Topology study for an Inductive Power Transmitter for Cordless Kitchen Appliances

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Abstract—In a conventional kitchen an appliance may require over 2.4kW of power during normal operation. In a cordless kitchen the used appliances are cordless, so this power needs to be transferred over a distance of at least 4cm, at high efficiency. Therefore an inductive power system needs to be installed with an inductive power transmitter below the table countertop and an inductive power receiver inside the appliance. Next to the inductive power system the cordless appliance is also equipped with a communication channel. There are many transmitter topologies which can satisfy the functional requirements (like amount of power transferred, high efficiency, etc.) of the inductive power system. But satisfying the non-functional requirements (like EMI) most often takes several design iterations. Presently, a transmitter based on a series resonance principle will be the first choice of the designer because it is relatively simple to understand and can satisfy the functional requirements of the system easily. The aim of the paper is to describe the present choice for the transmitter and to explain a new transmitter topology that can intrinsically minimize the problems associated with the present choice. The paper proceeds with a brief introduction about the cordless kitchen concept, in section I. Section II gives the description and analysis of the presently used topology in the system. The requirements of a new transmitter topology are explained in section III and a new transmitter topology along with its analysis, is proposed in Section IV. Section V explains the experimental measurements and gives the comparison between the present and the proposed topologies. The conclusion is stated in section VI.

I. INTRODUCTION

An alternating current flowing through a coil produces its own alternating magnetic field. If another coil is placed in the proximity of the first one then, this alternating magnetic field gets linked to it. Such an arrangement of the coils is called as an inductive coupler. In a cordless kitchen the inductive coupler consists of a transmitter coil and a mutually coupled receiver coil. The transmitter coil is located beneath the counter top of the kitchen table, whereas, the receiver coil is located inside an electrical appliance, as shown in Figure 1.

As shown in Figure 1, the system operates with the help of two channels: a power channel and a communication channel. The power transfer is executed by the power channel, whereas the information or data exchange is done using the communication channel. The power transfer and the communication is done at a coil to coil distance of 4cm. One of the operating principles for the power channel is Inductive Power Transfer (IPT). This principle is used in kitchen appliances which are equipped with a motor or a resistive heating element. A food processor or an oven are good examples of appliances working according to the IPT principle. The basic block diagram of the IPT power channel is shown in Figure 2. The IPT power channel consists of an AC source, an AC-AC conversion stage, a resonant inductive coupler, an AC-DC conversion stage (optional hence shown as dotted block) and a load. The AC-AC conversion stage along with the transmitter coil of the resonant inductive coupler is called the transmitter (Tx). The receiver coil of the resonant inductive coupler along with the AC-DC conversion stage (optional) and the load is called the receiver (Rx). The transmitter and the receiver coils are mutually coupled to each other with a certain coupling factor (k). The mains voltage is converted to a high frequency AC voltage by the AC-AC conversion stage. Due to the inductive coupling, the high frequency AC voltage is induced on the receiver side. This induced voltage can be converted to a DC voltage with the help of AC-DC conversion stage. The load can be a DC motor or simply a resistor.

In order to get the maximum system efficiency, the principle of resonance is used in the inductive coupler. To achieve this resonant nature, different combinations of interconnections of both the coils (L) and the capacitors (C) can be used. These different combinations are known as topologies. Each topology has its own operating principle with advantages as well as disadvantages, depending on the application.
II. PRESENT TOPOLOGY

Figure 3 shows the topology which is used in the present inductive power system. This system has the series resonance principle applied in the transmitter as well as in the receiver side.

![Fig. 2. IPT with motor](image)

![Fig. 3. Series resonant topology](image)

A. Description of present topology

In Figure 3, $V_{in}$ is the DC voltage obtained after rectifying the mains voltage. As mentioned earlier, the transmitter is nothing but the AC-AC conversion stage comprising $V_{in}$, an inverter stage and a resonant tank comprising of the series connection of the transmitter coil ($L_1$) and a capacitor ($C_1$). On the other hand, the receiver comprises of the receiver coil ($L_2$) and a capacitor ($C_2$) connected in series with the load ($R_L$).

