CHAPTER 2

SIMULATION AND EXPERIMENTAL INVESTIGATION
OF THE LCL AND LCC RESONANT INVERTERS
AND LCL RESONANT CONVERTER

2.1 INTRODUCTION

Any independent system like aircraft and space systems depend on the battery/solar collectors for its power requirement. Hence the need for DC-DC converters and DC-AC inverters are vital in these applications. Also harvesting of renewable solar energy using distributed power systems (DPS) also require above DC-DC converters and DC-AC inverters to supply for its loads. These applications need high efficiency power conversion techniques for both converters and inverters.

Above applications require high frequency, high efficiency conversion systems thus making resonant inverters more ideal. Resonant inverters have been recognized as the next generation power conversion circuits that can best suit for the above requirements because of their low component costs, small component sizes and high efficiency.

Also the LCL and LCC Resonant Full Bridge Converters (RFB) are new, high performance DC-DC converters which cater above requirements. Also owing to their small size and lightweight they could find their place in the above applications. The common limitation of two element resonant
topologies can be overcome by adding a third reactive element termed as modified series resonant converter (SRC).

This chapter consists of two parts, one dealing with LCL and LCC Resonant Inverters and the other briefing the LCL Resonant converter.

### 2.2 LCL AND LCC RESONANT INVERTERS

Recent research by several authors has shown that LCL topology resonant inverters have a number of desirable features as compared to series or parallel topologies. Designing an LCL-type resonant inverter that can operate under resistive and reactive loads is still a challenging process. Efforts to improve the efficiency under varying loads have shown good promise. It is intended to focus primarily on designing a LCL topology to handle varied loads while maintaining constant output voltage with low total harmonic distortion (THD). While LCL topologies have been designed and tested for various applications and intend to replace the FPGA phase shifter circuitry with a microcontroller. The microcontroller approach is ideal considering the rising cost of high frequency FPGAs. A High-Frequency DC/AC Power Inverter is proposed which is capable of supplying power to an AC bus with varying loads.

Zero voltage switching (ZVS) at full load ensures very low losses in the bridge circuit while at reduced loads the mode of operation in the MOSFET switches in one arm of the bridge is zero current switching (ZCS). By adjusting the resonant frequency of the LCL tank, it is intend to maximize the range of ZVS and therefore increase the total efficiency of the bridge.
Figure 2.1 LCL resonant inverter simulation circuit

Figure 2.1 shows a simulation model of LCL resonant inverter. The proposed resonant tank consisting of three reactive energy storage elements (LCL) unlike the conventional resonant converter that has only two elements. The first stage converts a DC voltage to a high frequency AC voltage. The major advantages of this series link load SRC are that the resonating properties of the elements block the DC supply voltage and also there is no commutation failure if MOSFET’s are used as switches. Moreover as the DC current is absent in the primary side of the transformer, there is no need for current balancing. Another advantage of this circuit is that the device currents are proportional to load current. This increases the efficiency of the converter at light loads because of the reduction in the device losses.

2.2.1 Specifications of LCL and LCC Resonant Inverters

Input voltage 40-60V DC
Output power -250 VA
Output voltage -28 VRMS
Output current -10 A
Switching frequency -20 kHz
Minimum load unity power factor, 0.9 leading, 0.9 lagging

Minimum load efficiency- 80%

Full load efficiency -90%

\[ f_{sw} = 20\, kHz \]

\[ f_r = 19\, kHz \]

\[ D = 1 \]

\[ V_{dc} = 48\, volts \]

\[ \omega_r = 2\pi f_r \quad (2.1) \]

\[ y = \frac{f_{sw}}{f_r} \quad (2.2) \]

\[ y = 1.053 \]

\[ k = \frac{38}{28} = 1.357 \]

### 2.2.2 Design of LCL and LCC Resonant Inverters

The application for the LCL resonant inverter is constructed here for a high frequency distributed bus. Choice of the LCL resonant tank components is a difficult task since it involves numerous calculations and recalculation.

\[ V_{in} = \frac{2}{\pi} \sqrt{2} V \sin \left( \frac{D \cdot \pi}{2} \right) \]

\[ V_{gain} = \frac{2}{\pi} \sqrt{2} \frac{V}{y} \left( 1 + \frac{1}{Cn} - \frac{y}{Cn} \right) \sin \left( \frac{D \cdot \pi}{2} \right) \quad (2.3) \]

\[ V_{out} = V_{in} |V_{gain}| \]
\[ V_{in} = 43.215V \]
\[ C_r = 2 \]
\[ Q_s = 4 \]
\[ |V_{gain}| = 0.873 \]
\[ V_{out} = 37.729 \]
\[ R_{\text{load}} = 2.842 \Omega \]
\[ \text{rating} = 271.902W \]

\[ L_s = \frac{1}{\omega \cdot r \cdot C_s} \] \hspace{1cm} (2.4)
\[ C_s = 400 \times 10^{-6} F \]
\[ C_s = 0.4 \times 10^{-6} F \]

\[ \text{rating} = \frac{V_{out} \cdot V_{out}}{k \cdot k \cdot R_{\text{load}}} \] \hspace{1cm} (2.5)
\[ L_s = 1.754 \times 10^{-4} H \]
\[ \omega_{sw} = 2\pi f_{sw} \] \hspace{1cm} (2.6)

