

HIGH POWER, HIGH EFFICIENCY, LOW COST DC/DC CONVERTERS FOR
LASER TEST EQUIPMENT AND RESIDENTIAL FUEL CELL APPLICATIONS

by

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ABSTRACT

In this work two low cost, high efficiency, high power DC/DC converters are developed. The first converter is targeted for industrial laser applications. The converter is designed for a 400 volt input voltage and a 0-36V output voltage and 0-40A output with a maximum power output of 1500 watts at a cost less than \$0.30 / watt. To achieve a high efficiency and low cost at this power level a zero-voltage switched full bridge converter is used. This technology increases the efficiency of the converter past 90% while reducing the size of the components. The converter was built and tested and achieved a 91.5% efficiency at full load. The total cost was \$0.28 / watt. This converter met all the design goals while exceeding the cost goals.

The second converter is targeted for residential fuel cell applications. This converter utilizes the technology developed for the industrial converter. This residential converter is designed for an input of 26-42 volts at 190 amps and an output of 400 volts and 12 amps at a power level of 5000 watts while maintaining a \$40/kilowatt cost goal. To achieve the low cost and high efficiency design goals the converter uses several technologies in its construction. Like the converter for industrial applications this converter utilizes zero voltage switching full bridge converter. To compensate for the high input current a unique multiphase design was designed for the application. A unique parallel input / series output topology and three interleaved converters split the input current to increase the efficiency of the converter. This unique topology increases the switching frequency on the secondary side which reduces the size of the passive components, reducing cost. The converter was built and tested at a light load to verify its operation versus the theory. An estimated 96% efficiency at full load is possible using this topology. The total cost was \$39 / kilowatt.

INTRODUCTION

DC/DC Converter for Industry

To stay competitive in a volatile telecommunications market, a local company, ILX Lightwave (ILX), specializing in laser test equipment, in conjunction with the Montana Board of Research and Commercialization Technology (MBCRT) asked the university to work with their staff and build a power supply. This supply is intended to replace a commercial power supply in many of their products. Presently the company is using an off-the-shelf power supply a 1500W 36V Cosel PBA1500F-36. At the time of publication it retails for \$880.00 or \$0.59 / watt at an online retailer [1]. To stay competitive the company marketing department requires a new power supply that has a cost of \$0.30/watt in quantities in the hundreds, \$0.29 less than their current supply. In addition to cost, the company requires several other specifications incorporated into the supply to stay competitive. Figure 1 shows current ILX products that will contain the new supply.



a) Industrial Rack Application



b) Benchtop Application

Fig. 1: Typical ILX products [2].

First, the supply must be zero voltage switched, this technology increases efficiency above 90%. This specification is important to marketing their products because several of their products use up to eight of these supplies. In a customer's laboratory environment, housing several of these racks, the radiated heat can be expensive for the customer to vent. A more efficient supply that is cooler is more desirable for the customer because they can spend less money on the cooling of their lab. Additionally, a more efficient supply means that the fans can run slower or less often making the customer's lab environment quieter.

Another required specification is that the supply has a more generalized output than their current supply. The requirements are that the new supply has a 0-36V output in addition to a 0-40A constant current mode. The company has two product categories that require a constant current mode. The first product requires a low noise constant current while the other product can output a more noisy current. Presently the Cosel power supply is used in both products. The Cosel supply only maintains a constant current at its maximum operating current of 40A. To operate at a smaller constant current value, the company uses external current control circuitry. Incorporating this feature into the power supply eliminates the external current control circuitry and opens up a new product line of low cost more noisy constant current supply that do not require the external control circuitry.

Finally, building an in-house power supply the company can modify the design at any time to work with future products that may require a slightly different output voltage

or current. Once the footprint for an in-house supply is designed the company can easily and cheaply modify the output versus buying a new model off-the-shelf supply.

Table 1 outlines the exact specifications of the supply needed by the company. The table outlines several details of the supply including input voltage, output voltage and currents, output power, efficiency, noise requirements, operating modes, safety requirements, operating temperature, as well as the final requirement of the supply, the size. The new supply must fit in the same form factor as the old supply. This means the supply must have a maximum size of 5.9"x2.4"x9.4".

Table 1: ZVS specifications.

	ZVS-1500	
INPUT		Notes
Voltage:	88-132, or 198 - 264 VAC	Universal input
Current:		
100VAC:		
200VAC:		
Frequency:	47-63 Hz	
Efficiency:	>90% @ 120 V >92% @ 220 V	
Power Factor:	0.93	
Inrush Current:		
100VAC:	20/40A	Primary/secondary inrush current
200VAC:	40/40A	
Leakage Current:		According to IEC60950-1
100VAC:	0.45A	
200VAC:	0.70A	
Control Voltage:	0 to 5V	Referenced to "-" output terminal
PFC power rating	1600W	
PFC Output Voltage	400VDC	
OUTPUT		
Voltage:	36 VDC	
Current:	0 to 40A	
Power:	1500 W	
Line Regulation:	0.50%	% change in output voltage to a change in input voltage
Load Regulation:	0.50%	%change in output voltage per amp of output load
Ripple:	0.50%	% full scale voltage p-p
Ripple Noise:	0.60%	% full scale voltage p-p
Temperature Regulation:	1%/°C	% full scale voltage
Drift:	0.40%	% full scale voltage, change in DC output over an 8 hour period at 25°C
Start-up Time:	400 to 500 ms	
Hold-up Time:	≤20 ms	

Table 1 Continued: ZVS specifications

Voltage Adjustment:	0 to 36V	See INPUT – Control Voltage	
Output Voltage Adjustment Range	30.0 to 40.0	Adjustment Range valid when not utilizing remote voltage set point control	
OPERATING MODES			
Constant Voltage	Series Connection		
Constant Current	Parallel Connection		
PROTECTION			
Overcurrent Protection:	105%	Of rated current	
Overvoltage Protection:	V _o + 6 V		
Operating Indication:	Green LED		
Remote on/off:	TTL level – 0=ON, 1=OFF	Referenced to “-” output terminal	
ISOLATION			
Input-Output:	3000VAC	1 minute, cut-off current = 25 mA	
Input-FG	2000VAC	1 minute, cut-off current = 100 mA	
Output RC Aux-FG	500VAC	1 minute, cut-off current = 100 mA	
Output RC Aux	500VAC	1 minute, cut-off current = 100 mA	
ENVIRONMENT			
Operating Temperature:	10° - 40°C		
Humidity:	10% to 90%	RH, non-condensing	
Storage Temperature:	-20°C to 70°C		
Vibration:	10 to 55 Hz	19.6m/s ² 60 min each along X-Y-Z axis	
Size:	150mm x 61mm x 240mm	W x H x D Maximum height 63.5mm (2.5")	
	5.9" x 2.4" x 9.4"		
Power Density:	5W/in ³		
Weight	<2.5 kg		
Connectors:			
Input:			
Output:			
Voltage Control:			
Safety:	EN60825-1 / EN60950 / EN61010 / EN61326 / 21CFR 1040.10		
EMC:	98/336/EEC / Low Voltage Directive		
Regulation:	CE		
	RoHs		

The design of this converter must meet all specifications outlined above while still maintaining a low cost, small size, and be easily reproducible in a manufacturing environment. A design must be chosen that incorporates all these factors.

DC/DC Converter for Residential Fuel Cell Applications

Distributed power generation has become a popular topic in recent years. Distributed generation consists of several small power generators spread over a larger power distribution system. This allows consumers to generate their own power during

high local demand and sell unused power back to the power company during low demand. Distributed generation requires small efficient generators that can easily fit in a home or business. Distributed generation can be marketed as a clean solution to consumer power because the most dominant technologies are solar cells, wind generators and fuel cells. All of these technologies can be marketed as clean energy.

Solid-Oxide Fuel Cells (SOFCs) are a promising technology for use in distributed generation systems. SOFCs are energy conversion devices that convert chemical energy from fuels such as natural gas and propane into electrical energy. The major advantages of fuel cells are that they are small, highly efficient, have low emissions, are quiet and have excellent fuel flexibility which is important in rural areas where certain fuels are not always available.

The typical structure of a residential SOFC generation system is shown in Figure 2. In this system a 22-41V, 5kWatt SOFC stack, according to the specifications of the 2003 US Department of Energy (DOE) and IEEE Future Energy Challenge [3] is used. Table 2 outlines the specifications in more detail. In this system a DC/DC converter is used to boost the low voltage from the SOFC to a working voltage of 400V as well as provide galvanic isolation. This voltage is required by the DC/AC second stage. This DC/AC inverter provides the residence or business with a split 120/240V 60Hz supply.

Table 2: DOE/IEEE challenge specifications [4].

Design Item	Minimum Target Requirement	
Manufacturing Cost	Less than US \$40/kW in high volume production	
Output Power	Nominal	5kW continuous at DPF 0.7
	Overload	10kW overload for 1 minute at DPF 0.7 5kW from fuel cell and 5kW from battery
Energy Source	Primary source (SOFC)	29V nominal, 22-41Vdc 275A max. from 5kW fuel cell
	Battery	48V nominal, +10%/-20% 500Wh
Output Voltage	Split single-phase 120V/240V, 60Hz	
	Voltage Regulation	+/- 6%
	Frequency Regulation	+/-0.1Hz
	THD	Less than 5%
Acoustic noise	Less than 50dBA at 1.5m distance	
Overall efficiency	Higher than 90%	
Protection	Over current, over voltage, short circuit, over temperature, and under voltage	

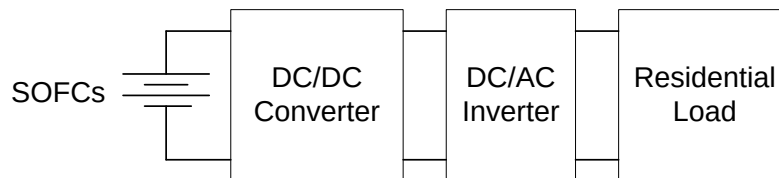


Fig. 2: A typical structure of a SOFC residential system [5].

This thesis will concentrate on the DC/DC converter subsystem. According to the specifications of the 2003DOE/IEEE Future Energy Challenge, the DC/DC converter must have an efficiency over 90% while maintaining a cost less than \$40 / kilowatt or \$200 per unit in large quantities. These requirements along with the input voltage and

power level make the DC/DC converter design challenging. First, the cost and efficiency requirements are mutually exclusive. To maximize the efficiency, expensive components are usually used. A compromise must be made by careful design and a smart selection of components. Next, the low input voltage and extremely high input current create the toughest challenge of the design. The high input current requires large high-current semiconductors to properly operate using traditional methods. Large semiconductors can inflate the cost of the converter past the \$40/kW goal. Additionally, these high currents create higher switching and conduction losses in the semiconductors, reducing the efficiency. Finally, the large voltage boost in the converter requires a transformer with a large turn ratio. The turn ratio makes the transformer large and expensive. The large turn ratio also creates large leakage inductance in the transformer. The large leakage inductance in the transformer can lead to stability problems and efficiency loss if soft switching is used in the design. A new DC/DC converter topology must be utilized to manage the input current and reduce the losses and cost.

DC/DC CONVERTER FOR INDUSTRY

Topology

Several factors must be considered when designing a power supply. The most important factor of the design is the general topology of power components. Several topologies are available, Buck, Boost, Buck/Boost, Cuk, Forward, Flyback, Half-Bridge, and Full-Bridge are some of the more popular configurations. Each of these topologies have advantages and disadvantages that must be evaluated to make a educated choice in the topology. In the case of the industrial power supply four specifications define the topology, isolation, soft switching, size, and cost. The most limiting specification is the cost. The cost limits the choices of topology to the most common topologies. There are hundreds of new power supply topologies proposed by people all over the world that are super efficient, but may require exotic components and complicated control strategies. Limiting the design choices to the common topologies allows off-the-shelf controller ICs and common components to be used, drastically reducing the cost. The second most limiting requirement is isolation. The need for isolation eliminates several topologies. Buck, Boost, Buck/Boost, and Cuk, all have no isolation capabilities. The next specification, the size, eliminates the Flyback converter and the Half-Bridge. The Flyback requires a large transformer because the transformer must have an air gap and more windings to double as an inductor. In both topologies the flux in the transformer swings from zero to a positive value requiring a larger transformer to prevent saturating the core. Finally, the soft switching specification eliminates the Flyback converter.

Although soft switching is available for the Flyback it is an exotic topology without a commonly available controller IC. All that remains is the Full-Bridge converter. It is a very popular topology which controller ICs are available from several manufacturers. It utilizes a transformer in its design providing isolation. The flux in the transformer swings from a negative to a positive value, requiring a smaller transformer. It has the ability to be soft switched without any additional components with an off-the-shelf controller.

Beyond satisfying the major specifications the full bridge offers other benefits. It is a very good topology for medium to high power applications at medium to high input voltages. The input current and voltage are split between the switching MOSFETs requiring modest requirements in the ratings of the FETs. On the secondary side the frequency is doubled due to the diode rectification requiring smaller filtering components.

The Full Bridge Converter

Figure 3 shows the full bridge topology. It consists of a DC bus supplying power to the system, the switching MOSFETs configured in an H-Bridge fashion, a transformer with isolation, a diode rectifier, a low pass LC filter, and the load. In this converter switch pairs QA, QD, and QB, QC are alternately switched at the switching frequency. The switching cycle as shown in Figure 4 starts at time $T=0$ with the pair QA and QD switched on. At this point the input voltage V_i is applied to the primary side of the transformer. This voltage is reflected to the secondary side as

$$V_S = \frac{N_S}{N_P} V_P. \quad (1)$$

Diode D1 is now forward biased and D2 is reversed biased. The voltage VL at this point is Vs-Vo. The current following the equation

$$VL = L \frac{dl}{dt} \quad (2)$$

increases linearly. At time $T=1$ the pair QA, QD is turned off. V_p and V_s are now zero volts. The diodes are reversed biased and not conducting. The voltage across the inductor is $0-V_o$. The current in the inductor decreases linearly. At time $T=2$ the pair QB, QC is turned on. V_p becomes $-V_i$ and V_s following equation 1 becomes negative. Diode D2 is now forward biased and D1 reverse biased. The inductor voltage becomes the positive V_s-V_o and its current rises linearly. At time $T=3$ all switches are off again. This state is the same as $T=1$, the inductor current decreases. Finally, at time $T=4$ switch pair QA, QD is turned on starting the switch cycle over.