If the transmitter (Tx) is powered, in the absence of the coupling ($k$), it has its natural resonant frequency ($f_{tx}$). Similarly, in the absence of the coupling ($k$), the receiver (Rx) has its natural resonant frequency ($f_{rx}$). In practice, the $f_{tx}$ and $f_{rx}$ are chosen to be equal [3] and the coupling factor ($k$) between the transmitter coil ($L_1$) and the receiver coil ($L_2$) is in the order of 0.1 to 0.35. Therefore, the whole inductive power system has its natural resonant frequency or frequencies ($f_{res}$). For low coupling factors the $f_{tx}$ and $f_{rx}$ are merged into a single resonant frequency $f_{res}$. For higher coupling factors the $f_{tx}$ and $f_{rx}$ frequencies are moved away from each other, resulting in $f_{res1}$ and $f_{res2}$. The inverter stage of the transmitter can be operated at different operating points ($f_{op}$). Choosing the optimal $f_{op}$ for the inverter is an important task because the high frequency AC voltage is applied to the transmitter coil, with the help of the power switches. Thus, one has to make sure that Zero Voltage Switching (ZVS) for the power switches is applied, avoiding high dV/dt’s in the inverter stage. Therefore, in this case, the optimal $f_{op}$ is chosen where the efficiency is maximum and input current ($I_{L1}$) is slightly lagging the inverter bridge voltage ($V(LH, RH)$). With the help of the mathematical analysis, one can determine the optimal operating frequency for the inverter stage of the transmitter, using the particular values of the circuit components shown in Figure 3 [2]. In the present topology, with the help of a frequency control loop the $f_{op}$ can be varied to achieve the maximum efficiency for the power transfer. Furthermore, at a particular $f_{op}$ the amount of power transfer can be controlled with the help of the duty (D) cycle control.

B. Analysis of present topology

The mathematical analysis along with time domain and frequency domain simulations of the series resonant topology is discussed in this section. The mathematical analysis can be used to derive the different transfer functions. The frequency domain simulations can be used to observe the bode plots of important parameters like impedance, input voltage (or current), output voltage (or current), etc. The time domain simulations can be used to observe the instantaneous voltages across and current through the different circuit components. For the time domain simulations, it should be noted that as the present topology is voltage source driven, there should be a ‘dead time’ between the switches of the same bridge leg. To proceed with the analysis following (realistic) values are chosen:

- $f_{tx} = f_{rx} = 33$ kHz
- $L_1 = 290$ uH
- $L_2 = 290$ uH
- $C_1 = 1/(4\pi^2 f_{tx}^2 L_1)$ F
- $C_2 = 1/(4\pi^2 f_{rx}^2 L_2)$ F
- $k = 0.35$
- $R_L = 10$ Ω
- $V_{in} = 325$ V DC (for time domain)
- $V_{ac} = 1$ V AC (for frequency domain)

1) Mathematical analysis: From Figure 3, a simplified equivalent circuit can be derived for the mathematical analysis as shown in Figure 4.

In the above circuit, $R_1$ represents the equivalent series resistance of the transmitter coil ($L_1$) along with the equivalent series resistance of the capacitor $C_1$. $R_2$ represents the load resistance along with the equivalent series resistance of the receiver coil ($L_2$) and the equivalent series resistance of the capacitor $C_2$. $V_{i12}$ represents the voltage induced in the transmitter coil due to the mutual inductance (say, $M$) and the current flowing through the receiver coil ($i_2$). $V_{21}$ represents the voltage induced in the receiver coil due to the mutual inductance and current flowing through the transmitter coil ($i_1$).

With the help of Laplace transform and transformer theory, the voltage transfer function can be derived [2]. From the
above equation, it should be noted that the order of characteristic equation (denominator) is four. Consequently, it is called as 4th order system. The roots of the characteristic equation converges to the resonance frequencies.

With the help of the derived transfer function, frequency domain simulations can be performed. Figure 5a shows the bode plot for the power delivered to $R_2$ whereas, Figure 5b shows the variation of the efficiency w.r.t. operating frequency. Figure 5a shows two resonant peaks. Thus, the transmitter can be operated at, or near two resonant frequencies for the maximum power transfer. There are two falling edges and two rising edges. The frequency control loop can operate either on the falling edge or on the rising edge. In any case, there are two operating points at which the gain is the same but the efficiency can be different.