The first step was to calculate arbitrary values of series inductance and series capacitance based on the resonant frequency that was required for this application. A detailed study of the reference papers, it was realized that a more rigorous method of choosing LCL tank values were needed. At this point, MathCAD was utilized to calculate the LCL resonant tank gain and the output voltage at various stages of the circuit. With this in mind, the remaining LCL tank values were calculated using proven design parameters.

\[ T_{sw} = \frac{1}{f_{sw}} \] \hspace{1cm} (2.7)

\[ X_{\text{load}}(n) = n \cdot \omega_{sw} \cdot L_{\text{load}} \] \hspace{1cm} (2.8)
Inherent Inductance  \[ X_{\text{load}} (1) = 0.729 \Omega \]

\[ RL_{\text{load}} (n) = R_{\text{load}} + \frac{i.X_{\text{load}} (n)}{k^2} \]  \hspace{1cm} (2.9)

Primary Side load  \[ |RL_{\text{load}}(1)| = 2.87 \Omega \]

\[ X_{LP}(n) = \frac{1}{\omega_{sw} n \frac{L_p}{K^2}} \]  \hspace{1cm} (2.10)

\[ \varphi(n) = a \tan \left( \frac{\text{Im}(Zeq(n))}{\text{Re}(Zeq(n))} \right) \]  \hspace{1cm} (2.11)

\[ X_{cs}(n) = \frac{1}{\omega_{sw} n C_s} \]  \hspace{1cm} (2.12)

\[ Z_{s}(n) = i.X_{l}(n) - i.X_{cs}(n) \]  \hspace{1cm} (2.13)

\[ X_{i}(n) = \omega_{sw} n L_i \]  \hspace{1cm} (2.14)

\[ Z_{eq}(n) = Z_{s}(n) + Z_{parallel}(n) \]  \hspace{1cm} (2.15)

\[ Z_{parallel}(n) = \frac{-iX_{LP}(n).RL_{\text{load}}(n)}{\left( (-iX_{LP}(n)) + (RL_{\text{load}}(n)) \right)} \]  \hspace{1cm} (2.16)

\[ V_{AB}(n, t) = \left\{ \frac{4V}{n \pi} \sin \left( n \cdot \frac{\pi}{2} \right) \sin \left( D \cdot n \cdot \frac{\pi}{2} \right) \sin \left( n \cdot \omega_{sw} t \right) \right\} \]  \hspace{1cm} (2.17)

\[ V_{\text{rms}}(n) = \left( \frac{2}{T_{sw}} \int_0^{T_{sw}} V_{AB}(n,t)^2 \, dt \right)^{\frac{1}{2}} \]  \hspace{1cm} (2.18)
\[ V_{AB\text{rms}}(1) = 43.215V \]
\[ V_{AB\text{rms}}(3) = 14.405V \]
\[ V_{AB\text{rms}}(5) = 8.643V \]
\[ V_{AB\text{rms}}(7) = 63.174V \]
\[ V_{AB\text{rms}}(9) = 4.802V \]
\[ V_{AB\text{rms}}(11) = 3.929V \]
\[ V_{AB\text{rms}}(13) = 3.324V \]

\[ I_s(n,t) = \frac{V_{ab}(n,t)}{Z_{eq}(n,t)} \] (2.19)

\[ V_{LP}(n,t) = \frac{I_s(n,t)Z_{\text{parallel}}(n)}{k} \]

\[ V_{LP \text{ rms }} \left( n \right) = \left( \frac{2}{T_{SW}} \int_0^{T_{SW}} V_{LP}^2 \left( n, t \right)^2 \, dt \right)^{\frac{1}{2}} \] (2.20)

| \[ |V_{LP\text{rms}}(1)| = 24.781V \] |
| \[ |V_{LP\text{rms}}(3)| = 0.581V \] |
| \[ |V_{LP\text{rms}}(5)| = 0.248V \] |
| \[ |V_{LP\text{rms}}(7)| = 0.166V \] |
| \[ |V_{LP\text{rms}}(9)| = 0.124V \] |
| \[ |V_{LP\text{rms}}(11)| = 0.079V \] |
| \[ |V_{LP\text{rms}}(13)| = 0.051V \] |

\[ I_{\text{rms}}(n) = \frac{V_{\text{LPrms}}(n)}{Z_{\text{parallel}}(n)} \] (2.21)

| \[ |I_{\text{rms}}(1)| = 8.571A \] |
| \[ |I_{\text{rms}}(3)| = 0.177A \] |
| \[ |I_{\text{rms}}(5)| = 0.06A \] |