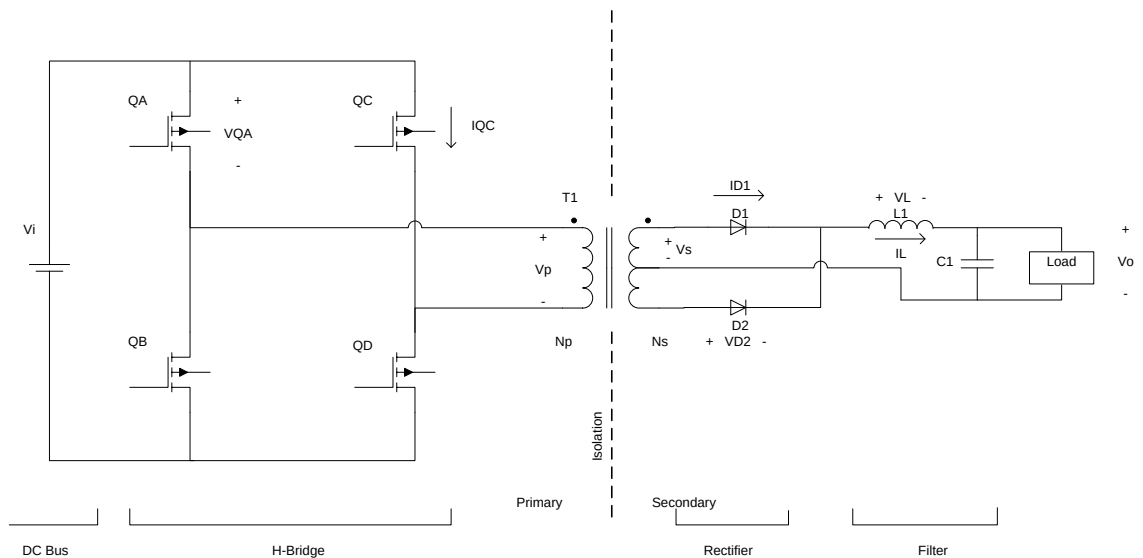


Fig. 3: Basic full-bridge topology.

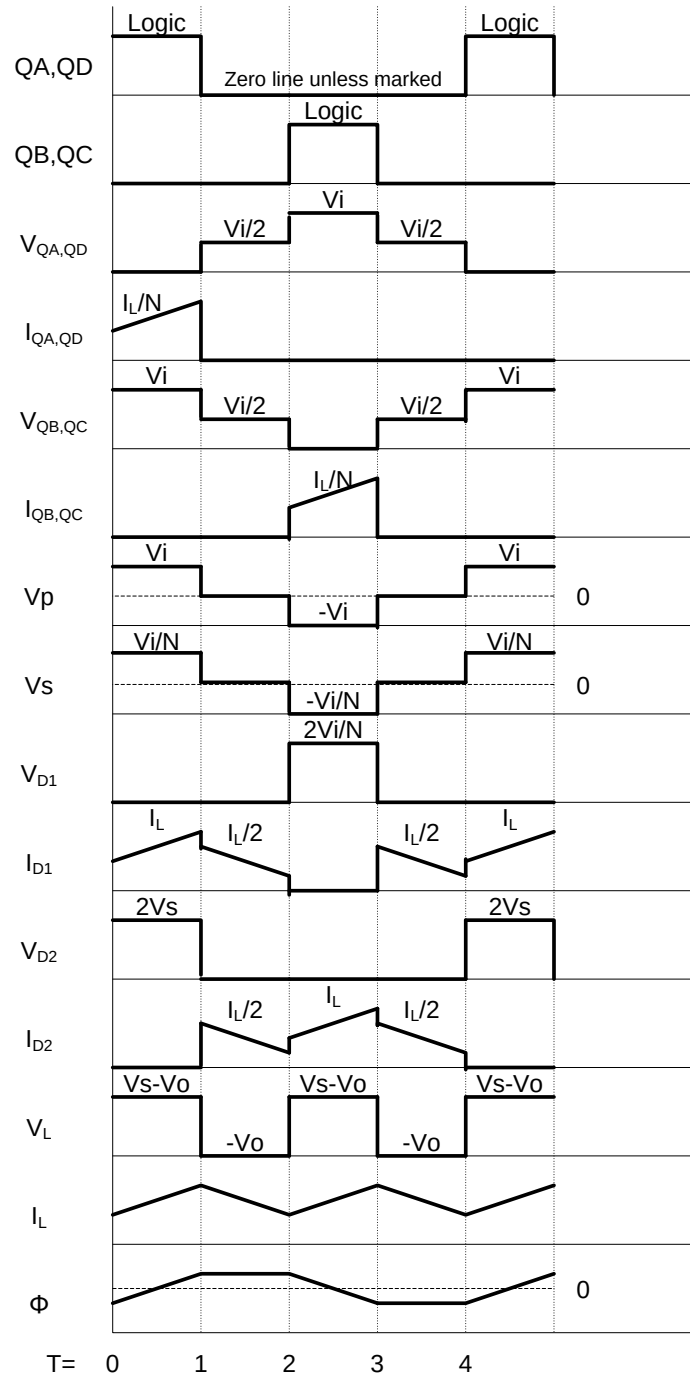


Fig. 4: Full-bridge switching waveforms.

The switching cycle creates a triangular current waveform through the inductor with a DC offset. If this triangular current were to travel through the load it would create

an AC waveform through the load making the converter a DC/AC-like converter. To remedy this, a capacitor is placed in parallel with the load. Since a capacitor looks like a short circuit to AC currents the triangular current passes through the capacitor while the DC component passes through the load.

The governing equation for this converter is the equation relating the input voltage to the output voltage. The derivation can be done by only looking at the secondary side. When any switch pair is on the inductor equation (2) becomes

$$\text{ON: } V_s - V_o = L \frac{I_{min} - I_{max}}{DT}. \quad (3)$$

Where L is the inductance, I_{max} and I_{min} are the maximum and minimum current through the inductor, D is the duty cycle or T_{on}/T defined in Figure 5, and T is the switching period. When the switches are off equation (2) becomes

$$\text{OFF: } -V_o = L \frac{I_{max} - I_{min}}{(1-D)T}. \quad (4)$$

Rearranging equations 3 and 4 yields

$$\frac{DT(V_s - V_o)}{L} = I_{min} - I_{max} \quad (5)$$

$$\frac{V_o(1-D)T}{L} = I_{min} - I_{max}. \quad (6)$$

Combining eqns. 1, 5, and 6 and solving for V_o the input voltage to output voltage equation becomes

$$V_o = V_{in}(D) \frac{N_s}{N_p}. \quad (7)$$

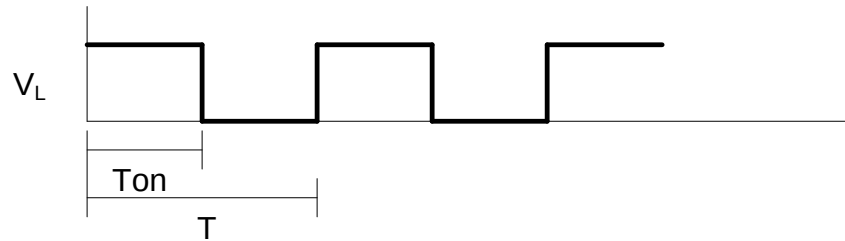


Fig. 5: Definition of duty cycle $D=T_{on}/T$.

Zero Voltage Switched Full Bridge Converter

The zero voltage switched full bridge converter, (ZVS) full bridge, or soft switched full bridge was invented as a way to increase the switching frequency while maintaining a low efficiency. Increasing the switching frequency has advantages that help reduce the cost of the converter. With a higher switching frequency, the magnetics can be smaller and cheaper. Additionally, the filter components can be smaller because they can have a higher cutoff frequency which improves the cost as well. However, as switching frequencies increase the switching loss increases. Switching loss occurs during the ON/OFF transitions in the power MOSFETs. Figure 6 shows the voltage and current waveforms associated with switching loss.

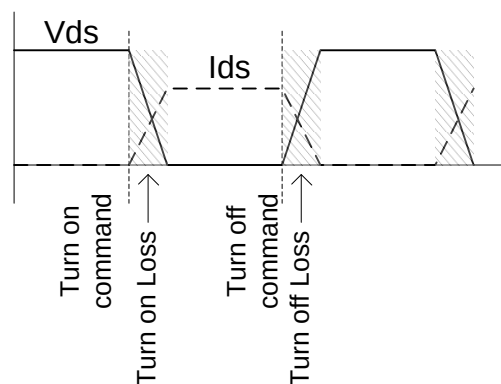


Fig. 6: Switching loss.

When the on or off command is applied to the gate the voltage and current in the switch do not immediately transition to their new values. Due to parasitic capacitances and inductances the voltage and current transition slowly. These slow transitions cause an overlap of voltage and current in the switch causing a power loss in the switch.

Increasing the switching frequency causes these transitions to happen more times, causing higher losses in the switch.

Soft switching aims to reduce the switching loss of the converter. The idea is to utilize resonance and the parasitic components in the converter to eliminate the voltage/current overlap in the switch. The soft switching technique used is the Phase Shifted, Zero Voltage Transition technique or Zero Voltage Switching (ZVS). It uses the parasitic diodes and capacitors in the MOSFETs as well as the leakage inductance of the transformer forming a resonant tank. The circuit diagram showing the power circuit with the parasitics is shown in Figure 7.

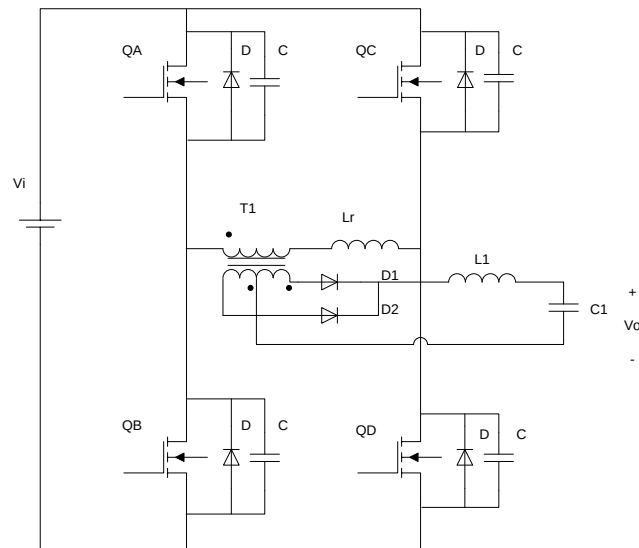


Fig. 7: ZVS circuit diagram with parasitics shown.

Unlike the conventional full bridge design, the switches are not driven in diagonal pairs. A deliberate dead time is introduced in the switching cycle where the switch remains off and is clamped at zero volts by the resonant tank. Figure 8 shows the switching waveforms for the ZVS converter with dead-time. Additionally, the duty cycle of each switch is not controlled like the hard switched converter. To regulate the output voltage a phase shift between the diagonal pairs is added. At the primary terminals this phase shift is the equivalent of a duty cycle change. Figure 8 shows the resultant PWM signal of the diagonal pairs.

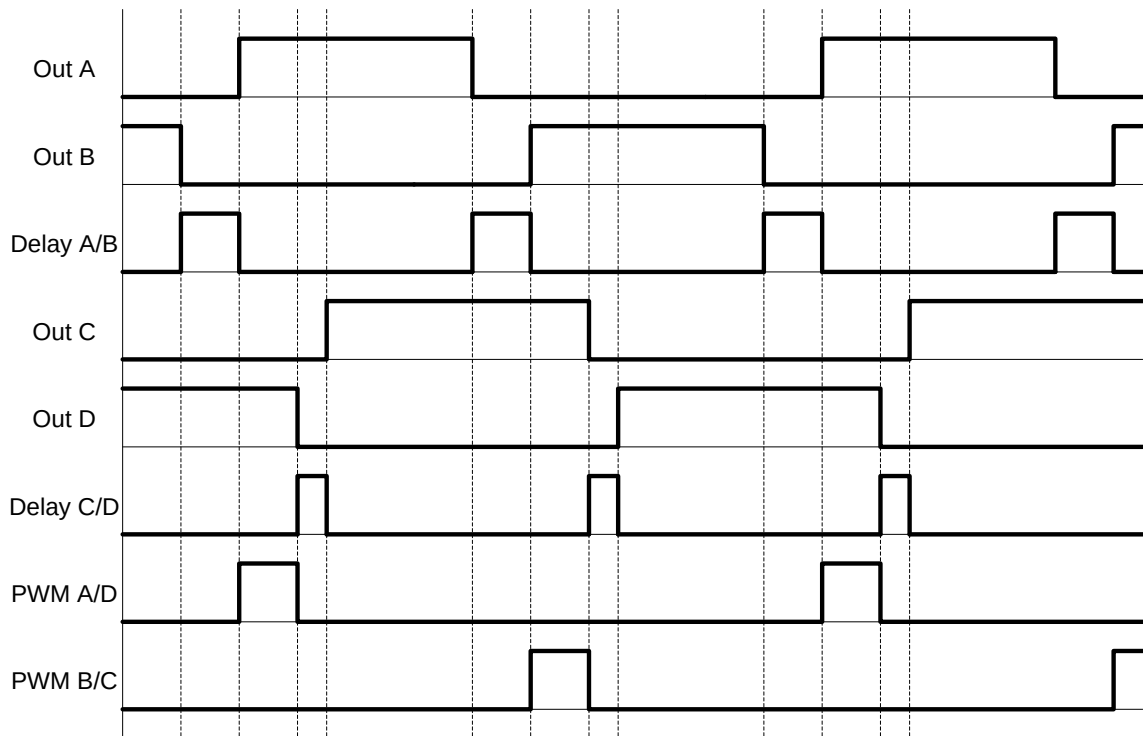


Fig. 8: Control signals with dead-time.

