack of availability of reverse blocking IGBTs for wide voltage range, MOSFET-diode combination is used at this stage.

A series capacitor C is connected to achieve the required resonance in RC . Thus, the large voltage drop across the RC leakage inductor L_{sr} is compensated. Next, diode bridge rectifies the RC

current and feeds the battery.

C. Proposed IPT Circuit With Lower Voltage Stress of VSI Switches

The parallel-parallel compensation topology is also a viable solution for WPT. However, parallel topology in the receiver side suffers from circulating current [19] and this leads to higher copper loss in the RC. Also, parallel compensation on the receiver side leads to inductive lter after the diode bridge rectifier. This will increase on-board charger weight as well as needs secondary snubbers to limit ringing and diode voltage stress. The switches S1 and S4, a diagonal pair, are modulated by identical gating signals. Similarly, other diagonal pairs of switches S2 and S3 are triggered by identical gating signals. The key message is that all four devices cannot be OFF at a time. Only one diagonal pair is turned OFF at a time to pass the current from source to the coils; otherwise, all four devices are

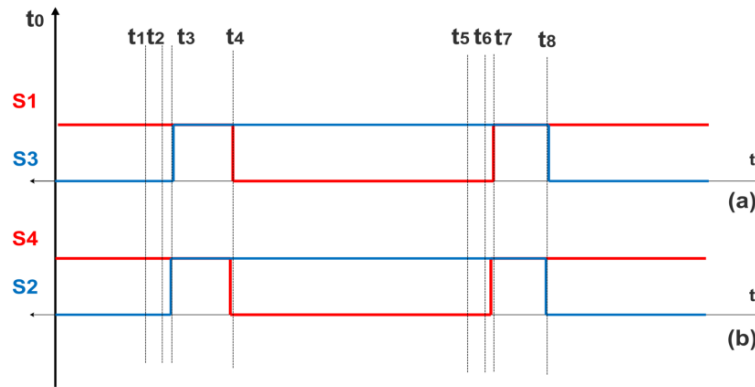


Fig.3. PWM pulses for Proposed WPT

ON during overlap allowing the inductor to store the energy. It is similar to boost converter principle. Referring to Fig. 2 when switching signal S1 is high, capacitor C_d always gets a continuous path irrespective of the status of S2. Similarly, when S1 is low, i.e., S1 is high, C_d always gets a continuous path. Calculation of circuit parameters and the voltage and current stress of the components are derived here.

D. Design of Compensation Capacitor

Applying Kirchhoffs current law (KCL) at inverter output, the inverter current is given as

$$I_{inv} = I_t + I_c \quad (1)$$

where I_t is TC current; I_c is capacitor C_p , current. Replacing the currents with branch voltage and impedance, (1) is modied as

$$I_{inv} = \left[\frac{V_{inv} + j\omega M I_r}{j\omega L_1 + \frac{1}{j\omega C_s}} \right] + j\omega C_p V_{inv} \quad (2)$$

$$= \left[\frac{\omega^2 M C_s I_r}{\omega^2 L_1 C_s - 1} \right] + j \left[\omega V_{inv} \left(C_p - \frac{C_s}{\omega^2 L_1 C_s - 1} \right) \right] \quad (3)$$

where L_1 is self-inductance of TC, ω is the angular frequency, M is mutual inductance, I_r is RC current, and TC-to-RC turns ratio is 1:1. From Fig. 3, it is clear that without parallel capacitor C_p the required high amount of TC current would pass through VSI switches. The magnitude of VSI output current is minimized by selecting C_{pt} and C_{st} such that second part of (2) is removed. On the receiver side, voltage injected at the input of the diode bridge is given as

$$V_{rD} = j\omega M I_t - I_r \left[j\omega L_2 + \frac{1}{j\omega C_r} \right] \quad (4)$$

where C_{sr} is receiver-side series compensation/resonance capacitor. The value of C_{sr} is chosen such that bridge input voltage V_{rD} is maximum which is done by removing the second part of (3). From (2) and (3), required capacitances are calculated as

$$\omega_t = \frac{1}{\sqrt{L_1 \left(\frac{C_p C_s}{C_p + C_s} \right)}}, \omega_r = \frac{1}{\sqrt{C_r L_2}} \quad (5)$$