\[ I_{\text{ILrms}} = \sum_{n=1}^{13} I_{\text{rms}}(n) \] (2.22)

\[ V_{\text{ILrms}} = \sum_{n=1}^{13} V_{\text{LPrms}}(n) \]
\[ |I_{\text{rms}}| = 8.57\text{A} \]
\[ |V_{\text{rms}}| = 24.781\text{A} \]

\[
THD = \sqrt{\left| I_{\text{rms}(3)^2} \right| + \left| I_{\text{rms}(5)^2} \right| + \left| I_{\text{rms}(7)^2} \right| - \left| I_{\text{rms}(9)^2} \right| + \left| I_{\text{rms}(11)^2} \right| - \left| I_{\text{rms}(13)^2} \right|}
\]

\[
THD = 6.7\%
\]

(2.23)

Once this approach was verified, construction of the LCL resonant tank began. It soon became clear that recalculations were necessary due to non availability of initial capacitances that were calculated. By selecting the available capacitors beforehand and using these values to calculate the gain of the LCL resonant tank it was able to construct a new set of boundary conditions for calculating the series inductor. Exact values of the series and parallel capacitors were not available and they were constructed from series and parallel combinations of 0.1µF high frequency, high voltage capacitors. The series inductor was hand wound around a high frequency magnetic core. The step down transformer was also hand wound around a similar high frequency magnetic core. Capacitance and inductance was verified using a 1 kHz RLC multi-meter.

![Figure 2.2 LCC resonant inverter simulation circuit](image)
The evaluation version of PSIM was used to model the LCC topology for variable loads and LCC configurations. The test circuit is shown in Figure 2.2. The following are the snubber circuit calculations.

\[ I_{\text{turnoff}} = 10\text{amp} \]

\[ T_{\text{fall}} = 48\text{ns} \]

\[ V_{dc} = 48\text{volt} \]

\[ T_{\text{deadgap}} = 800\text{ns} \]

\[ C_{\text{snubber}} \max = \frac{I_{\text{turnoff}} \cdot T_{\text{fall}}}{V_{dc}} \]

\[ R_{\text{snubber}} = \frac{T_{\text{deadgap}}}{4. C_{\text{snubber}}} \]  \hspace{1cm} (2.24)

\[ C_{\text{snubber} \max} = 1 \times 10^{-8} \text{F} \]

\[ C_{\text{snubber}} = 1.9\text{nF} \]

\[ R_{\text{snubber}} = 105.263\Omega \]

The ratio of \( L_s \) (series inductance) to the load resistance for example, is an important design parameter. Below are the two most important equations for LCC resonant tank design.

\[ C_w = \frac{C_s}{C_p} \]  \hspace{1cm} (2.25)

\[ Q_i = \frac{\omega_s \cdot L_s}{R_{\text{load}}} \]  \hspace{1cm} (2.26)
Due to the limitations of available components and measurement equipments, the capacitors and inductors in the LCC tank were subject to some drift from the ideal. If it is possible to construct the exact components needed then it would be possible to achieve a more ideal tank gain. Another method of optimizing the circuit is to adjust the switching frequency slightly to vary the ratio of the resonant frequency to the switching frequency. In order to vary the values of the resonant tank, recalculation is required for the above method. A limitation of winding the inductor and the transformer was realized in the 1 kHz test frequency output by the multi-meter. Ideally, a more precise method of measuring the inductance and the transformer inductance would have aided in the implementation of the LCC tank and could be utilized to optimize the design.

2.2.3 Comparison Results of LCL and LCC Resonant Inverters

Simulation was carried out for various load conditions as listed in Table 2.1 for both LCL and LCC resonant configurations. The PSIM simulink model was used to simulate the LCL and LCC resonant configurations.

Table 2.1 Summary of load values

<table>
<thead>
<tr>
<th>Load</th>
<th>Resistance (Ω)</th>
<th>Inductance (µH)</th>
<th>Capacitance (µF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full Load</td>
<td>3.30</td>
<td>5.80</td>
<td>-</td>
</tr>
<tr>
<td>50% Load</td>
<td>6.50</td>
<td>8.90</td>
<td>-</td>
</tr>
<tr>
<td>11% Load</td>
<td>30.00</td>
<td>41.40</td>
<td>-</td>
</tr>
<tr>
<td>Inductive load</td>
<td>3.30</td>
<td>18.40</td>
<td>-</td>
</tr>
<tr>
<td>Capacitive load</td>
<td>3.30</td>
<td>5.80</td>
<td>3.00</td>
</tr>
</tbody>
</table>
Figure 2.3 Device switching loss bar chart

Figure 2.3 shows bar chart comparison of switch loss and diode loss of both converters. The bar chart clearly shows that switching loss, for turning on and off of the MOSFET switches, in the LCL Resonant converter is minimum as compared to the LCC Resonant converter.