The operation of the phase shifted converter as described in [6] will begin with the initial conditions.

Initial Conditions: $t=t(0)$, Figure 9 a). The initial conditions begin with the conclusion of a power transfer cycle where the transformer has been delivering power to the load. Two diagonal switches were conducting and the initial current into the transformer is $i_p(t(0))$.

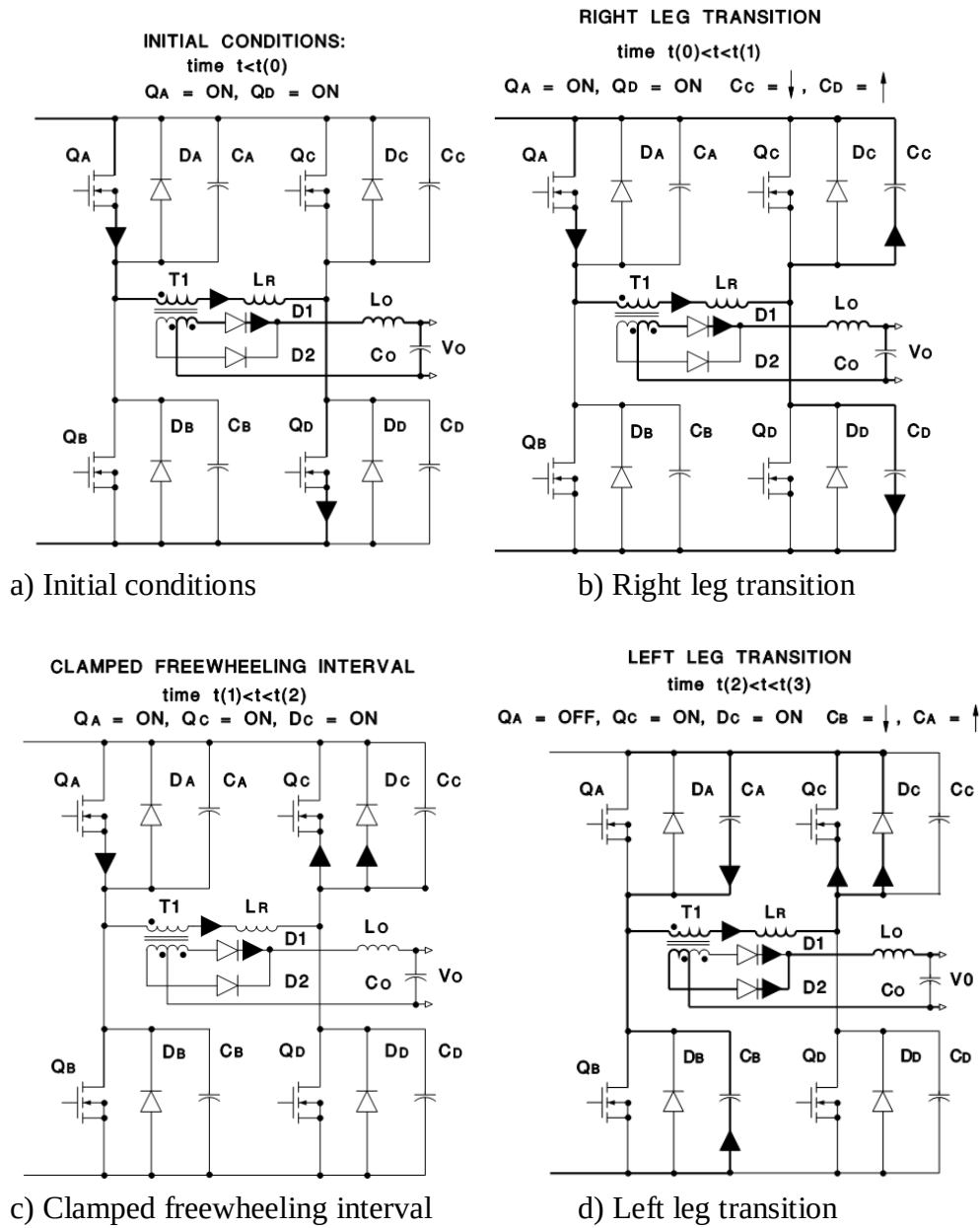
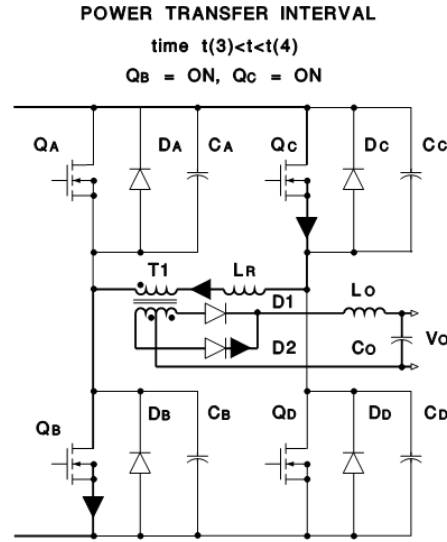


Fig. 9: ZVS operational cycles [6].



e) Power transfer interval

Fig. 9 Continued: ZVS operational cycles [6].

Right Leg Resonant Transition: $t(0) < t < t(1)$, Figure 9 b). The primary current at time $t=t(0)$ is equal to $i_p(t(0))$. This current is being conducted through the diagonal pairs of QA and QD. At time $t(0)$ the gating signal is removed from QD which begins the resonant transition of the right leg. The primary current is maintained through the resonant leakage inductance of the transformer. With QD off the primary current begins to flow through the output capacitance of QD. This charges the capacitor from zero volts to the upper voltage rail. At the same time, the capacitance of QC is discharged from the upper rail to zero. This transition places zero volts from drain to source on transistor QC before the turn-on signal is applied. This allows switch QC to be switched with zero volts across it creating zero voltage switching.

During the right leg transition the primary transformer voltage decreases from V_{in} to zero. During this transition the primary voltage drops below the reflected secondary voltage. At this point the primary is no longer supplying power to the secondary and the

output inductor changes polarity. The energy stored in the output inductor begins to supply the primary voltage until the voltage reaches zero.

When the right leg transition has been completed there is no longer voltage across the transformer primary, the transformer secondary and no power transfer.

Clamped Freewheeling Interval: $t(1) < t < t(2)$, Figure 9 c). At this time the turn-on signal has been applied to QC. Both QA and QC are on. The primary current freewheels through transistor QA, the channel of QC, and the body diode of QC. Although current is flowing in the opposite direction in QC than normal, the current is split between the R_{ds} and the body diode. This point causes the biggest disadvantage to ZVS, the duty cycle loss on the secondary side. During this freewheeling interval the secondary side produces no output which creates a duty cycle loss. Although the duty cycle loss produces no efficiency drops on its own, it does require a larger turn ratio in the transformer to overcome the voltage loss.

Left Leg Transition: $t(2) < t < t(3)$, Figure 9 d). At this point switch QC is fully on and QA is ready to be turned off. When QA is turned off the primary current will continue to flow, but it will flow through the capacitance of QA instead of its channel. The direction of the current causes the drain to source voltage of QA to increase and lowers the source voltage to zero volts. In switch QB the opposite has occurred. QB previously had the full input voltage across it now is lowered to zero volts. The resonant transition has allowed QB to be switched with zero volts across it allowing lossless switching.

Primary current continues to flow and is clamped by the body diode of QB which is still off. The clamping is necessary to facilitate fixed frequency switching. Switching QB on the instant it reaches zero volts will cause variable frequency operation which is not the desired operation. With QB on the transformer has the full V_{in} across it and begins to transfer power.

It should be noted that the left leg transition takes longer to complete due to conduction losses in the switches, transformer windings, and connections resulting in a DC voltage drop. Energy stored in the leakage inductor is no longer clamped to zero volts. This loss along with the losses in the previous transition reduce the primary current below its initial $i_p(t(0))$ causing a longer transition time in the left leg than the right.

Power Transfer Interval: $t(3) < t < t(4)$, Figure 9 e). This interval is basically the same as the standard “hard switched” full bridge converter. Two diagonal pair are on, which apply the full input power to the primary of the transformer. The current rises with a rate determined by V_{in} and the leakage inductance. However, the current begins at a negative value versus zero with the hard switched converter. Power is transferred to the secondary side where the voltage is rectified, filtered, and supplied to the load.

Figure 10 outlines the voltages and currents during the soft switched transitions. Each time interval is labeled on the figure.

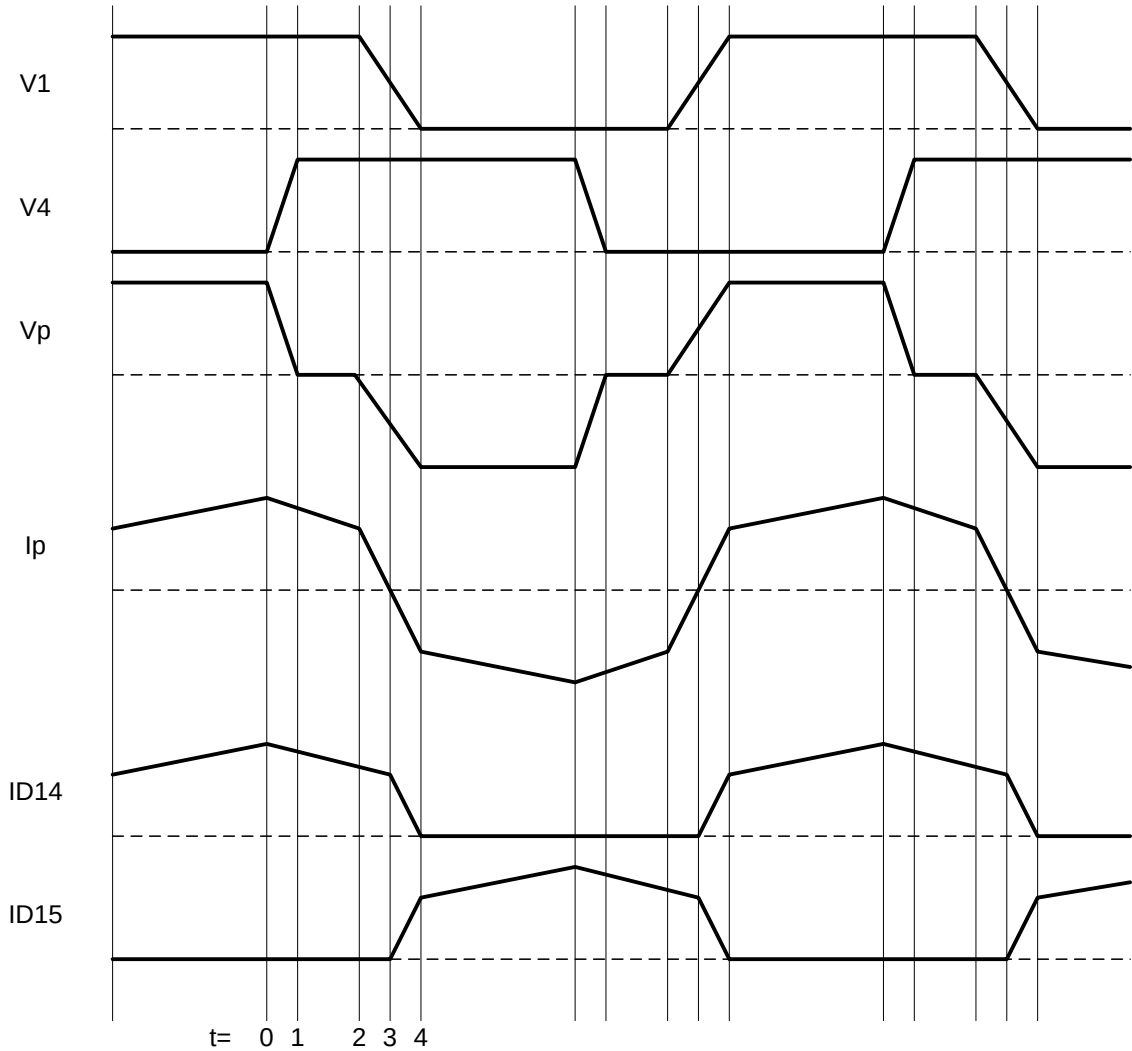


Fig. 10: ZVS voltage and current waveforms.

Design of the Converter

The DC/DC converter for industry has two major components, the power component and the control component. The power component consists of all the high current and voltage components that do the heavy work converting the input voltage to the desired output voltage. The control component consists of low voltage components

that monitor the output voltage and current and regulate the power circuit. The control circuit makes sure that the desired output is within tolerance. The design of the circuit begins with the power circuit and then the control circuit.

The Power Circuit

The design of the DC/DC converter for industry begins with the power circuit following the outline for a soft switched full bridge converter in Figure 7. There are five major components to choose in the power stage, the switching MOSFETs, the flux-runaway capacitor, the transformer, the rectifier diodes, the inductor and the capacitors. Each ratings of each component must be calculated based off the specifications of the converter. Parts must then be chosen that meet the calculated ratings and minimize cost.

The design of the converter begins with the design of the transformer. The design goals of the transformer include primary and secondary voltage and current ratings, turn ratio, and size. These values can be calculated by the converter specifications. From the specifications the minimum input voltage is 385V and the maximum output voltage and current of the converter must be 40V and 40A. The turn ratio is found from the input to output voltage transfer function in equation 7. Filling in the numbers eqn. 7 becomes

$$40 = 385(D) \frac{N_s}{N_p}. \quad (8)$$

To maximize efficiency, the duty cycle (D) can be set to 95% which assumes that due to component tolerances the duty cycle cannot reach 100%. Solving eqn. 8 using D=0.95 N_s/N_p is 0.1053 or 1/9.45. However due to the duty cycle loss from the soft switching the turn ratio must be modified a little. A turn ratio of 0.149 or 1/6.7 is calculated using a

lower duty cycle of 70%. Using the turn ratio and the output current, the peak and RMS input current can be calculated using

$$I_{in} = I_{out} \frac{N_s}{N_p} \quad (9)$$

$$I_{rms} = I_{peak} \sqrt{D} \quad (10)$$

where D is the primary square wave duty cycle. Using eqns. 9 and 10 the input current is found to be 5.97A_{peak} and 4.22A_{rms}. The transformer specifications, input voltage, input current, output voltage, output current, power rating, turn ratio, and height were sent to a transformer manufacturer.