Effective power transfer is achieved by equalizing VSI switching frequency $\omega_s (= 2\pi f_s)$, TC resonance frequency ω_t , and RC resonance frequency ω_r

E. Derivation of Switch Voltage and Current Rating

In order to simplify the calculation, coil resistance and switch losses are neglected which is reasonable since voltage drop across coil is mainly due to its inductance. It is clear that the reverse blocking voltage rating of the diodes D5, D6, D7, and D8 are the same as battery voltage V_b . The current waveform of the RC is sinusoidal since it passes through the series resonating circuit. Hence, the peak and RMS current rating of the diodes are given as

$$\widehat{I_D} = \frac{\pi}{4} I_0, I_D = \frac{\pi}{2\sqrt{2}} I_0 \quad (6)$$

where I_0 is average battery charging current. Transmitterside switch and diode current rating are the same as dc-link Capacitor C_d . Peak forward-blocking voltage of the switches or reverse voltage-blocking voltage of the series diodes are dependent on peak voltage of capacitor C_{pt} voltage. The active load, i.e., battery is replaced by its ac equivalent voltage V_{eq} . Since the internal resistance of rechargeable battery is in the order of milliohm or fraction of ohm, and the voltage drop across it is much less when compared with the battery voltage, it is not included in the derivation. Although V_{eq} wave-shape is square wave, only its fundamental component is involved in active power transfer, since current I_r is sinusoidal. It is clear from Fig. 3 that I_r and V_{eq} are in same phase, since the receiver-side diode bridge toggles the conduction of diode pairs just after the change of polarity of I_r . In order to derive the voltage and current stress on circuit components, consider that the current I_r or voltage V_{eq} is reference phasor. Applying power balance and using Fourier analysis, RMS value of equivalent load voltage and RC current are derived as

$$V_{eq} = \frac{2\sqrt{2}}{\pi} V_o, I_r = \frac{\pi}{2\sqrt{2}} I \quad (7)$$

Applying Kirchhoffs voltage law (KVL) in the receiverside loop, TC current is calculated as

$$I_t = -j \frac{2\sqrt{2}}{\pi} \frac{V_o}{\omega_s M} \quad (8)$$

Applying KVL and KCL in transmitter side and using (6) and (7), RMS value of voltage across TC and voltage across series capacitor

C_s is derived as

$$V_t = \frac{2\sqrt{2}}{\pi} \frac{L_1}{M} V_o - j \frac{\pi}{2\sqrt{2}} \omega_s M I_0 \quad (9)$$

$$V_{cs} = -\frac{2\sqrt{2}}{\pi} \frac{V_0}{\omega_s^2 C_s M} \quad (10)$$

Applying KVL, RMS value of inverter output voltage V_{inv} is calculated by adding (8) and (9) and applying KCL, inverter output current is calculated as

$$V_{inv} = \frac{2\sqrt{2}}{\pi} \frac{V_0}{\omega_s M} \left(\omega_s L_1 - \frac{1}{\omega_s C_s} \right) - j \frac{\pi}{2\sqrt{2}} \omega_s M I_o \quad (11)$$

$$I_{inv} = \frac{\pi}{2\sqrt{2}} \omega_s^2 M C_p I_o - j \frac{2\sqrt{2}}{\pi} \frac{V_0 C_p}{M} X \left(\frac{1}{\omega_s C_s} + \frac{1}{\omega_s C_p} - \omega_s L_1 \right) \quad (12)$$

From (10) and (11), power factor at the inverter output is derived as

$$\cos \phi = \cos(\phi_i - \phi_v) = \cos \left[\tan^{-1} \left\{ \frac{\pi^2}{8R_0} \frac{\omega_s^2 M_2}{\left(\omega_s L_1 - \frac{1}{\omega_s C_s} \right)} \right\} - \tan^{-1} \left\{ \frac{8R_0}{\pi^2} \frac{1}{\omega_s^2 M^2} \left(\frac{1}{\omega_s C_{st}} + \frac{1}{\omega_s C_{pt}} - \right. \right. \right. \quad (13)$$