Figure 2.4 Load versus % efficiency plot
From Figure 2.4, very little difference is observed between the efficiency of a LCC resonant inverter and a LCL resonant inverter. Though the difference in efficiency is not appreciable, but this may lead to a significant power loss in inverters of large power ratings.

Hence, it was concluded that LCL resonant inverter has better efficiency and less switching power losses. For this reason LCL has been selected for further research in this thesis by investigating its performance with fuzzy and PI controller. Considerable efforts have been devoted in designing a microcontroller circuit to mimic existing fixed frequency phase shift circuits.

Having decided an LCL resonant inverter topology, the primary efforts will be towards optimizing the LCL parameters for the highest efficiency.

2.2.4 Simulation Results for LCL Resonant Inverter

The simulation study on LCL topology for varied loads was carried out. All components are assumed to be ideal because of the insignificance in their effect in the overall performance of the system. However the influence of the leakage inductance of a transformer is significant and therefore it cannot be obviated. For this reason, in simulation, measured leakage inductance of the transformer is incorporated as the series inductor (Ls). The resistors had an intrinsic inductance which has been added in L component in its load. Simulation was conducted for different load configurations. Below are some of the simulation results.
Figure 2.5 $V_{AB}$ and $I_{L_s}$ at full load resistive

Figure 2.5 shows voltage across the terminal A and B and current through the series inductor ($L_s$). The voltage is square wave because of the harmonics produced in the input side of the transformer. These harmonics were eliminated by connecting inductor $L_p$ across secondary of the transformer. So the output voltage becomes sinusoidal. This has been well illustrated from the Figure 2.6.
Figure 2.6  $V_{out}$ and $I_{out}$ at full load resistive

Figure 2.6 shows output voltage across the load and current through load. Shape of the voltage and current wave forms are sinusoidal.
Figure 2.7 $V_{AB}$ and $I_{ls}$ at 11% resistive load

In Figure 2.7, since capacitive reactance due to inter-winding capacitance of resistive load is greater than the inductive reactance of the resonant circuit, capacitive behavior is obtained.
Figure 2.8  \( V_{\text{out}} \) and \( I_{\text{out}} \) at 11% resistive load

Figure 2.8 shows output voltage across the 11% load and current through 11% load. Shape of the voltage and current wave forms are sinusoidal.

Table 2.2 shows summary of RMS value of current and voltage from simulated results at different loading conditions.
Table 2.2 RMS current and voltage from simulation results

<table>
<thead>
<tr>
<th>Load</th>
<th>$I_{LS}$ ($A_{RMS}$)</th>
<th>$I_{out}$ ($A_{RMS}$)</th>
<th>$V_{out}$ ($V_{RMS}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full Load Resistive</td>
<td>12.73</td>
<td>9.14905</td>
<td>30.41</td>
</tr>
<tr>
<td>11% Load Resistive</td>
<td>0.28</td>
<td>1.414</td>
<td>30.41</td>
</tr>
<tr>
<td>Full load Inductive</td>
<td>9.89</td>
<td>7.07</td>
<td>28.28</td>
</tr>
<tr>
<td>Full load Capacitive</td>
<td>19.80</td>
<td>13.5684</td>
<td>53.03</td>
</tr>
</tbody>
</table>

From the above table it is concluded that the RMS value of output voltage is 30.41V and is same at both full load and 11% of load. But in case of inductive and capacitive loads, output voltage was decreased and increased respectively.

2.2.5 Experimental Results

The experimental setup for Figure 2.1 has been fabricated and presented in Figure 2.9. Testing was done on the fabricated setup and oscilloscope plots obtained for variety of actual loads are presented in Figures 2.12 to 2.15. The used scaling factor on the high frequency current probe was not gave consistent results and as such, digital multi-meters were used to measure the RMS current through both the load and the series inductor. Digital multi-meters were also used to measure the RMS output voltage, the average DC input current and the DC Bus voltage. Inductance and capacitance were measured and verified using the 1 kHz RLC probe in ELWB303.

In this work the applicability of the ATMEL 89C51 is investigated as the controller for the LCL resonant converter. The time sharing feature of the ATMEL 89C51 offers ample possibility for its use in the designed LCL RC this has a resonance frequency of 20 KHz.
Figure 2.9 Proto-type model for LCL resonant inverter

A 20 KHz, 250VA, 28V prototype, shown in Figure 2.9, is built to verify the proposed LCL resonant converter. The $L_s$ is chosen to be 0.1754mH, $L_p$ is chosen to be 0.4 $\mu$H, and the resonant capacitor (C) is chosen to be 0.4$\mu$F. The transformer turns ratio is 38:28. The primary and secondary-side switches are selected to be IRF 540. The transformer core is chosen to be EER4242.
Figure 2.10 Fabricated microcontroller unit (89C51)

Figure 2.10 shows fabricated microcontroller unit (89C51). It is a low power, high performance CMOS 8-bit microcomputer with 4K bytes of flash programmable and erasable read only memory (PEROM). The device is manufactured using Atmel’s high-density nonvolatile memory technology and is compatible with the industry-standard MCS-51 instruction set and pin out. The on-chip flash allows the program memory to be reprogrammed in-system or by a conventional nonvolatile memory programmer. By combining a versatile 8-bit CPU with flash on a monolithic chip, the Atmel AT89C51 is a powerful microcomputer which provides a highly-flexible and cost-effective solution to many embedded control applications.
2.2.5.1 Programming algorithm

Figure 2.11 shows square pulse output for port1. Before programming the AT89C51, the address, data and control signals should be set up according to the flash programming mode table. To program the AT89C51, the following steps are used.