After several quotations from various transformer manufacturers a design by Thomas Magnetics was chosen. In quantities of 100, it costs \$27.10 each. The datasheet of the finished transformer is found in the appendix.

Based the input voltage and current calculated by designing the transformer, the MOSFETs can be chosen. The primary ratings of the FETs include voltage rating, current rating, and cost. Using the current and voltage ratings several transistors can be found. The FET specification that narrows down the choices is the drain to source resistance or R_{ds}. A small R_{ds} reduces the conduction loss of the switches. Transistors with small R_{ds} are typically high current devices. Choosing a FET with a current rating higher than the required 6A will reduce the R_{ds}. However, any high current FET cannot be used because a high current FET costs more money. Careful selection is the key to selecting a FET that has a small R_{ds} and low cost.

After some time was spent looking at MOSFETs, the Infineon SPW20N60C3 was chosen. This transistor has a voltage rating of 650V, a current rating of 20.7A and an R_{ds}

of 0.19Ω . In quantities of 100, it retails for \$3.56 each. A higher voltage rating was chosen to minimize failure due to high voltage spikes during the switching cycle due to parasitic inductances.

The flux-runaway capacitor prevents any DC component from the switching components from reaching the transformer due to possible errors in the switching cycle. A DC component at the transformer will cause the transformer to saturate, heat up, and quit working. A bipolar capacitor with a sufficient current rating and large capacitance must be chosen. The current rating of this capacitor is equal to the primary current in the transformer or 4.22A. The capacitor B32676*6106 by Epcos has a current rating of 12 A and a capacitance of $10\mu\text{F}$. It costs \$7.73 each in large quantity.

Based on the transformer design, the rectifier diodes on the secondary side can be chosen. According to the waveforms in Figure 4, the diode current rating is equal to the output current (40A) and the voltage rating is $2V_s$ or $2*57.5=115\text{V}$. Choosing a diode is much like choosing the MOSFET. A diode must be found that exceeds the voltage and current ratings by at least 10% to minimize failure, has a low cost, has fast switching characteristics, and minimize losses. The losses in a diode come from the forward voltage drop across the junction. When current flows through this voltage drop, the losses are $V_f*I=P_{\text{loss}}$. Minimizing the voltage drop minimizes loss. Of all the common diodes available, a Schottky diode has the smallest voltage drop of any power diode and has good switching characteristics. The 60A, 150V, 0.66V_f, IXYS DSSK 60-015A was chosen. It has a cost of \$6.50 in large quantities.

The next component to choose in the power design is the inductor. The inductor has two specifications, the inductance and the current rating. The inductance is calculated based off the desired current ripple in the output. The specifications call for a 20% current ripple at full load. Equation 2 is used to calculate the inductance. Modifying eqn. 2 using the known values and current ripple the equation becomes

$$V_s - V_o = L \frac{\Delta I}{T_{on}}. \quad (11)$$

Inserting the values and rearranging to solve for L the equation becomes

$$L = \frac{(57.5-36)0.5}{0.2(40)100000} = 13.4\mu H. \quad (12)$$

The current rating for the inductor is simply the rated peak output current of 40A. An inductor by Cooper Bussmann (HC3-3R3-R) was chosen meeting these specifications. Its cost is \$12.68 in large quantities.

The final component in the power stage is the capacitor. The specifications of the capacitor include the voltage capability and the capacitance. The voltage capacity is simply the maximum output voltage of 40V. The capacitance required is based off the desired voltage ripple and spiked current demand of the output. In this case, no calculations were performed to determine the capacitance. The amount of capacitance needed was determined by reverse engineering the model Cosel power supply. This was done to match the Cosel's peak pulsed current capacity. Eight Panasonic ECA1HM471 capacitors with a working voltage of 50V and capacitance of 1000 μ F were chosen to match the Cosel's performance.

The schematic of the power circuit is shown in Figure 10. The schematic includes two other components that are not needed for the basic full bridge, but are needed to meet

CE requirements and generally make the converter perform better. These components are snubbers and EMI capacitors.

Snubbers are small R-C networks that are placed across active switching devices such as the MOSFETs and diodes. Snubbers reduce the voltage spikes that occur in these devices when they are turned off. Voltage spikes in the system create noise problems in the control circuit, reduce the lifetime of the active devices and cause voltage ringing in the system. It is desirable to reduce these spikes whenever possible.

The voltage spikes are created due to the parasitic inductances in the wires connecting the components resonating with the capacitances in the active devices. When the active device is turned off the current previously flowing in the device wants to keep flowing in the same direction due to the parasitic inductances. The large rate of change in the current creates a voltage spike that is seen across the active device. The snubber adds a capacitor and resistor across the active device to absorb the voltage spike.

Although there are theoretical methods to design snubbers [7], it is often easier to empirically find the values needed. In the design snubbers are placed across the primary MOSFETs and the secondary diodes. An alternate method of placing the snubbers across the transformer was used to reduce the number of snubbers and reducing cost.

The EMI capacitors are used to provide current paths to earth ground. Circulating current in the system causes unnecessary oscillations in the system and creates unnecessary noise, both radiated from the system and conducted out the output. Adding paths to earth ground reduces the circulating current and minimizes oscillations.

Since these capacitors are connected to earth ground they must meet certain safety requirements to meet CE specifications. The capacitors must be rated X1/Y2. This means that during a failure they will not conduct any potentially dangerous voltage to the ground which is usually the same potential as the metal case.

The snubbers and EMI capacitors are outlined in the power circuit in Figure 10. Their cost is outlined in the Bill of Materials for the Industrial Converter in the appendix.

The Control Circuit

The control circuit consists of several sections. These sections include the PWM modulator, gate drive circuitry, voltage feedback, current feedback, automatic crossover system, voltage and current command circuitry, current sense circuitry, short circuit protection, remote off circuitry, fan fault circuitry, power good indicator, output good indicator, and negative voltage supply.

PWM Modulator: The heart of the control circuit is the PWM modulator. It takes a voltage input and converts it into four square wave signals to drive the switching MOSFETs. The IC used in this design is the UC3875 by Texas Instruments. It is specifically designed to control a phase shifted ZVS full bridge converter. Several application notes have been published to aid engineers designing a converter using the UC3875 [6,8].

The UC3875 is designed to incorporate the time delay needed for the resonant tank in the ZVS to properly circulate. The time delays for each leg can be calculated to properly achieve ZVS. [8] provides a spreadsheet to calculate the resonant tank transition

times to properly design the leg delay times. All the values needed are added to the spreadsheet and shown in the appendix. The equations in the spreadsheet calculate the delays as 457ns for the right leg C/D and 718 for the left leg A/B. The UC3875 datasheet [9] provides equations to translate the delay times into resistor values.

Voltage Control: The PWM controller can do little without the current and voltage control circuitry to regulate the PWM signals. The current and voltage control look at the output voltage and current and compare it with a reference value. The error amplifier is the main component of this control. It compares the output value and the reference value as well as applies the control math to keep the output stable. To properly design the error amplifier the transfer function of the power circuit must be calculated.

The transfer function is an equation that translates small changes in the control signal into small changes in the output. The voltage transfer function of the full bridge can be calculated based off the simpler buck converter which the full bridge is derived, plus a gain [10]. An analysis method using state-space averaging is used in [11] to derive the transfer function

$$\frac{\hat{V}_o(s)}{\hat{d}(s)} = \frac{R V_s(1+s R_c C)}{R+s(L+R R_c C)+s^2(R+R_c)LC}. \quad (13)$$

Where R is the load resistance, V_s is the secondary voltage, R_c is the ESR of the output capacitor, C is the output capacitance, and L is the output inductance. The bode plot for this system is shown in Figure 11.

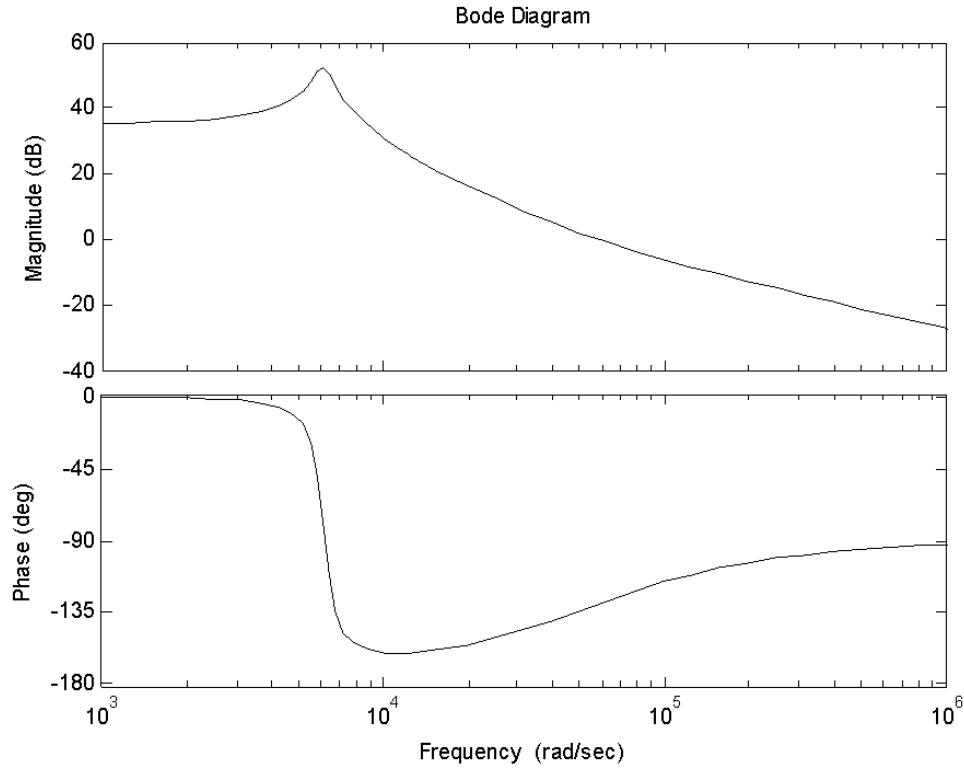


Fig. 11: Bode plot for converter.

Following the procedure outlined in [12] an error amplifier can be designed. The procedure outlines some desired characteristics of the closed loop transfer function. First, the gain at low frequencies should be high to minimize the steady-state error in the output. Next, the crossover frequency, the frequency at which the gain of the closed loop transfer function falls to 0 dB, should be as high as possible but approximately an order of magnitude below the switching frequency to allow the power supply to respond quickly to transients. Finally, the phase margin should be in the 45° - 60° range.

The schematic of the error amplifier is shown in Figure 12. The values calculated for the error amplifier are $R_1=150\Omega$, $R_2=681\Omega$, $C_1=0.18\mu\text{F}$, and $C_2=0.01\mu\text{F}$. The transfer function of this converter is

$$T_c(s) = \frac{1}{R_1 C_2} \frac{s + \omega_z}{s(s + \omega_p)} \quad (14)$$

where

$$\omega_z = \frac{1}{R_2 C_1} \quad (15)$$

$$\omega_p = \frac{C_1 + C_2}{R_2 C_1 C_2}. \quad (16)$$

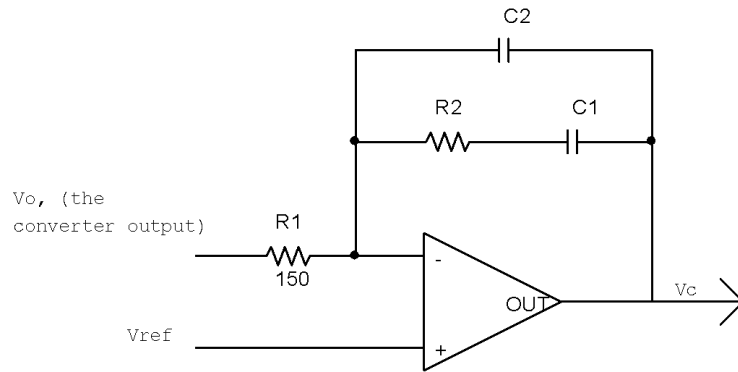


Fig. 12: Schematic of error amplifier.

The bode plot from Matlab is shown in Figure 13 and the Spice simulation of the amplifier is shown in Figure 14. They are similar except for the roll off of the spice simulation at higher frequencies due to the op amp used in the simulation.

As a verification, the step response of the closed loop system was simulated in Matlab. Figure 15 shows the step response, it has an acceptable settling time as well as rise time. For comparison, Figure 16 shows the open loop step response to a unit step. It has a steady state error of about 59 and large oscillations.

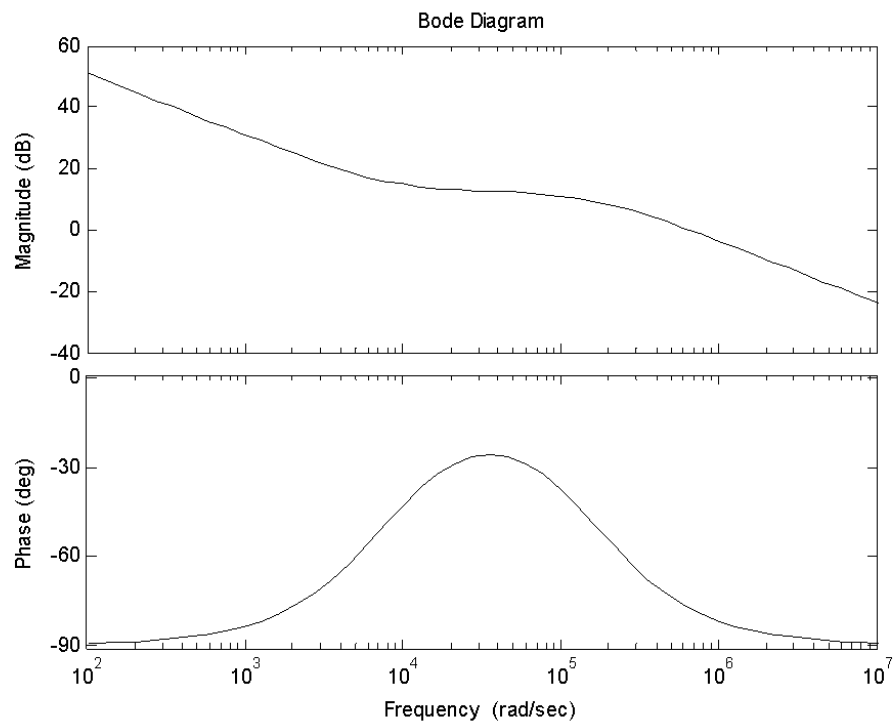


Fig. 13: Matlab calculated error amplifier transfer function.