From (12), it is clear that if C_{pt} and C_{st} are selected using (4), then the power factor at inverter output is leading and it gives soft switching at device turn OFF of inverter switches. Hence, the parallel capacitor C_{pt} is calculated using (12) such that at rated output power, inverter output is slightly lagging which enables soft switching at device turn ON. This modified value of capacitor C_{pt} for soft switching at device turn ON is slightly lower than value calculated using (4). From (10) and (11), VSI peak switch voltage and both peak and RMS current ratings are derived as

$$\hat{V}_{sw} = \hat{V}_{inv} \approx \frac{4}{\pi} \frac{V_0}{\omega_s M} \left(\omega_s L_1 - \frac{1}{\omega_s C_{st}} \right) \hat{I}_{sw} \approx \frac{\pi_2}{8} \omega_s^2 M C_{pt} I_0 \quad (14)$$

In order to calculate natural voltage gain of the dc-dc wireless power transfer stage, volt-sec balance is applied on the dc inductor L_d .

Average voltage (VA) at the input side node of Ld is same as input dc voltage V . Average voltage at the other side d of inductor, i.e., H-bridge inverter side is calculated as

$$V_B = \frac{1}{\pi} \int_{\phi}^{\pi+\phi} \widehat{V}_{inv} . \sin \omega t . dt = \frac{2\sqrt{2}}{\pi} V_{inv} \cos \phi = \frac{8}{\pi^2} \frac{V_0}{\omega_s M} X \left(\omega_s M - \frac{1}{\omega_s C_s} \right) \cos \phi \quad (15)$$

At steady state, both node voltages of inductor are same, i.e., VA = VB . Hence, natural voltage gain of dc-dc wireless power transfer stage is calculated using (14) as

$$\frac{V_0}{V_d} = \frac{\pi^2 \omega_s M}{8 \cos \phi} \frac{1}{\left(\omega_s L_1 - \frac{1}{\omega_s C_s} \right)} \quad (16)$$

From (15), it is clear that without presence of series capacitor C_s , the voltage gain of the converter would be

$$\frac{V_0}{V_d} = \frac{\pi^2}{8} \frac{1}{\cos \phi} \frac{M}{L_1} \quad (17)$$

Hence, another advantage of the proposed converter is improved voltage gain compared with conventional parallelseries compensated topology. In order to describe steady-state operation of one full switching cycle with all subintervals, the capacitor C_p is selected using (12) such that the inverter output power factor is lagging.

F. Steady state operation for one complete switching cycle Interval

Interval 1 $[t_0 - t_1 - t_2]$: Consider that at time instant t_0 , diagonal switch pair of the inverter S1 and S4 are in conduction as shown in Fig.3. Also consider that the inverter output voltage, V_{inv} leads the inverter output current, I_{inv} . During time interval, t_0 to t_1 switches S1 and S4 take the complete inverter current and off-diagonal switch pair S3 and S4 blocks a voltage with same magnitude as inverter output voltage, V_{inv} . The receiver coil current, I_r is positive during the interval t_0 to t_1 and diode pair D5, D8 conducts and feeds the output filter capacitor, C_o . At time instant t_2 receiver coil current,

I_r changes its direction and diode pair D6, D7 commutates and D5, D8 start conduction.

Interval 2 [$t_2 - t_3 - t_4$]: At time instant t_2 inverter output voltage, V_{inv} changes its direction since, the operating power factor at the output of the inverter is considered as lagging. At time instant t_3 gating signal of switch pair S3 and S2 become high but since, the voltage across these switches are negative, the other switch pair S1 and S4 keeps on conducting are shown in Fig.3. A negative voltage with same magnitude as V_{inv} appears across the switch diode pairs (S3, D3 and S2, D2) and the diodes D5 and D2 blocks this negative voltage. Throughout this time interval $t_2 - t_3 - t_4$ receiver side diode pair D6, D7 conducts.