1. Input the desired memory location on the address lines.
2. Input the appropriate data byte on the data lines.
3. Activate the correct combination of control signals.
4. Raise EA/Vpp to 12V for the high-voltage programming mode.
5. Pulse ALE/PROG once to program a byte in the flash array or the lock bits.
The byte-write cycle is self-timed and typically takes no more than 1.5ms. Repeat steps 1 through 5, changing the address and data for the entire array or until the end of the object file is reached. The AT89C51 standard features and program are provided in Appendix 1 and Appendix 2 respectively.

![Figure 2.12 V<sub>out</sub> and I<sub>out</sub> at full load resistive](image)

**Figure 2.12 V<sub>out</sub> and I<sub>out</sub> at full load resistive**

Figure 2.12 shows output voltage is 24.57 V (RMS), output current is 7.6A (RMS) and frequency is 20.02 kHz at 100% resistive load.
Figure 2.13 $V_{AB}$ and $I_{Ls}$ at 50% resistive load

Figure 2.13 shows voltage across the terminals A and B and current through the series inductor ($L_s$) with the frequency is 20.04 kHz at 50% resistive load.

Figure 2.14 Full load voltage spectrum

Figure 2.14 shows full load voltage spectrum for FFT sample rate is 400ksa/s and sampling period is 810µs.
Figure 2.15 15 - 50% Load current spectrum

Figure 2.15 shows 50% load current spectrum for FFT sample rate is 400ksa/s and sampling period is 52µs.

Hence the sampling frequency is 19.23 kHz. From the experimental results the values are tabulated and presented below.

Table 2.3 Summary of experimental results and percentage efficiency

<table>
<thead>
<tr>
<th>Load</th>
<th>$I_{DC}$ (Amps)</th>
<th>$V_{DC}$ (V)</th>
<th>$I_{out}$ (A&lt;sub&gt;RMS&lt;/sub&gt;)</th>
<th>$V_{out}$ (V&lt;sub&gt;RMS&lt;/sub&gt;)</th>
<th>p.f</th>
<th>% Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full Load Resistive</td>
<td>4.36</td>
<td>44.78</td>
<td>7.60</td>
<td>24.57</td>
<td>Unity</td>
<td>94.86</td>
</tr>
<tr>
<td>50% Load Resistive</td>
<td>2.04</td>
<td>48.00</td>
<td>3.81</td>
<td>24.56</td>
<td>Unity</td>
<td>90.24</td>
</tr>
<tr>
<td>11% Load Resistive</td>
<td>0.51</td>
<td>48.10</td>
<td>1.52</td>
<td>24.50</td>
<td>Unity</td>
<td>78.63</td>
</tr>
<tr>
<td>Full load Inductive</td>
<td>3.20</td>
<td>48.00</td>
<td>6.05</td>
<td>20.21</td>
<td>0.81 (lag)</td>
<td>90.73</td>
</tr>
<tr>
<td>Full load Capacitive</td>
<td>9.30</td>
<td>48.20</td>
<td>6.42</td>
<td>31.59</td>
<td>0.91 (lead)</td>
<td>94.80</td>
</tr>
</tbody>
</table>
The LCL resonant inverter provides better efficiency at 100% load (both resistive and capacitive). For the first case, power factor is unity and final case power factor is 0.91(lead).

At 11% load also better efficiency (78.63%) is obtained. Losses are also taken into account while calculating efficiency. The sample efficiency calculation (for 100% Resistive Load) is given below.

**2.2.5.2 Calculation for efficiency**

\[ V_{DD} = \text{Drain source voltage} = 50V \]
\[ I_D = \text{Drain current} = 10A \]
\[ f_{SW} = \text{Switching frequency} = 20 \text{ KHz} \]
\[ T_{ON} = \text{ON time period} = \frac{T}{2} \]
\[ T_R = \text{Rise time} = 21 \text{ ns} \]
\[ T_{fall} = \text{Fall time} = 48 \text{ ns} \]
\[ V_{ON} = \text{ON state voltage drop} = 2V \]
\[ I_{OFF} = \text{OFF state current} = 250\mu A \]
\[ P_T = \text{Total losses} = P_{ON} + P_{OFF} + P_{SW} \]