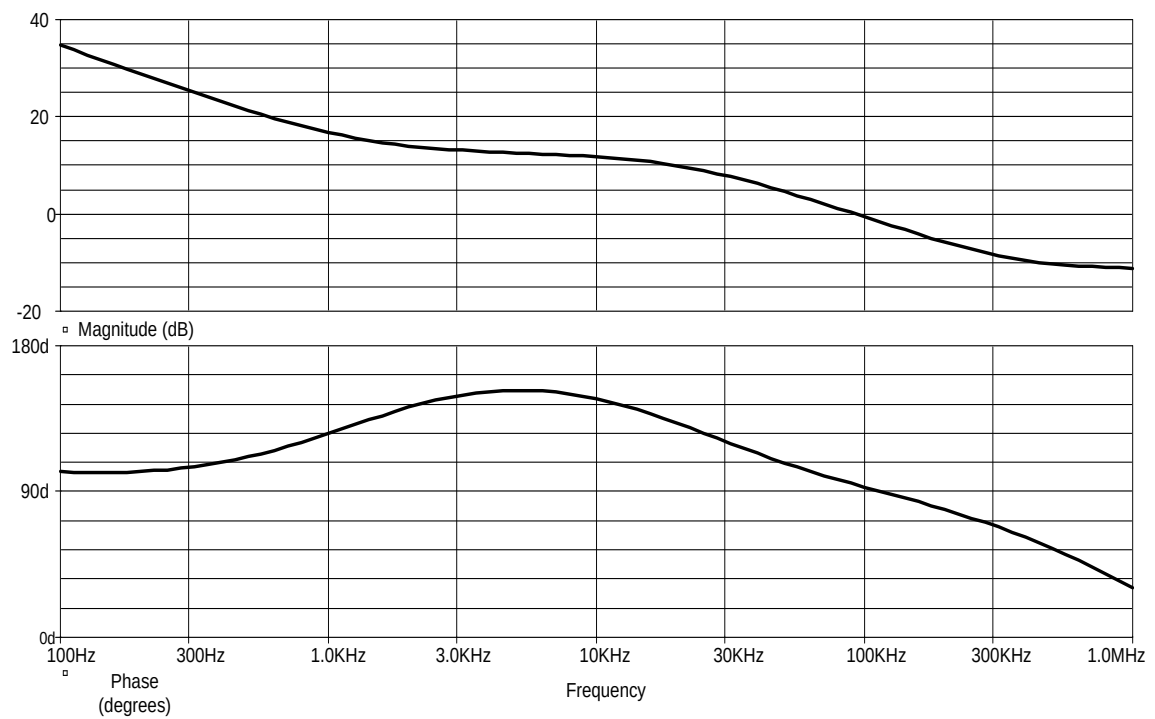


Fig. 14: Spice calculated error amplifier transfer function.

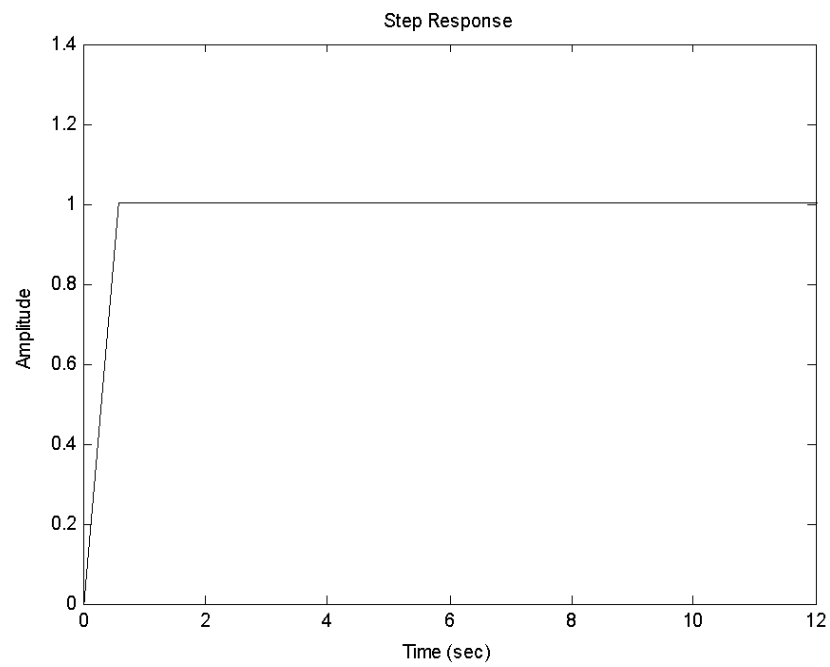


Fig. 15: Matlab calculated closed loop step response.

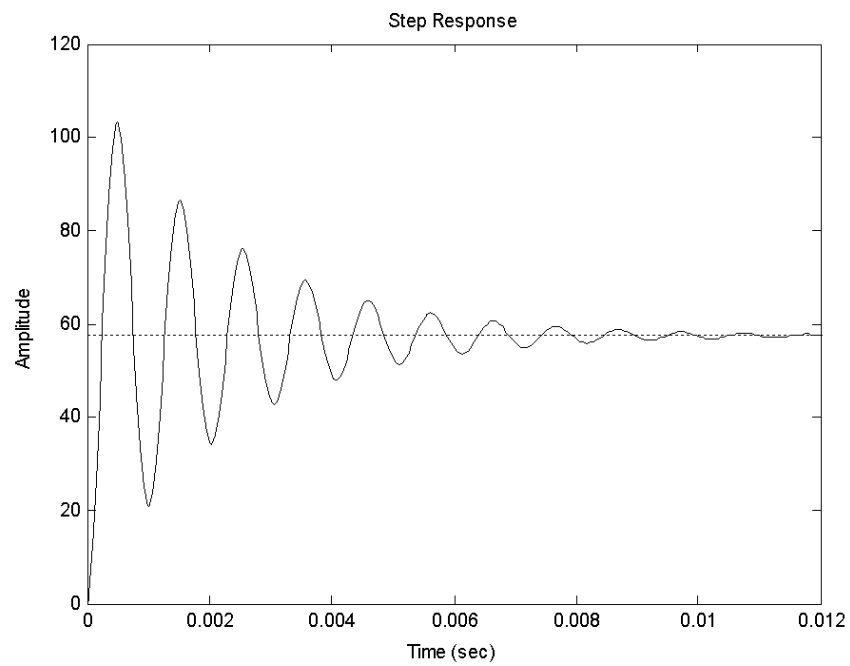


Fig. 16: Matlab calculated open loop step response.

Current Control: The current control circuitry works in a similar fashion as the voltage control. The output current is sensed and compared to a reference value through an error amplifier with control circuitry attached. The design of the current feedback system is similar to the voltage feedback; however, the current feedback must work at a low voltage, high current operating point where the gain of the open loop transfer function is very large. The gain is very large because at a low output resistance a small change in control makes a very large change in current. Figure 17 shows the bode plot of the current transfer function. It does not have the resonance peak that the voltage transfer function has.

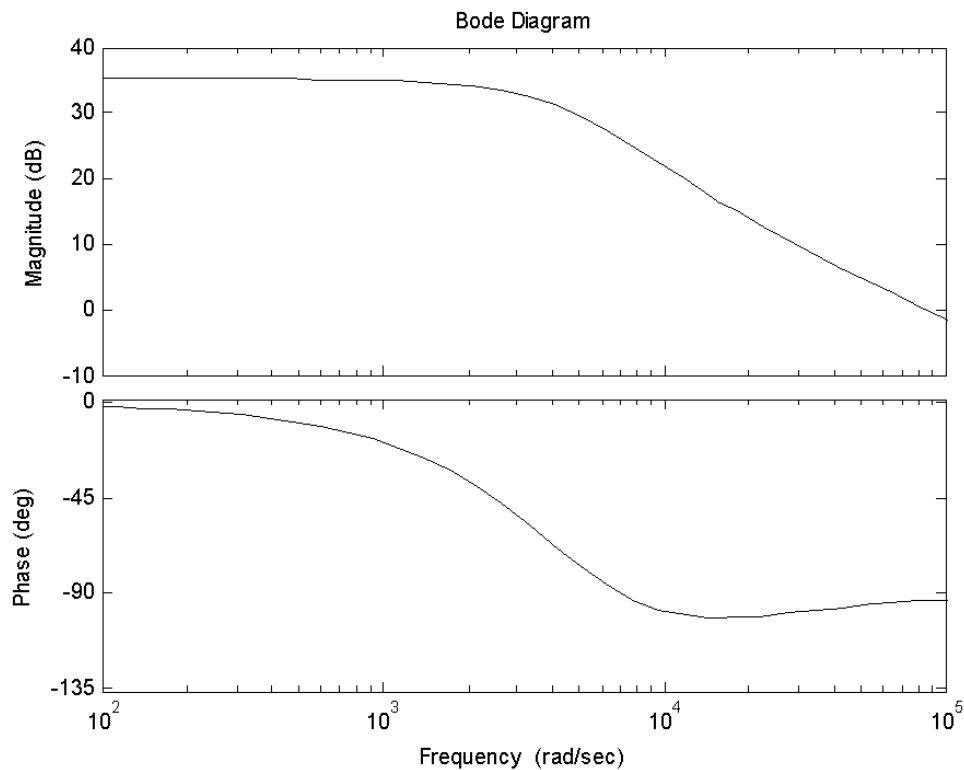


Fig. 17: Bode plot of current transfer function.

The control method outlined in [12] were developed and tested in the prototype with poor results. Instead, the current feedback from [10] was used. This feedback system does not have the extra pole in the transfer function. The error amplifier from [10] is shown in Figure 18.

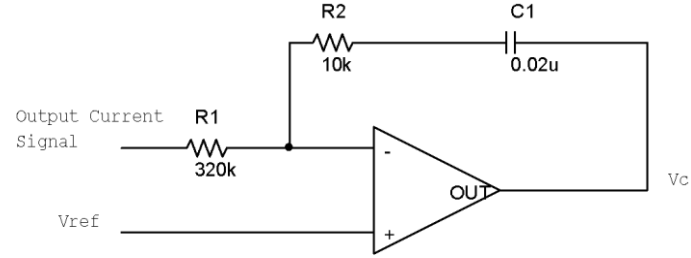


Fig. 18: Schematic of current EA.

The transfer function of this circuit is

$$T_c(s) = \frac{R2}{R1} \frac{s + 1/R2C1}{s}. \quad (17)$$

The error amplifier's transfer function using the values shown in Figure 18 is shown in Figure 19.

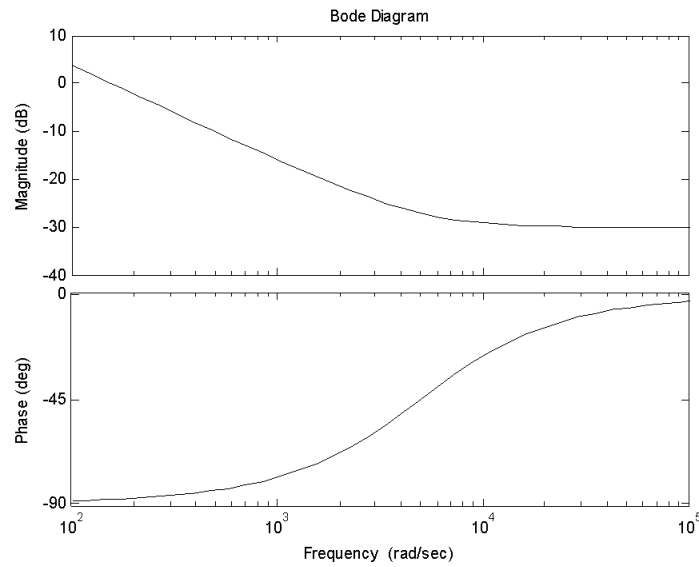


Fig. 19: Current EA bode plot.

The values given in [10] for R1-2, and C1 were tested experimentally in the circuit. It was found that the values $R1=320k\Omega$, $R2=10k\Omega$, and $C1=0.02\mu F$ work better than the original values. These new values have the same pole and zero as the original system, however, the gain is much smaller.

Automatic Crossover System: The voltage and current controllers work fine by themselves, but they need to work simultaneously. The desired effect of simultaneous operation is that the power supply has two modes, a constant voltage mode and a constant current mode. In regular operation the supply is in constant voltage mode where the output voltage is held steady and mirrors the control signal and the output current depends on the load. If the load begins to draw more current than a set limit, the supply goes into a constant current mode where the output voltage is variable while the current is set by an external reference. The supply can operate in either mode but not both at the same time. The automatic crossover circuit ensures that both modes do not happen at the same time but effortlessly change from one to the other.

The automatic crossover circuit is outlined in [13]. The circuit is shown in Figure 20. The circuit carries out an OR function between the outputs of the two error amplifiers. The amplifier with the lower output will cause its associated diode to conduct causing the other diode to be reverse biased. Therefore the lower output of the two amplifiers prevails and controls the output of the converter.