Interval 3 [$t_4 - t_5$]: At time instant t_4 gating signal of switch S1 and S4 is removed and immediately switch S3 and S2 takes the inverter current. It is clear that to get the zero voltage switching, time lag of inverter output current, I_{inv} from output voltage V_{inv} has to be equal or more than the overlap time of the switches. The switch pair S3 and S2 keep on conducting in the interval t_4 to t_5 and the switch pair S1 and S4 keep on blocking the positive voltage. At this interval rectifier diode pair D6, D7 in the receiver side keeps on conducting. Interval 4 [$t_5 - t_6 - t_7 - t_8$]: At instant t_5 receiver coil current, I_r changes its direction and diode pair D5, D8 conducts and feeds the output filter capacitor, C_0 . At instant t_6 inverter output voltage, V_{inv} changes its direction and at time instant t_7 gating signal of switch pair S1 and S4 becomes high but since, the voltage across these switches are negative, the switch pair S3 and S2 keeps on conducting. At time instant t_8 gating signal of switch S3 and S2 is removed and immediately switch S1 and S4 takes the inverter current and this cycle repeats in every switching cycle.

3 WIRELESS PAD AND COIL DESIGN

While designing a wireless IPT system then main factor is Magnetic coupling. Generally, Inductive pads are classed as inductive pad without ferromagnetic core and IPT pad with ferromagnetic core, i.e., IPT pad with ux-guided material. Although the IPT pad with ferrite core is comparatively expensive, it has some advantages.

The major advantage of the ferrite core is that it minimizes the magnetic field emissions around the coils by reducing the stray or fringe fields since ferrites keep the magnetic field in between the coils. Various types of ux-guided IPT pads are discussed in [9], [12], [20] and [21][26]. Since the main objective of this paper is to examine the suitability of the proposed Voltage-fed topology in IPT, a UU-type core is selected for its simple structure and easy to implement. Considering that the main focus of this paper is power electronics and control and not on IPT couplers; hence, detail design is not carried out here. The coil design and size optimization of the pads are detailed in [20] and [25].

The change in self-inductance of IPT pad is very slow with change in pad misalignment, whereas the change in mutual inductance is much faster. It is clear that for the ux generated due to current flow in TC have two possible ux paths. One is through transmitter core - air-transmitter core and the other one is through transmitter core - air - receiver core - air-transmitter core. The reluctance of the second parallel ux path, i.e., via receiver core increases with the increase in air gap between TC and RC. Hence, this slight reduction in TC self-inductance with increase in air gap is observed. Also, increase in air gap directly leads to increase in reluctance to the mutual ux; therefore, mutual inductance reduces with increase in air gap [18], [27].

The predominant loss in these large air gap coils is the copper loss. From TC current and RC current expressions given in (6) and (7), the copper loss in the coils is calculated as

$$P_{cul} = \frac{8}{\pi^2} X \left(\frac{V_0}{\omega_s M} \right)^2 r_1 \quad (18)$$

$$P_{cul} = \frac{\pi^2}{8} X I_0^2 r_2 \quad (19)$$

where r_1 and r_2 are transmitter and RC ac resistances, respectively. Using (17) and (18), the coil-to-coil efficiency is calculated as

$$\eta_{coil} = \frac{1}{1 + \frac{8}{\pi^2} \frac{V_0}{I_0 (M\omega_s)^2} r_1 + \frac{\pi^2}{8} \frac{I_0}{V_0} r_2} \quad (20)$$

From (19), it is clear that coil-to-coil power transfer efficiency increases with increase in impedance due to mutual coupling ($M\omega_s$)

and reduction in coil ac resistance. For a given system, all parameters in (19) other than charging current are known. Hence, the optimum charging current for maximum coil-to-coil power transfer efficiency and corresponding maximum efficiency are given as

$$V_0 = \frac{\pi}{8} M \omega_s \left(\frac{r_1}{r_2} \right)^2 I_{0max} \quad (21)$$

$$I_{0max} = \frac{8}{\pi^2} \frac{V_0}{M \omega_s} \sqrt{\frac{r_1}{r_2}} \quad (22)$$

$$\eta_{coilmax} = \frac{1}{1 + 2 \frac{\sqrt{r_1 r_2}}{M \omega_s}} \quad (23)$$