From the marked crosses in Figure 5, it can be seen that the operating point for the maximum power gain and maximum efficiency is not the same. The maximum efficiency point lies in between the two resonant peaks.

2) Detailed analysis: As mentioned earlier, for this system the optimal operating point, where the efficiency is maximum and input current ($I_{L_1}$) is slightly lagging the inverter bridge voltage (V(LH,RH)). can be calculated with the help of mathematical analysis as shown in [2]. For this topology and the calculated values of circuit components, the optimal operating point ($f_{op}$) can be calculated to be 32.7kHZ[2] and the frequency control loop operates on the falling edge. At the calculated $f_{op}$, the time domain simulations can be performed.

Figure 6 shows the voltage across the inverter stage of the transmitter (V(LH,RH)) and the input current ($I(L_1)$). The lagging current indicates the system operating in an inductive mode. The inductive mode of operation is always preferred to get the advantage of Zero Voltage Switching (ZVS) of the power switches. ZVS of the power switches improves the efficiency of the system by minimizing the switching losses. Due to ZVS the voltage across the switches rises steadily, (i.e. no high dV/dt’s) minimizing the EMI problem. As mentioned earlier, at a particular $f_{op}$ the duty cycle (D) control is used to control the power delivered to the load. Figure 6 shows the waveforms for D=100%. Figure 7 shows the same waveforms but at D=50%. It can be observed from Figure 6 and Figure 7 that, introduction of duty cycle control results in loss of ZVS. The loss of ZVS causes high dV/dt on the corresponding node.

Figure 8 shows the voltage across the transmitter coil, when $f_{op} < f_{res1}$, $f_{op}=f_{res1}$ and $f_{op} > f_{res1}$. It can be clearly noted that high dV/dt’s are observed in all the three cases. These high dV/dt’s may create a changing electric field between the node which carries the high dV/dt and surrounding conducting components, which has some kind of capacitive coupling to the environment. This can cause common mode currents to flow from the system to the environment and
back via the mains, causing common mode noise emission, as shown in Figure 9.

There are many protective means of reducing the common mode noise but, often these require additional and expensive components, like filters.

Another important point to note, is the voltage across the inverter power switches. From Figure 3, the peak amplitude of $V_{in}$ can be expected and observed across the inverter power switches, irrespective of the load changes. Therefore, one can state that:

$$V_{sw_{\text{peak}}} = V_{\text{mains}_{\text{peak}}}$$

### III. REQUIREMENTS OF THE SYSTEM

In the previous section, the presently used series resonant topology is discussed in detail. It is capable of transferring the power at high levels (kW) at high efficiency but it is difficult to meet the non-functional requirements of the inverter stage, i.e. EMI. In the cordless kitchen system, the transmitter (Tx) remains the same and the receiver (Rx) changes, depending on which appliance is located on the table top. Thus, it is beneficial if the drawback of the cordless kitchen system is minimized by modifying only the transmitter side. Thus, in the further studies different topologies for the transmitter are analyzed keeping the receiver part the same. The basic requirement from the new transmitter topology is to meet all the non-functional requirements intrinsically without compensating for the efficiency.

As mentioned earlier, the duty cycle control might result in the loss of ZVS of the power switches resulting in high dV/dt’s at the bridge nodes. The change in the input current w.r.t. time (dI/dt) as well as the change in the voltage across the transmitter coil w.r.t. time (dV/dt) may create electromagnetic interference (EMI) problems. This gives rise to the first requirement of the system: having a low EMI.

Thus, one can summarize the requirements for the transmitter topology as, functional requirements:

- Power transfer up to 2.4 kW
- Efficiency > 90%
- Simple control system

and the most important non-functional requirement:

- Low EMI

Based on these requirements a new topology is identified and analyzed in further sections.