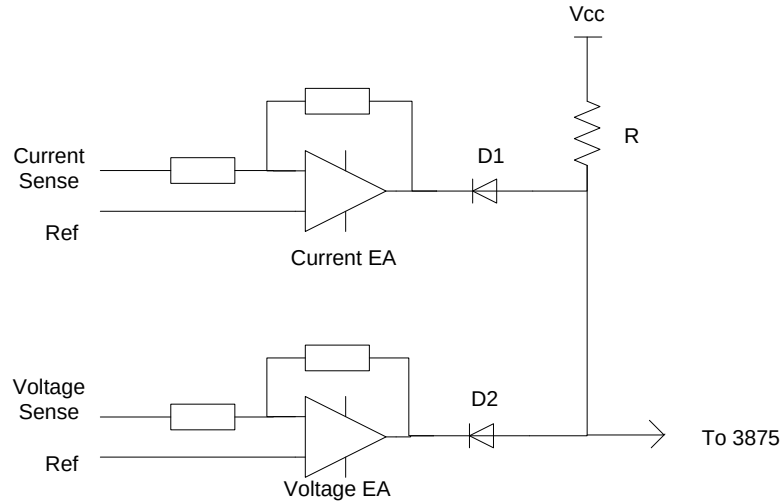


Fig. 20: Automatic crossover circuit block diagram.

Experimental Results

The Prototype

The prototype unit is shown in Figure 21. The unit consists of two components, the power factor corrector (PFC) and the full bridge unit. The PFC rectifies the input voltage and boosts the input voltage to 385V while ensuring that the input current is in phase with the input voltage. The ZVS component is the design developed in this thesis. The MOSFETs, diodes, and output filter are labeled in Figure 21. The primary side is at the top of the figure while the secondary side is at the bottom. The input connection to the PFC is near the large square capacitor while the output connector is the large metal bus bar connector at the bottom. The control circuit is a vertical mounted board, it is located on the secondary side close to the output. The control board is not easily visible in Figure 21; Figure 22 shows the control board in detail.

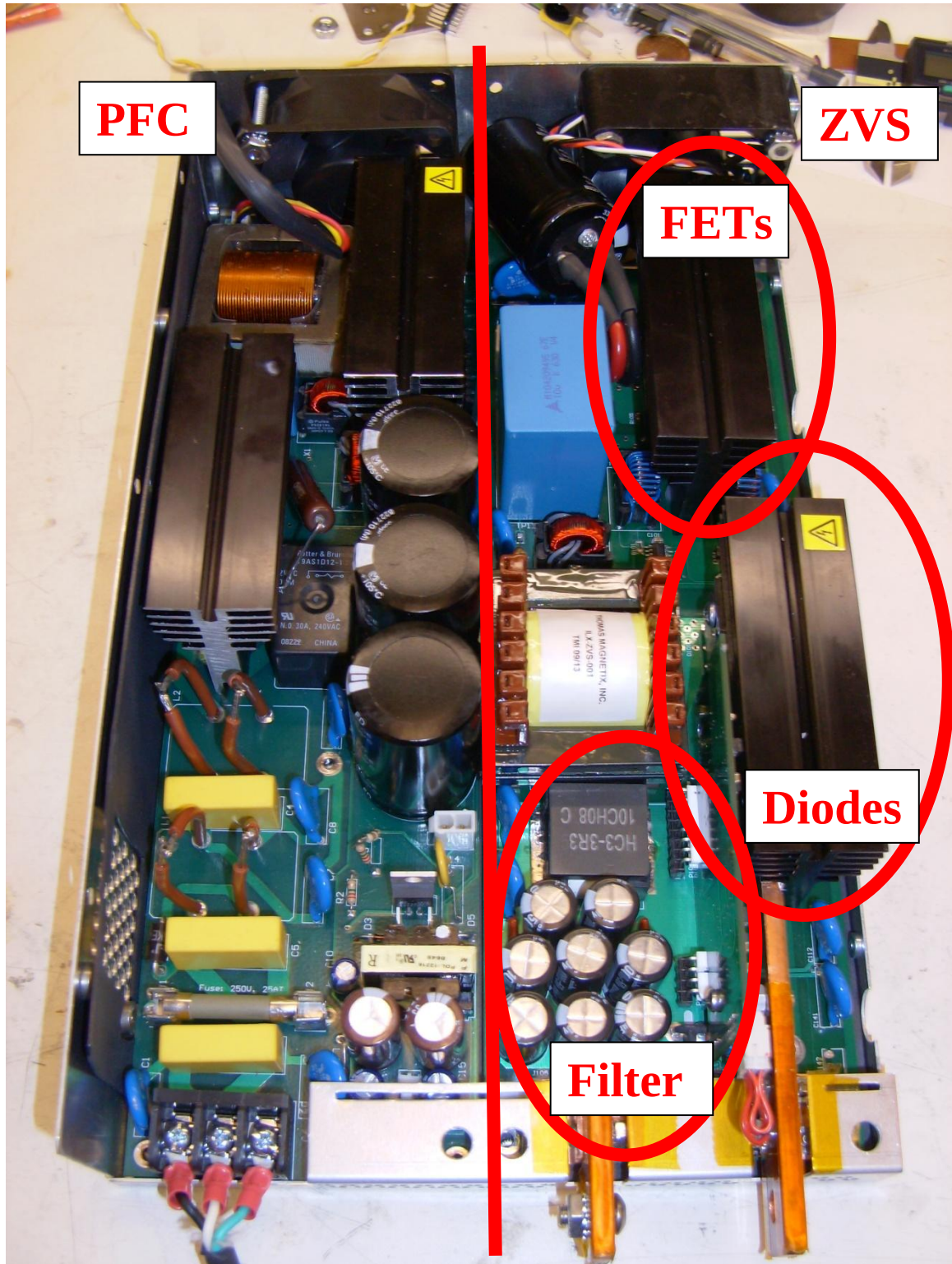


Fig. 21: ZVS prototype.

Figure 23 shows the switching MOSFETs in closer detail. They are clipped to a heatsink to manage the thermal waste. The clips are used for ease of manufacture. Figure 24 shows the control circuit in greater detail. It contains the UC3875 as well as all the control circuitry. Figure 25 shows the ZVS in its chassis. The chassis was custom made to house the ZVS. A design requirement is that the new converter matches the Cosel's form factor. Figure 26 shows the ZVS next to the Cosel for comparison. The ZVS is on the left and the Cosel is on the right. Note the four mounting holes on the top are in the same location for both. Additionally, the input and output connections are in the same location.

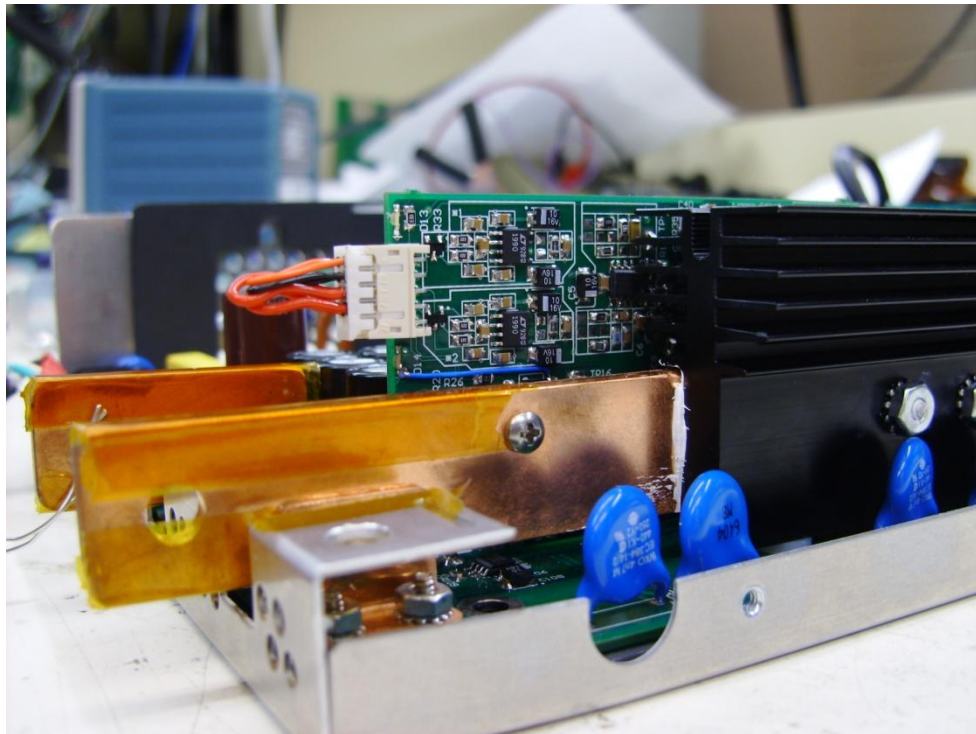


Fig. 22: Secondary side and control board.

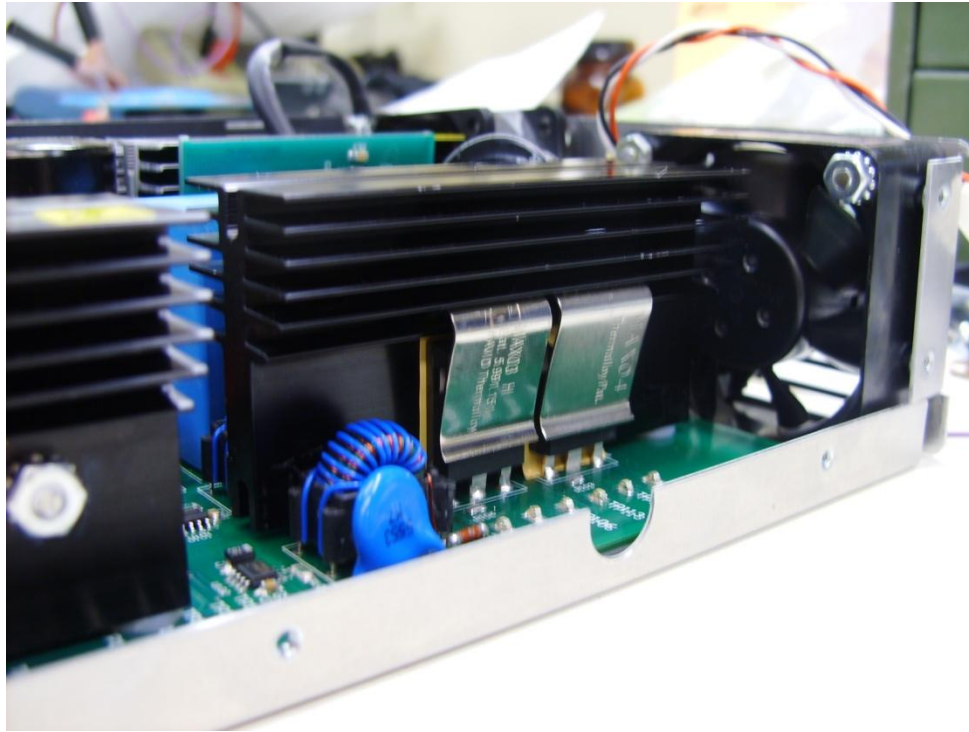


Fig. 23: Primary side

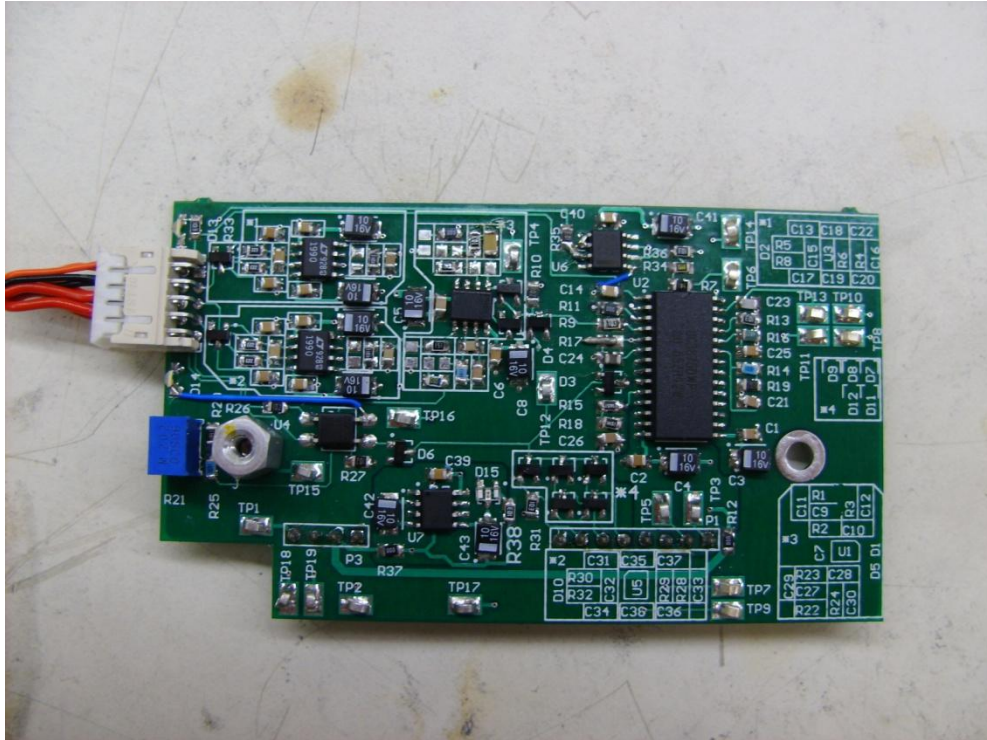


Fig. 24: Control board.