4 SIMULATION RESULTS

The WPT circuit shown in Fig. 2 is simulated using PSIM 9.03. Simulation model for power ratings rated at 630 W was developed. The main application of the proposed converter is EV; therefore, to demonstrate the feasibility of the concept for high power, the proposed converter is simulated for 630-W power output. The IPT pad parameters, viz. self-inductances and mutual inductance values for 630-W system are selected from [7]. In order to control battery charging current, xed switching frequency and variable duty cycle modulation is adopted in the inverter. The current ratings of the inverter switches are almost same, whereas the voltage rating of the proposed topology is close to half of the conventional LC resonance tank network. Hence, the inverter VA rating is also almost half. Fig. 4 shows steady-state voltage and current waveforms of transmitter and RCs for 630-W power transfer. Equations (6) and (7) show that TC current I_t lags RC current I_r by 90 degree. Also, from (8), TC voltage V_t lags I_r . Since the mutual inductance is very low, the real part of (8) predominantly determines the voltage magnitude. Fig. 4 verifies this analytical prediction. With the given simulation parameters, calculated RMS value of V_t is 512 V for 630-W power output. Simulation result also gives the same value of V_t . Fig.4 shows that the TC voltage and current waveforms are very close to sinusoidal. It is because the parallel capacitor C_p provides

much lesser impedance to higher order harmonic currents than TC, and the TC predominantly receives fundamental component.

Fig. 5 shows the simulation results of the different voltage waveforms of the passive elements of transmitter-side tank network. Since the impedance of the capacitor C_s is half of the impedance of TC self-inductance at resonance frequency, hence, parallel capacitor C_p gets approximately half of the TC voltage V_t as derived in (8), (9), and (10). Simulation

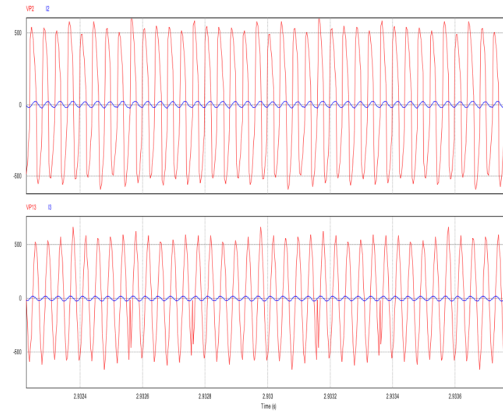


Fig.4. Simulation results: Steady-state waveforms of TC voltage v_t current i_t , and RC voltage v_r and current i_r

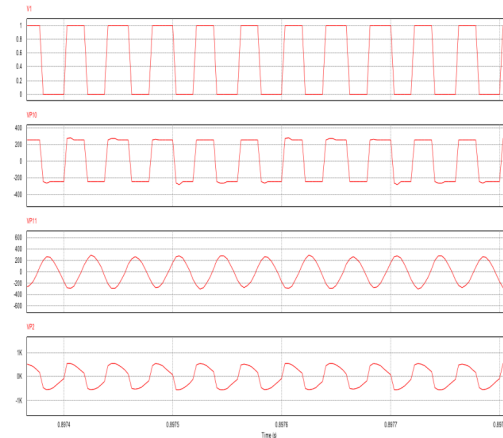


Fig.5. Simulation results: Steady-state waveforms of gating signal

of switch S1, inverter output voltage v_{inv} , series capacitor voltage v_{cs} and TC voltage

results given in Fig. 5 verify this mathematical analysis. This is a major advantage of this topology where the inverter switches get half of the TC voltage stress, whereas the switch voltage stress is exactly similar to TC voltage in case of conventional voltage-fed topology. Fig. 6 shows steady-state dc input voltage, input current waveforms and dc output voltage, output current waveforms for 630-W power output, and Fig. 6 shows voltage and current waveforms at diode bridge rectifier ac input for 630-W output, respectively.

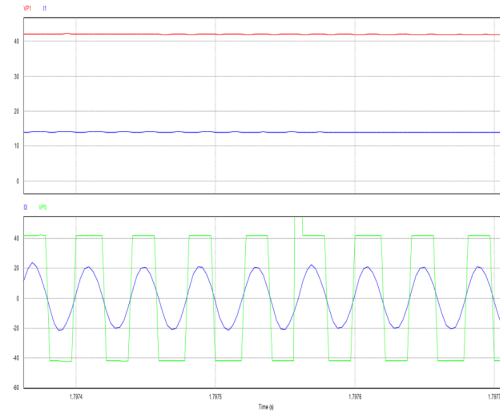


Fig.6. Steady-state waveforms for Output voltage v_b and current i_0 and Diode bridge rectifier input voltage v_{rD} and current i_d