### IV. PROPOSED TOPOLOGY

Figure 10 shows the proposed topology which might be easier to meet the non-functional requirements. An attempt is made to meet the functional and non-functional requirements by modifying the transmitter side. And the best suitable topology is presented further in this section.

#### A. Description of Current Source parallel resonant topology

Figure 10 shows the current source (CS) parallel resonant topology. $L_{\text{choke}}$ is placed at the DC side of the inverter stage to realize a current source driven parallel resonant inductive coupler. It should be noted that, as it is a current source driven topology, there should be some ‘overlap time’ between the switches of the same bridge leg. Again, to analyze the properties of this topology, mathematical analysis together with the simulations in the time and frequency domain are performed. It can be observed that, due to the presence of $L_{\text{choke}}$, it is harder to perform the frequency domain simulations for this topology. One can reduce the circuit with the proper mathematical analysis and proceed with the further evaluation. In this case, to proceed with the analysis following (realistic) values are chosen:

- $f_{tx} = f_{rx} = 33$ kHz
- $L_1 = 290$ uH
• \( L_2 = 290 \text{ uH} \)
• \( C_1 = 1/(4\pi^2 f_{tx}^2 L_{res}) \text{ F} \)
• \( C_2 = 1/(4\pi^2 f_{tx}^2 L_2) \text{ F} \)
• \( L_{choke} = 1 \text{ mH} \)
• \( L_{res} = L_1||2L_{choke} \)
• \( k = 0.35 \)
• \( R_L = 10 \Omega \)
• \( V_{in} = 325 \text{ V DC (for time domain)} \)

### B. Analysis of CS parallel resonant topology

1) Mathematical analysis: The topology is based on the principle of parallel resonance in which the circuit offers maximum impedance at the resonant frequencies.

For this topology and calculated circuit component values, the resonant frequency of the whole system \( f_{res1} \) is observed to be 28.2kHz. With the help of these calculated circuit component values, the time domain simulations of the circuit shown in Figure 10 can be performed. Figure 11 shows the voltage across the transmitter coil of the resonant inductive coupler when the inverter stage is operated at \( f_{op}<f_{res1}, f_{op}=f_{res1} \) and \( f_{op}>f_{res1} \).

![Fig. 11. Voltage across transmitter coil (a)\( f_{op}<f_{res1}, (b)\)\( f_{op}=f_{res1} \) and (c)\( f_{op}>f_{res1} \)](image)

Interestingly from Figure 11, it can be seen that when the inverter stage of the transmitter is operated at or below the \( f_{res1} \), high dV/dt’s are not observed. On the other hand, when the inverter stage is operated above the \( f_{res1} \), high dV/dt’s are observed. Therefore, based upon the requirements stated in the earlier section, the operation of this topology at or below \( f_{res1} \) is preferred. When the system is operated below \( f_{res1} \), the transmitter coil voltage waveform is almost a sine wave. It has an event near the zero crossing of the sinewave where the voltage is almost zero (say \( T_{freeze} \)). Thus, for the analysis of the system operating at or below \( f_{res1} \), First Harmonic Approximation (FHA) can be used [1]. With the help of FHA, a simplified equivalent circuit of the system can be derived as shown in the Figure 12.

The FHA allows to replace the current source \( (I_{ac}) \) and the capacitor \( (C_1) \) with an AC voltage source \( (V_{ac}) \). The transmitter capacitor \( (C_1) \) does not influence the receiver coil voltage gain but it does affect the phase of the input current \( (I_1) \). Thus, the capacitor \( (C_1) \) can be regarded as a phase compensator for the input current drawn from the AC voltage source \( (V_{ac}) \). In the equivalent circuit shown in Figure 12, \( R_1 \) represents the equivalent series resistance of the transmitter coil \( (L_1) \), equivalent series resistance of the choke coil \( (L_{choke}) \) along with the equivalent series resistance of the capacitor \( (C_1) \). \( R_2 \) represents the load resistance along with the equivalent series resistance of the receiver coil \( (L_2) \) and the equivalent series resistance of the capacitor \( (C_2) \). \( V_{12} \) represents voltage induced in the transmitter coil due to the mutual inductance (say, \( M \)) and the current flowing through the receiver coil \( (i_2) \). \( V_{21} \) represents the voltage induced in the receiver coil due to the mutual inductance and the current flowing through the transmitter coil. By applying the Laplace transformation and the transformer theory, a voltage transfer function can be derived as shown below.