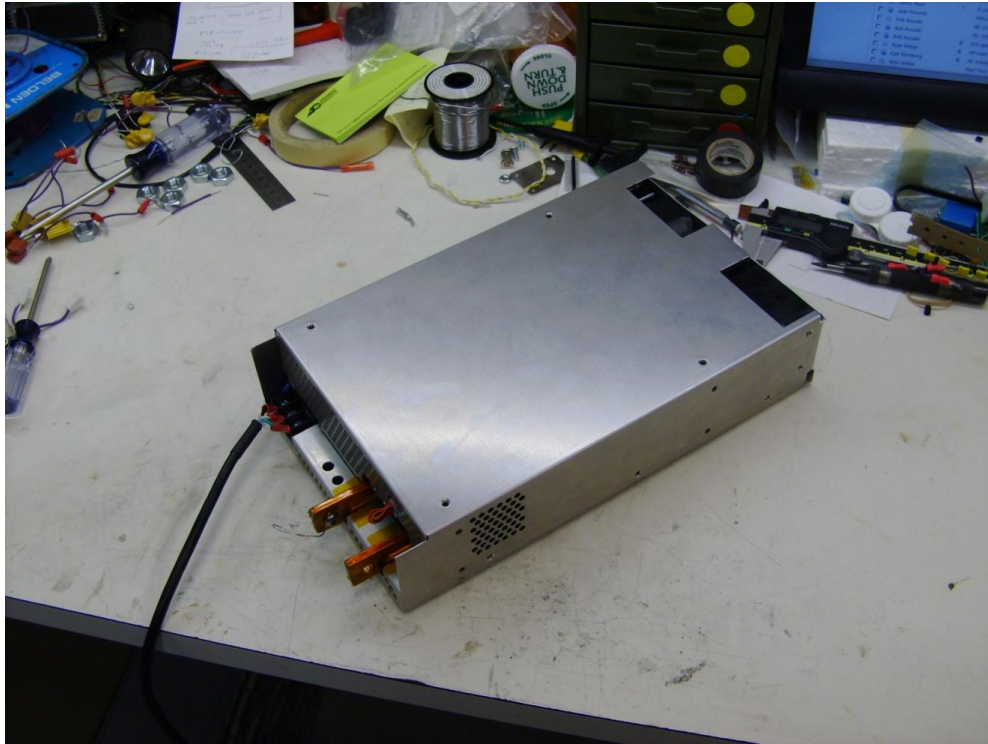


Fig. 25: ZVS with chassis.

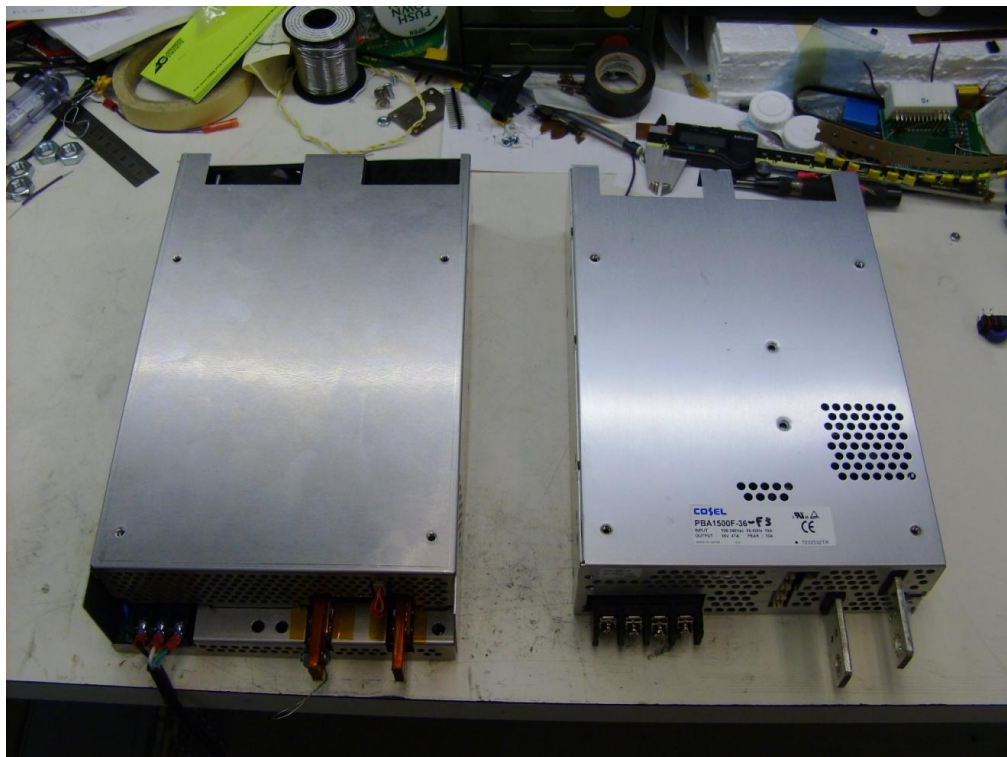


Fig. 26: ZVS versus Cosel.

Operation

The prototype was tested to see if it meets all the design criteria. Several waveforms were measured to verify the correct operation of the converter. These measurements include primary voltage, secondary voltage, output voltage, input current, and output current. These measurements are used to verify that the converter is operating properly, has achieved soft switching, calculate efficiency, determine output voltage quality and output current quality.

Figure 27 shows the drain to source voltage of two MOSFETs in the same leg. The figure shows that both FETs are switching in phase. Out of phase operation will cause a short of the input voltage through the FETs causing immediate failure or extreme losses.

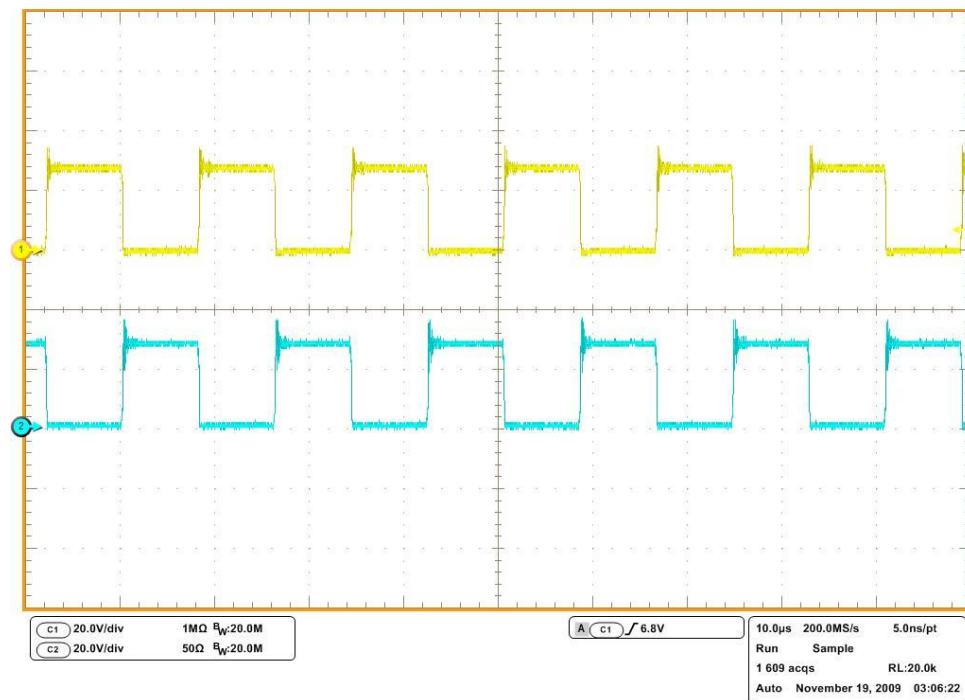


Fig. 27: Drain to source voltages in one leg.

To verify that the soft switching is working as expected, the gating signal and drain to source voltage of one MOSFET can be examined. To achieve zero voltage switching, the drain to source voltage must reach zero before the gating signal is applied. Figure 28 shows the gating signal and drain to source voltage simultaneously. In Figure 28 the drain to source voltage reaches zero before the gating signal is applied.

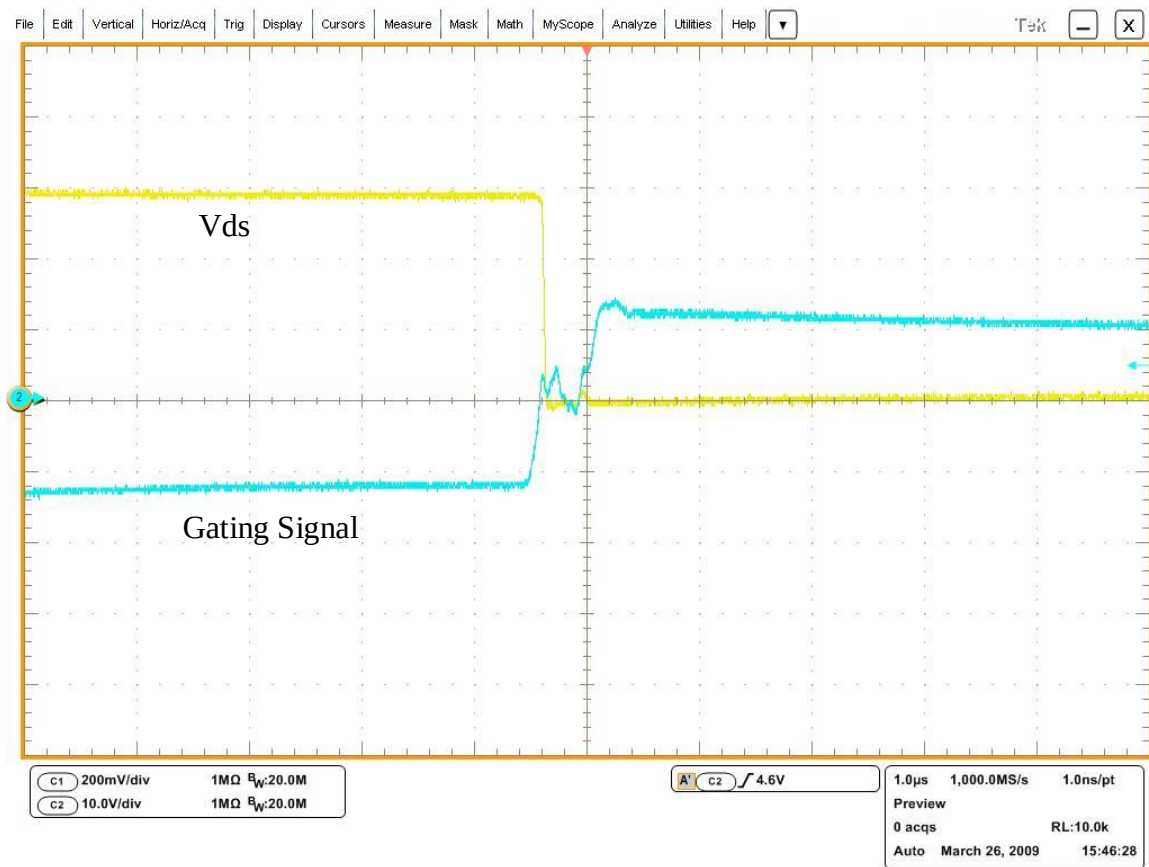


Fig. 28: Soft switching.

Figure 29 shows the primary and secondary voltage of the transformer at the same time. This figure shows three things. First it shows the H-Bridge is operating properly by applying a proper square wave to the transformer. Secondly, the secondary voltage

shows the transformer is working properly. Finally, the figure shows the duty cycle loss on the secondary side due to soft switching.

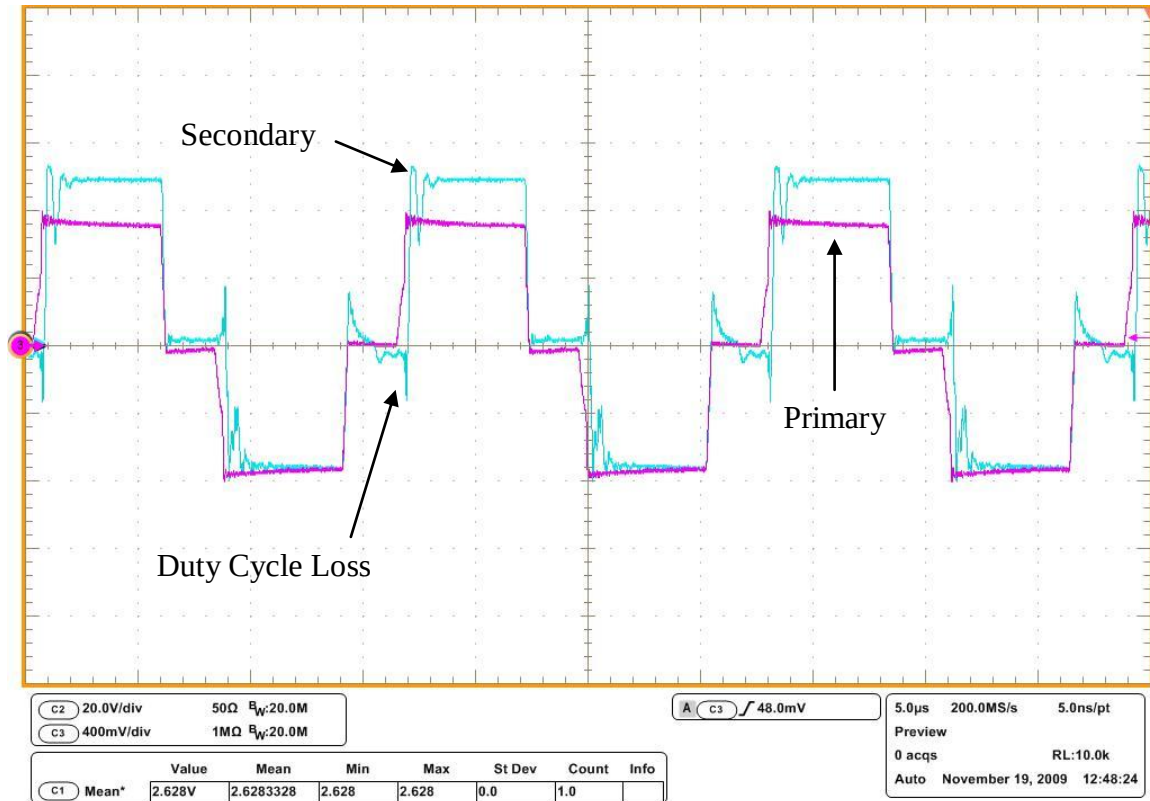


Fig. 29: Transformer voltages.

Figure 30 shows the output voltage and current. The figure shows that the output voltage and current are stable and have low noise, within the required specifications. In the figure the current is displayed as a voltage that is 1:1 equivalent to the current.