where,

\[ P_{ON} = \text{Average Power Lost during conduction in the Switch} \]
\[ P_{OFF} = \text{Average Power Lost in the Switch when it is OFF} \]
\[ P_{SW} = \text{The Switching loss is given by sum of losses during rise time and fall time of the switch.} \]
\[
P_{\text{turnon}} + P_{\text{turnoff}} = \frac{V_{DD} \times I_D \times T_R \times F_{SW}}{6} + \frac{V_{DD} \times I_D \times T_F \times F_{SW}}{6}
\]

\[
= \frac{50 \times 10 \times 21 \times 10^{-9} \times 20 \times 10^3}{6} + \frac{50 \times 10 \times 48 \times 10^{-9} \times 20 \times 10^3}{6}
\]

\[
= 0.035W + 0.08W
\]

\[
= 0.115W
\]

\[
P_{\text{ON}} = \frac{V_{ON} \times I_{ON} \times T_{ON}}{T}
\]

\[
= \frac{2 \times 10 \times 0.5 \times T}{T}
\]

\[
= 10W
\]

\[
P_{\text{OFF}} = \frac{V_{DD} \times I_{OFF} \times T_{OFF}}{T}
\]

\[
= \frac{50 \times 250 \times 10^{-6} \times 0.5T}{T}
\]

\[
= 6.25 \times 10^{-3}W
\]

\[
\therefore P_T = 10 + 6.25 \times 10^{-3} + 0.115
\]

\[
= 10.12125W
\]
Output Power \( P_{out} \) = \( V_{out} \times I_{out} \times \cos \phi \)  
\[ (2.28) \]

\[ = 24.57 \times 7.6 \times 1 \]

\[ = 186.732\text{VA} \]

Efficiency (\( \eta \)) = \frac{Output Power (P_{out})}{Output Power (P_{out}) + Total Losses (P_L)}  
\[ (2.29) \]

\[ = \frac{186.732}{186.732 + 10.12125} \]

\[ = 94.86\% \]

Table 2.4 Total current harmonic distortion for varied resistive loading

<table>
<thead>
<tr>
<th>Load</th>
<th>Fundamental</th>
<th>3(^{rd})</th>
<th>5(^{th})</th>
<th>7(^{th})</th>
<th>9(^{th})</th>
<th>11(^{th})</th>
<th>13(^{th})</th>
<th>THD (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full Load Resistive</td>
<td>8.29</td>
<td>30</td>
<td>25</td>
<td>27</td>
<td>20</td>
<td>17</td>
<td>10</td>
<td>6.7</td>
</tr>
<tr>
<td>50% Load Resistive</td>
<td>11.5</td>
<td>52</td>
<td>39</td>
<td>42</td>
<td>30</td>
<td>26</td>
<td>15</td>
<td>7.6</td>
</tr>
<tr>
<td>11% Load Resistive</td>
<td>13.1</td>
<td>62</td>
<td>51</td>
<td>50</td>
<td>41</td>
<td>36</td>
<td>25</td>
<td>8.6</td>
</tr>
</tbody>
</table>

The 3\(^{rd}\), 5\(^{th}\), 7\(^{th}\), 9\(^{th}\), 11\(^{th}\), and 13\(^{th}\) harmonics are dominant in calculating percentage THD for the LCL Resonant inverter. Hence considering up to 13\(^{th}\) harmonics is sufficient and is presented in Table 2.4. Sample calculation for THD at full load (Resistive) is given below.
The Value of THD for Full Load (Resistive) is 6.7%. It is same value for both the (Analytical and Experimental) cases.

2.3 LCL RESONANT CONVERTER

The LCL Resonant Full Bridge Converters (RFB) is a new, high performance DC-DC converter. High frequency DC-DC resonant converters are widely used in many space and radar power supplies owing to their small size and lightweight. In Converter applications solid-state devices are operated at very high frequency. So the switching losses are more than the conduction losses and it becomes a major cause of poor efficiency of the converter circuit. This leads to the search of a converter that can provide high efficiency, lower component stress, high power, high switching frequency, lightweight as well as low cost and high power operation. In order to keep the switching power losses low and to reduce the problem of EMI, the resonant converter is suggested.

The limitations of two element resonant topologies can be overcome by adding a third reactive element termed as modified series resonant converter (SRC). The SRC has voltage regulation problems in light load conditions; to overcome this problem the modified SRC is presented. The LCL-resonant converter using voltage source type load has nearly load independent output voltage under some operating conditions. The
LCL-Resonant FBC is simulated using PSPICE software. The proposed results are improved power densities in air borne applications.

A series resonant full bridge converter (SRC) modified by adding an inductor in parallel with the transformer primary is presented. This configuration is referred to as an “LCL resonance full bridge converter”. A three element (LCL) resonance full bridge converter capable of driving voltage type load with load independent operation is analyzed.