\[
\frac{V_{21}}{V_{ac}} = \frac{s^2(R_2MC_2) + sM}{s^2(L_1sC_1 - M^2C_2) + s^2(L_3sC_2 + L_{out}C_2R_2) + s(R_1R_2C_2 + L_{out}) + R_1} \quad (1)
\]

The above equation shows that the order of the characteristic equation (denominator) is 3. The more poles in the system, the more difficult it is to control. Again interestingly, it implies that the proposed topology is simpler to control than the present topology. Therefore, in the further analysis, the operation of the current source parallel resonant topology above the \( f_{res1} \) is discarded and the operation at or below the \( f_{res1} \) is considered.

2) Detailed analysis: As mentioned earlier, for this topology and calculated circuit component values, the resonant frequency \( (f_{res1}) \) is observed to be 28.2kHz where \( T_{freeze} \) is zero. Therefore the system is operated at or below 28.2kHz, and other parameters like the voltage across switches, behaviour on load disconnection etc. are examined in this section.

With the help of time domain simulations of the circuit shown in Figure 10, at \( f_{op} \) say 27kHz, the voltage across all the circuit components, in steady state conditions can be observed.

Figure 13 shows the voltage observed across the choke coil \( (L_{choke}) \) and the transmitter coil. It shows that there are no high dV/dt’s present in any of the coils on transmitter side, under normal operating conditions. This is beneficial from the EMI point of view.

As shown in Figure 11a, when the inverter stage of the transmitter is operated below \( f_{res1} \), the voltage across the
transmitter coil appears to be a discontinuous function. With the help of time domain simulations, it can be observed that the peak voltage across the transmitter coil ($V_{tx}$) and $T_{freeze}$ increases with decrease in operating frequency. Therefore, it can be stated that the amplitude of the first harmonic also changes with the change in operating frequency ($f_{op}$). And thus, it is comparatively difficult to perform the frequency domain simulations for the equivalent circuit shown in Figure 12. For better results from the frequency domain simulations, one has to model the source $\hat{v}$ such that the amplitude of the first harmonic changes according to the operating frequency. The relation between the amplitude of the first harmonic ($V_{ac FHA}$) and the operating frequency ($f_{op}$) can be derived with the help Fourier analysis of the transmitter coil voltage waveform shown in Figure 14.

From Figure 14 the relation between the peak amplitude ($V_{tx}$) and operating frequency ($f_{op}$) can be derived and is as shown in the equation below:

$$V_{tx} = f(f_{op}, f_{res})$$  \hspace{1cm} (2)

And, at $f_{op} = f_{res}$, the peak amplitude can be given by:

$$V_{tx_{res}} = \frac{V_{in} \pi}{2}$$  \hspace{1cm} (3)

And with the help of Equation 2, the amplitude of the first harmonic ($V_{ac}$) can be given by:

$$V_{ac FHA} = f(V_{tx}, f_{op}, f_{res})$$  \hspace{1cm} (4)

$$V_{ac FHA} = \frac{V_{ac FHA}}{V_{tx_{res}}}$$  \hspace{1cm} (5)

The normalized value ($\hat{V}_{ac FHA}$) can be used in Equation 1 for the frequency domain simulations. Figure 15 shows the variation in normalized $V_{ac}$ with respect to operating frequency.

It should be noted that the FHA is valid only at or below $f_{res}$. Therefore, the whole mathematical model for the frequency domain analysis is valid only at or below $f_{res}$.