Table I Selected simulation parameters

Parameters	Selected Values (for 630W)	Unit
Input dc Voltage, v_d	250	volt
DC link Capacitor, c_d	500	μF
Self-inductance, L_1, L_2	177, 177	μH
Mutual inductance, M	40	μH
Capcitors, C_p, C_s, C_r	485, 504, 252	nF
Switching Frequency, f_s	25	kHz
Battery voltage, v_b	42	volt
Output filter capacitor, C_0	350	μF
Coil turns ratio	1:1	

5 CONCLUSION

A new inductive wireless power transfer circuit using Voltage-fed converter is presented in the paper. This topology provides an alternative wireless power transfer solution for EV charging. Simulation results show that the required voltage in the transmitter coil is supplied by the parallel compensation capacitor. Capacitor in the series of the transmitter coil reduces voltage rating of the inverter switches. Soft switching at turnon of VSI switches ensures lower switching loss. Also, soft recovery of the rectifier diodes eliminates reverse recovery loss. All the analysis and simulation are verified using PSIM.

References

- [1] D. H. Tran, V. B. Vu, and W. Choi, Design of a High Efficiency Wireless Power Transfer System with Intermediate Coils for the Onboard Chargers of Electric Vehicles, *IEEE Transactions on Power Electronics*, vol. 8993, no. c, pp. 11, 2017.
- [2] S. Li and C. C. Mi, Wireless power transfer for electric vehicle applications, *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 3, no. 1, pp. 417, 2015.
- [3] W. Eberle and F. Musavi, Overview of wireless power transfer technologies for electric vehicle battery charging, *IET Power Electronics*, vol. 7, no. 1, pp. 6066, 2014.
- [4] U. Surendrakumaran and Alamelu Nachiappan., Performance analysis of wireless power transfer(WPT) through two-coil and three-coil structure, in *2017 International Conference on Innovative Research In Electrical Sciences (IICIRES)*, pp. 16, IEEE, jun 2017.
- [5] S. Samanta, A. K. Rathore, and D. J. Thrimawithana, Bidirectional Current-Fed Half-Bridge (C) (LC)(LC) Conguration for InductiveWireless Power Transfer System, *IEEE Transactions on Industry Applications*, vol. 53, pp. 40534062, jul 2017.
- [6] A. Kurs, A. Karalis, R. Moffatt, J. D. Joannopoulos, P. Fisher, and M. Soljacic, Wireless power transfer via strongly coupled

- magnetic resonances., Science, vol. 317, no. 5834, pp. 836, 2007.
- [7] O. C. Onar, J. Kobayashi, D. C. Erb, and A. Khaligh, A bidirectional high-power-quality grid interface with a novel bidirectional noninverted buck-boost converter for PHEVs, IEEE Transactions on Vehicular Technology, vol. 61, no. 5, pp. 20182032, 2012.
 - [8] S. Dusmez and A. Khaligh, A charge-nonlinear-carrier-controlled reduced-part single-stage integrated power electronics interface for automotive applications, IEEE Transactions on Vehicular Technology, vol. 63, no. 3, pp. 10911103, 2014.
 - [9] H. H. Wu, A. Gilchrist, K. D. Sealy, and D. Bronson, A high efficiency 5 kW inductive charger for EVs using dual side control, IEEE Transactions on Industrial Informatics, vol. 8, no. 3, pp. 585 595, 2012.
 - [10] A. P. Hu, J. T. Boys, and G. A. Covic, Frequency Analysis and computation of a current-fed resonant converter for ICPT power supplies, PowerCon 2000 - 2000 International Conference on Power System Technology, Proceedings, vol. 1, pp. 327332, 2000.
 - [11] U. K. Madawala and D. J. Thrimawithana, A bidirectional inductive power interface for electric vehicles in V2G systems, IEEE Transactions on Industrial Electronics, vol. 58, no. 10, pp. 47894796, 2011.
 - [12] P. Ning, J. M. Miller, O. C. Onar, C. P. White, and L. D. Marlin, A compact wireless charging system development, Conference Proceedings - IEEE Applied Power Electronics Conference and Exposition APEC, pp. 30453050, 2013.
 - [13] H. L. Li, A. P. Hu, and G. A. Covic, A direct AC-AC converter for inductive power-transfer systems, IEEE Transactions on Power Electronics, vol. 27, no. 2, pp. 661668, 2012.
 - [14] D. J. Thrimawithana and U. K. Madawala, A novel matrix converter based bi-directional IPT power interface for V2G