Figure 2.16 DC-DC converter employing LCL resonant full bridge converter

The resonant tank of this converter consists of three reactive energy storage elements (LCL) as opposed to the conventional resonant converter that has only two elements. The first stage converts a DC voltage to a high frequency AC voltage. The second stage of the converter is to convert the AC power to DC power by suitable high frequency rectifier and filter circuit. Power from the resonant circuit is taken either through a transformer in series with the resonant circuit (or) across the capacitor comprising the resonant circuit as shown in Figure 2.16. In both cases the high frequency feature of the link allows the use of a high frequency transformer to provide voltage transformation and ohmic isolation between the DC source and the load.
In series resonant converter (SRC), the load voltage can be controlled by varying the switching frequency or by varying the phase difference between the two inverters where the switching frequency of each is fixed to the resonant frequency. The phase domain control scheme is suitable for wide variation of load condition because the output voltage is independent of load.

The major advantages of this series link load SRC is that the resonating blocks the DC supply voltage and there is no commutation failure if MOSFET are used as switches. Moreover, since the DC current is absent in the primary side of the transformer, there is no possibility of current balancing. Another advantage of this circuit is that the device currents are proportional to load current.

The filter circuit has some disadvantage. It is a capacitor input filter and the capacitor must carry large ripple current. It may be as much as 48% of the load current. The disadvantage is more severe for large output current with low voltage. Therefore, this circuit is suitable for high voltage low current regulators.

2.3.1 Specifications of the Resonant Converter

- Power output = 133W
- Minimum input voltage = 100V
- Minimum output voltage = 100V
- Maximum load current = 1.33A
- Maximum overload current = 4A
- Inductance ratio ($K_L$) = 1

The high frequency transformer turns ratio is assumed to be unity.
The input RMS voltage to the diode bridge is

\[(D1-D4) (V_{L2}) = \frac{2\sqrt{2} \cdot Vo}{\pi}\]  

(2.30)

The input RMS current at the input of the diode bridge is

\[I_d = \frac{\pi I_o}{2\sqrt{2}}\]  

(2.31)

The load impedance \(Z_L = \frac{100}{1.33} = 75.18 \Omega\)

\[f_o = \frac{1}{2\pi\sqrt{(L_1 + L_2)C}}\]  

(2.32)

From the above equation, the values of \(L_1, L_2\) and \(C\) are

\[L_1 = L_2 = 185 \mu H\]

\[C = 0.052 \mu F\]

From the availability of the capacitors, it is chosen as 0.05\(\mu\)F. The inductance \(L_1\) and \(L_2\) are obtained as 202\(\mu\)H. In the experimental setup, the actual inductance used is 200\(\mu\)H, which is close to the designed value.

### 2.4 SIMULATION OF LCL RESONANT FULL BRIDGE CONVERTER

The simulated circuit of LCL Resonant Full Bridge Converter is shown in Figure 2.17. Power MOSFET’s are used as switches S1, S2, S3 and S4 in the converters for an operating frequency of 50KHz. The anti parallel diodes, D1, D2, D3 and D4 connected across the switches are not need because they have inherent anti-parallel body diodes. The forward current and the reverse voltage ratings of the diode are the same as the current and voltage ratings of the MOSFET. The internal diode is characterized by forward voltage drop and reverse recovery parameters like a discrete diode.
The resistor, inductor, capacitors, the power diodes and the power MOSFETs are represented by their PSPICE Model. MOSFET IRF 330 is selected as the switching device which meets the peak current and voltage requirements.

Figure 2.17 Simulated circuit of LCL resonant full bridge converter

Figure 2.18 Flow chart for embedded controller 89C51
Figure 2.19  Program result for embedded controller output using Keil software

Figure 2.20  Prototype model for embedded based LCL resonant converter
Figure 2.21  Experimental results of gate voltage (X axis represents time in micro seconds and Y axis represents Gate Voltage. The amplitude is 5V. If S1 and S4 conduct 0-20µs then S2 and S3 are in OFF State. If S2 and S3 conduct 20µs - 40µs then S1 and S4 are in OFF State)

Figure 2.22  Experimental result obtained from the model for a 133W, 50 KHZ DC-DC LCL Resonant Converter output voltage = 127.5 V (L_1= L_2=185µH, C = 0.052 µF, R= 20Ω, C_o = 1000 µF, Load L=10µH, E=10V, Input Voltage = 100 V)
Table 2.5 Comparison of PSPICE simulation, theoretical and experimental results obtained from the model for a 133W, 50 KHz DC-DC LCL resonant converter  
(L1 = L2=185μH, C = 0.052 μF, C0 = 1000 μF, Load L=10μH, E=10V, Input Voltage = 100 V)