Another important point to observe is the switch voltage. It can be seen from Figure 16 that, the peak voltage ($V_{sw_{peak}}$) appearing across the switches is more than $V_{in}$. With the help of previously derived relation of the peak amplitude ($V_{tx}$), it can be calculated that $V_{sw_{peak}}$ is $\pi/2$ times the $V_{in}$ at resonance, and it increases further with decrease in the $f_{op}$, as shown in Figure 16.
the experimental validations. The topology for the system remains the same however, as the power requirement is scaled down (from 2.4kW to 60 W), one should also scale down the circuit component values. Therefore a new load resistance ($R_{L_{\text{new}}}$), a new operating frequency ($f_{\text{op}_{\text{new}}}$) and the new component values for the inverter stage of the transmitter has been found out using the scaling formulae. The new chosen values of $R_{L_{\text{new}}}$ and $f_{\text{op}_{\text{new}}}$ along with the are given below [4]:

$$
R_{L_{\text{new}}}=4\,\Omega \\
f_{\text{op}_{\text{new}}}=135\,kHz \\
L_{\text{new}}=24\,\mu H \\
C_{\text{new}}=41.6\,nF
$$

After scaling one can compare the behavior of the scaled down system against the behavior of the system with non-scaled component values as mentioned in previous sections. An example of scaling down the CS parallel resonant system is shown in Figure 17. One can observe that the scaled down system behaves exactly the same but at higher frequencies. Hence to verify the calculated and simulated results, a new prototype with the proposed topology was built. Figure 18 shows an image of the PCB with a transmitter prototype platform with a transmitter coil connected to it.

The EMC tests according to the standard CISPR14 were performed on the present system (series resonant topology) as well as the new prototype (CS parallel resonant topology). Figure 19 and Figure 20 shows the EMC measurements recorded for the scaled down series resonance topology and CS parallel resonance topology, respectively.

Both the figures indicates a pass for the maximum allowed noise level for the Quasi Peak (QP) noise measurement according CISPR14 (uppermost line) and maximum allowed noise level for the average noise measurement according CISPR14 (middle line).

However as per expectations, from Figure 20, a considerable reduction (10 dBuV) in the high frequency noise level can be observed in case of the CS parallel resonant topology. Therefore, the size of the mains filter can reduced.

From the detailed analysis explained in section II and experimental results, various factors can be noted down, as shown in table I.

Table I shows that, though the series resonant topology is easy to implement and widely used, it possesses some
TABLE I
COMPARISON OF DIFFERENT TOPOLOGIES

<table>
<thead>
<tr>
<th></th>
<th>Series topology</th>
<th>CS parallel topology (in proposed region)</th>
</tr>
</thead>
<tbody>
<tr>
<td>High dV/dt in transmitter coil</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>Voltage across power switches</td>
<td>( V_{in} )</td>
<td>((\pi/2)\sqrt{2}V_{in} ) or more</td>
</tr>
<tr>
<td>Use of duty cycle control</td>
<td>Possible</td>
<td>Not possible</td>
</tr>
<tr>
<td>Loss of ZVS</td>
<td>Possible</td>
<td>Not possible</td>
</tr>
<tr>
<td>Control</td>
<td>Comparatively difficult</td>
<td>Comparatively simpler</td>
</tr>
</tbody>
</table>

drawbacks when applied in the cordless kitchen system. It has advantages like the use of the low voltage switches, easy power control but, the EMI performance is relatively poor.

It can be observed that CS parallel topology is more suitable topology for the cordless kitchen system, when it comes to satisfy the non-functional requirements. High dV/dt’s do not occur, which offers good electromagnetic compatibility. The only disadvantages of the proposed topology are the high voltage rating of the power switches (more than \( 2V_{in} \)) and the loss of control over the delivered power, as duty cycle control cannot be used. One can make use of the varying input voltage (\( V_{in} \)) to control the delivered power.

VI. CONCLUSION

The paper explained the basic principle of cordless kitchen system along with its requirements. There are functional as well as non-functional requirements. The present series resonant topology for the transmitter has the drawback that it is harder to meet the non-functional requirements. With the help of mathematical analysis, simulations as well as the experimental results, the paper suggested the ‘current source parallel resonant’ topology for the cordless kitchen system which can meet the non-functional requirements more easily. To achieve the control over the delivered power to the load, one can employ a pre-power stage on the transmitter side or a post-power stage on the receiver side. Thus, it can be concluded that for the cordless kitchen system, the operation of the current source parallel resonant topology at or below resonant frequency of the system is a good alternative.

REFERENCES