- applications, 2010 IEEE International Energy Conference and Exhibition, EnergyCon 2010, pp. 495500, 2010.
- [15] J. Shin, S. Shin, Y. Kim, S. Ahn, and S. Lee, Design and Implementation of Shaped Magnetic Resonance Based Wireless Power Transfer System for Roadway-Powered Moving Electric Vehicles, no. c, pp. 1 14, 2013.
 - [16] R. Huang and B. Zhang, Frequency, impedance characteristics and HF converters of two-coil and four-coil wireless power transfer, IEEE Journal of Emerging and Selected Topics in Power Electronics, vol. 3, no. 1, pp. 177183, 2015.
 - [17] S. Samanta and A. K. Rathore, Analysis and design of current-fed (L)(C) (LC) converter for inductive wireless power transfer (IWPT), in 2015 IEEE Energy Conversion Congress and Exposition (ECCE), no. L, pp. 57245731, IEEE, sep 2015.
 - [18] J. Kan, S. Xie, Y. Tang, and Y. Wu, Voltage-fed dual active bridge bidirectional DC/DC converter with an immittance network, IEEE Transactions on Power Electronics, vol. 29, no. 7, pp. 35823590, 2014.
 - [19] N. A. Keeling, G. A. Covic, and J. T. Boys, A unity-power-factor IPT pickup for high-power applications, IEEE Transactions on Industrial Electronics, vol. 57, no. 2, pp. 744751, 2010.
 - [20] J. Shin, S. Shin, Y. Kim, S. Ahn, S. Lee, G. Jung, S.-J. Jeon, and D. H. Cho, Design and Implementation of Shaped Magnetic Resonance Based Wireless Power Transfer System for Roadway-Powered Moving Electric Vehicles, IEEE Transactions on Industrial Electronics, vol. 61, no. 3, pp. 11791192, 2014.
 - [21] M. Budhia, G. A. Covic, and J. T. Boys, Design and optimization of circular magnetic structures for lumped inductive power transfer systems, IEEE Transactions on Power Electronics, vol. 26, no. 11, pp. 30963108, 2011.
 - [22] C. Park, S. Lee, G. H. Cho, and C. T. Rim, Innovative 5-m-off-distance inductive power transfer systems with optimally

- shaped dipole coils, *IEEE Transactions on Power Electronics*, vol. 30, no. 2, pp. 817827, 2015.
- [23] M. Budhia, J. T. Boys, G. A. Covic, and C. Y. Huang, Development of a single-sided ux magnetic coupler for electric vehicle IPT charging systems, *IEEE Transactions on Industrial Electronics*, vol. 60, no. 1, pp. 318328, 2013.
- [24] Z. N. Low, R. A. Chinga, R. Tseng, and J. Lin, Design and test of a high-power high-efficiency loosely coupled planar wireless power transfer system, *IEEE Transactions on Industrial Electronics*, vol. 56, no. 5, pp. 18011812, 2009.
- [25] S. Y. Choi, B. W. Gu, S. Y. Jeong, and C. T. Rim, Advances in wireless power transfer systems for roadway-powered electric vehicles, *IEEE Journal of Emerging and Selected Topics in Power Electronics*, vol. 3, no. 1, pp. 1836, 2015.
- [26] A. Khaligh and S. Dusmez, Comprehensive topological analysis of conductive and inductive charging solutions for plug-in electric vehicles, *IEEE Transactions on Vehicular Technology*, vol. 61, no. 8, pp. 3475 3489, 2012.
- [27] U. R. Prasanna and A. K. Rathore, Small-signal modeling of activeclamped ZVS current-fed full-bridge isolated DC/DC converter and control system implementation using PSoC, *IEEE Transactions on Industrial Electronics*, vol. 61, no. 3, pp. 12531261, 2014.