<table>
<thead>
<tr>
<th>Load Resistance (Ω)</th>
<th>Simulation Result (Volts)</th>
<th>Theoretical Results (Volts)</th>
<th>Experimental Results (Volts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>127.9</td>
<td>125.0</td>
<td>127.0</td>
</tr>
<tr>
<td>20</td>
<td>128.5</td>
<td>125.0</td>
<td>127.5</td>
</tr>
<tr>
<td>30</td>
<td>128.7</td>
<td>125.0</td>
<td>128.1</td>
</tr>
<tr>
<td>50</td>
<td>129.0</td>
<td>125.0</td>
<td>128.7</td>
</tr>
<tr>
<td>100</td>
<td>129.2</td>
<td>125.0</td>
<td>129.0</td>
</tr>
<tr>
<td>200</td>
<td>129.2</td>
<td>125.0</td>
<td>129.0</td>
</tr>
<tr>
<td>300</td>
<td>129.2</td>
<td>125.0</td>
<td>129.0</td>
</tr>
<tr>
<td>400-1K</td>
<td>129.2</td>
<td>125.0</td>
<td>129.0</td>
</tr>
</tbody>
</table>

From Table 2.5, it is known that the hardware result for open loop LCL resonant converter varies from 127 V to 129V. But theoretical result was not go beyond 125V.

Table 2.6 Comparison of theoretical and PSPICE simulation results obtained from the model for a 133W, 50 KHz DC-DC LCC resonant converter operating in discontinues current mode  
(L1 =10.53μH, C1 = 0.0087, C2=0.033 μF, C0 = 1000 μF, Load L=10μH, E=10V, Input Voltage = 100 V)

<table>
<thead>
<tr>
<th>Load Resistance (Ohm)</th>
<th>Simulation Result (Volts)</th>
<th>Theoretical Results (Volts)</th>
<th>Experimental Results (Volts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>135.4</td>
<td>125.0</td>
<td>137.0</td>
</tr>
<tr>
<td>20</td>
<td>134.0</td>
<td>125.0</td>
<td>136.5</td>
</tr>
<tr>
<td>30</td>
<td>132.7</td>
<td>125.0</td>
<td>134.1</td>
</tr>
<tr>
<td>50</td>
<td>130.0</td>
<td>125.0</td>
<td>133.0</td>
</tr>
<tr>
<td>100</td>
<td>129.2</td>
<td>125.0</td>
<td>131.0</td>
</tr>
<tr>
<td>200</td>
<td>129.2</td>
<td>125.0</td>
<td>131.0</td>
</tr>
<tr>
<td>300</td>
<td>129.2</td>
<td>125.0</td>
<td>131.0</td>
</tr>
<tr>
<td>400-1K</td>
<td>129.2</td>
<td>125.0</td>
<td>131.0</td>
</tr>
</tbody>
</table>
From Table 2.6, it is known that the hardware result for open loop LCC resonant converter varies from 131 V to 137 V. But theoretical result was not go beyond 125V.

![Figure 2.23 Efficiency plot for both LCL and LCC resonant converters](image)

**Figure 2.23 Efficiency plot for both LCL and LCC resonant converters**

### 2.5 DISCUSSION

The LCL topology was taken as a starting point for reasons of its simplicity, stability of the design and previously explored design curves. Our challenge was of two-fold: (1) design a microcontroller solution to generate gating signals to the MOSFET switches and (2) choose and optimize an LCL configuration from the design parameters. There were a number of obstacles to overcome during the design of both challenges.

Choosing the correct microcontroller was the first hurdle to overcome. Due to the high frequency of the output signals it was not immediately clear whether standard output pins on the microcontroller IC
would be fast enough. Eventually, test algorithms were created to determine whether the microcontroller outputs would be fast enough. Initially, it was thought that only the PWM output pins would be fast enough but this proved to be untrue and standard output pins were used to output the gating signals. The 89C51 microcontroller is used for this experimental setup.

The second challenge and the more pertinent, was choosing an algorithm for gating pulse generation. A variety of approaches were explored and finally a solution was reached using a control signal at double the output gating frequency. The rising edge of the control waveform is used to trigger the gating pulses to one arm of the H-Bridge while the falling edge is used to trigger the gating pulses to the other arm of the bridge. By varying the elapsed time between the rising and falling edge pulse width modulation was achieved. Due to time constraints a method to vary the pulse width modulation externally (ie. an adjustable reference voltage via a potentiometer) was not completed. Further work on this design would begin by first incorporating an external control leading naturally into a closed loop design.

From the analysis of the above results (Figure 2.23), it can be concluded that LCL resonant converter is better in performance than LCC resonant converter. The output voltage for LCL resonant converter is close to the theoretical value and the deviation is less. Similarly for LCL resonant converter efficiency is higher than LCC resonant converter. To have constant output voltage irrespective of the load conditions, closed loop control techniques may be employed.

DC gain characteristics and operating region of LCL and LCC resonant converters were analyzed and described in APPENDIX 6.
2.6 SUMMARY

In this chapter, a detailed analysis was done for LCL and LCC resonant converter using simulation supported by hardware. Based on the results obtained, this research work confined its scope with LCL rather than LCC. Optimal parameter selection was briefed and results for various load conditions were tabulated to project the superiority of the LCL over LCC. Total Harmonic Distortion (THD) was also measured for various load conditions for LCL inverter, which supports its high operating efficiency and the ability to handle varying loads.