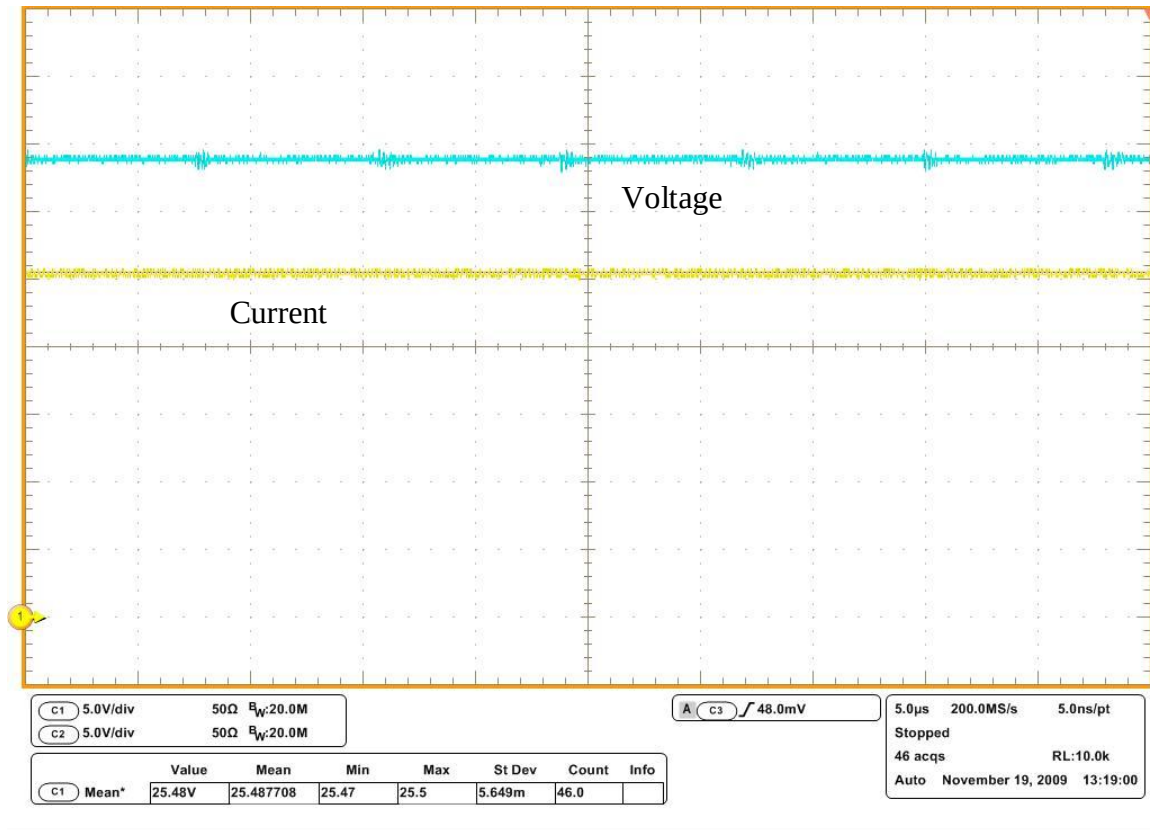


Fig. 30: Output voltage and current.

Efficiency

The efficiency was obtained simply by measuring the input and output voltage and current at full load. The efficiency is then calculated by comparing the output power versus the input power.

Running the converter at full load, the input voltage was measured at 385V and the input current was measured at 4.15A. The output voltage was measured at 37V and the output current was measure at 40A. The total output power is 1480W while the input power is 1598.4. This corresponds to an efficiency of 91.5%

Figure 31 shows the input voltage (via the primary voltage), input current, and output voltage on the same screen. Using a resistive load of 1.333Ω the efficiency of 91.5 is verified.

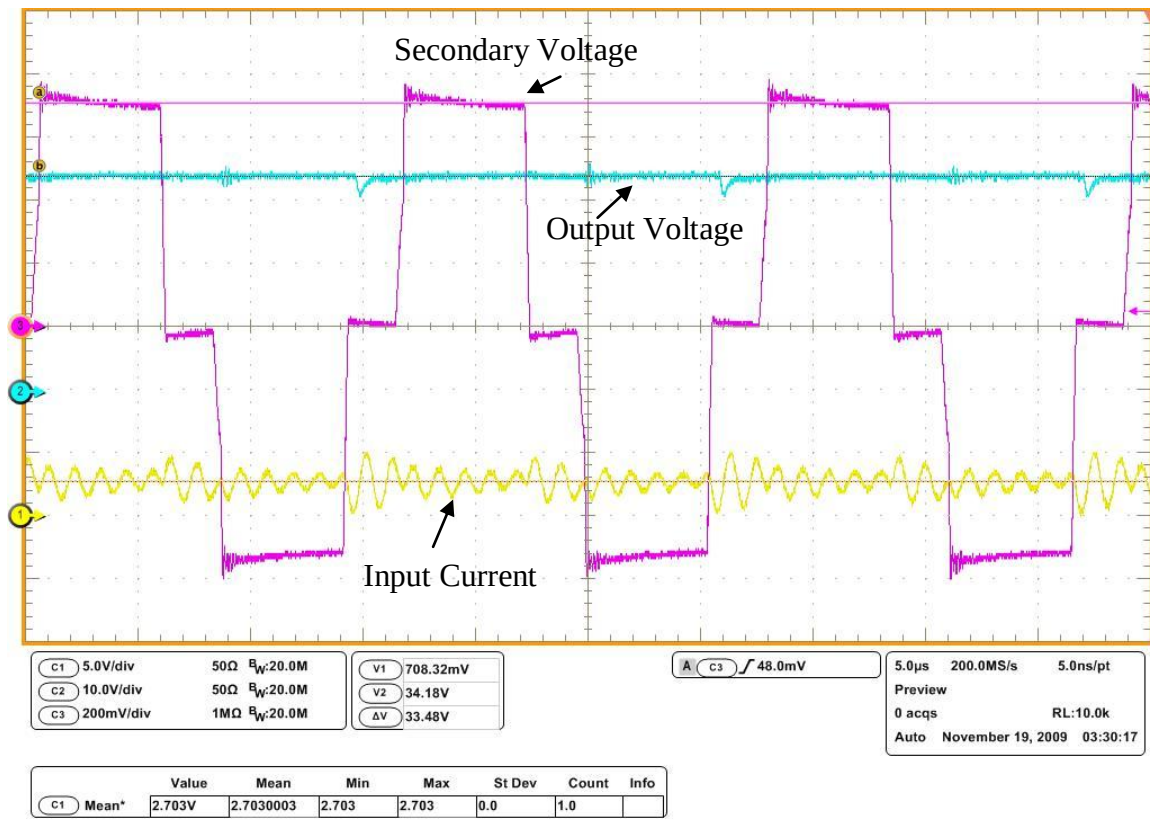


Fig. 31: ZVS efficiency.

Cost

The total cost of the converter can be broken down into four sections, the power circuit, the control circuit, the PFC, and packaging / ancillary hardware. The cost of the power circuit includes the power components, the circuit board and ancillary components such as the snubbers and EMI caps. The cost of the control circuit includes the control components like the UC3875, op amps, resistors and capacitors, and the circuit board.

The cost of the PFC includes all the power and control components of the power factor corrector as well as the circuit board and hardware. Finally, the cost of the packaging / ancillary hardware includes the cost of the metal box, the fans, the heatsinks, the screws, and other small non electrical components.

Table 3 outlines the cost each section of the converter. Additionally, table 3 outlines the cost of the major power components due to their large contribution to the overall cost.

Table 3: ZVS bill of materials.

<i>Item</i>	<i>Manufacturer</i>	<i>Model</i>	<i>Quantity</i>	<i>Price Each</i>	<i>Extended Price</i>
Power Circuit					
MOSFETs	Infineon	SPW20N60C3	4	\$3.56	\$14.24
Flux Runaway Cap	Epcos	B32676*6106	1	\$7.73	\$7.73
Transformer	Thomas Magnetics	Custom	1	\$27.10	\$27.10
Diodes	IXYS	DSSK 60-015A	2	\$6.50	\$13.00
Inductor	Cooper Bussmann	HC3-3R3-R	1	\$12.68	\$12.68
Capacitor	Panasonic	ECA1HM471	8	\$0.40	\$3.20
PCB	Various	ZVS36V40A	1	\$15.00	\$15.00
Power Circuit Subtotal					\$92.95
Control Circuit	Various		1	\$39.71	\$39.71
PFC	Various		1	\$149.97	\$149.97
Packaging	Various		1	\$132.68	\$132.68
Total Cost					\$415.31
Price Per Watt					\$0.28

Conclusion

A high efficiency, low cost, DC/DC converter for laser test equipment was designed, built and tested. Using ZVS full bridge technology, an efficiency of 91.5% was achieved at a cost of \$0.28 per watt. The cost beat the goal by \$0.02 per watt. The prototype was tested for efficiency, output noise, stability, and overall performance. The converter performed within all specifications. Using these measurements it was verified that soft switching was achieved during full load conditions.

Several tests that could not be performed at the university are currently being done at ILX Lightwave. These tests include HIPOT testing, thermal testing, safety verification, durability, and compatibility with the PFC. A properly designed full bridge converter can easily pass all these tests.

A possible improvement in efficiency could be achieved by careful transformer design. A high quality transformer coupled with careful soft switched delay times can improve the efficiency by another 2-3%. An engineer must carefully tweak the delay times to improve the efficiency the 2-3%. Design time was not allotted for this adjustment due to time requirements.

DC/DC CONVERTER FOR RESIDENTIAL FUEL CELL APPLICATIONS

Previous Work

Several designs have been proposed to meet the low cost and high efficiency goals set by the DOE. The first designed analyzed in this paper was the winner of the 2003 DOE/IEEE Energy Challenge. The design was proposed by the Seoul National University of Technology in Seoul, South Korea [4]. This DC/DC converter is shown in Figure 32. The reported efficiency and cost of this design is 90% and \$45.18/kW.

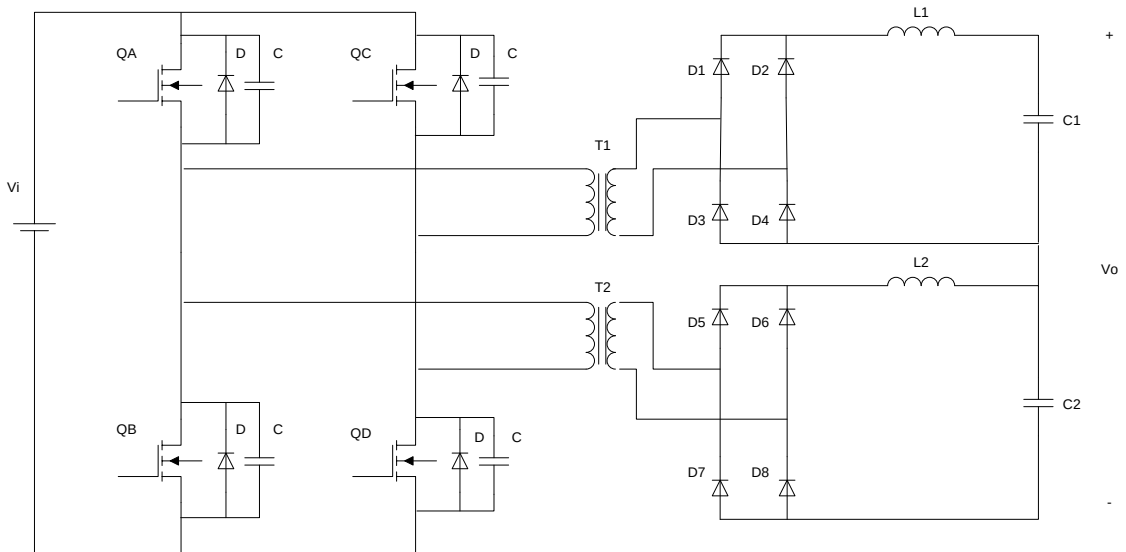


Fig. 32: Topology proposed by the Seoul National University of Technology [4].

This design achieves the low cost high efficiency goals using two primary methods. First, the design uses zero-voltage switching to minimize the switching loss. The second method is a combination parallel/series connection of DC/DC converters. The basic system consists of two transformers, rectifiers, filtering inductors and

capacitors that are used to share the power. The two transformers are connected in parallel on the primary side and the two DC outputs created by the secondary windings of the transformer, the rectifiers, filter inductors and capacitors are connected in series. The parallel connection of the transformers on the primary side allows the two transformers to share the high input current evenly. The series connection of the two DC output allows the turn ratio of each transformer to be smaller. This improves the stability of the system and improves the duty cycle loss from the soft switching circuitry by reducing the leakage inductance of the transformers. The dual transformer design also helps reduce the material cost by reducing the current rating of the transformer.

Other secondary factors help reduce the cost of the design as well. Careful selection of the switching MOSFETs is a large contributing factor to cost reduction. This design uses the IXYS IXFN340N07 MOSFET. This FET is a high quality, low cost semiconductor. It is rated for 70V and 340A. It may seem that this FET is overqualified for the 190A required of it. The high current rating is chosen to minimize the drain-to-source resistance. The IXFN340N07 has a R_{ds} of 4 milliohms. Minimizing the R_{ds} minimizes the conduction loss of the FET, improving the overall efficiency of the converter. The IXFN340N07 retails for about \$6 in large quantities making the cost due to the FETs \$24 per unit.

Despite the success of this design, it does suffer from some drawbacks. First, the MOSFET in the H-bridge carry the full input current. When carrying the full input current the conduction losses in each FET is quite high. Additionally, the switching loss, even with soft switching, is higher because of the large rate of change and large currents

moving through stray inductances in the wires reduce the effectiveness of soft switching. Secondly, the series connection of the DC outputs creates a complicated control scheme. Due to unavoidable mismatches in the components the system must be carefully controlled so that the current and voltage is evenly shared between the two transformers [4]. Finally, at light loads the efficiency of the converter drops because at light loads the criteria for soft switching cannot be maintained. The failure of the soft switching reduces efficiency because the duty cycle loss in the soft switching is no longer present causing a higher output voltage. This higher output voltage requires a smaller duty cycle in the switching FETS, causing higher conduction losses.

The second topology analyzed in this thesis is the topology proposed by Virginia Tech in Blacksburg, Virginia, USA [14]. Their design is shown in Figure 33.

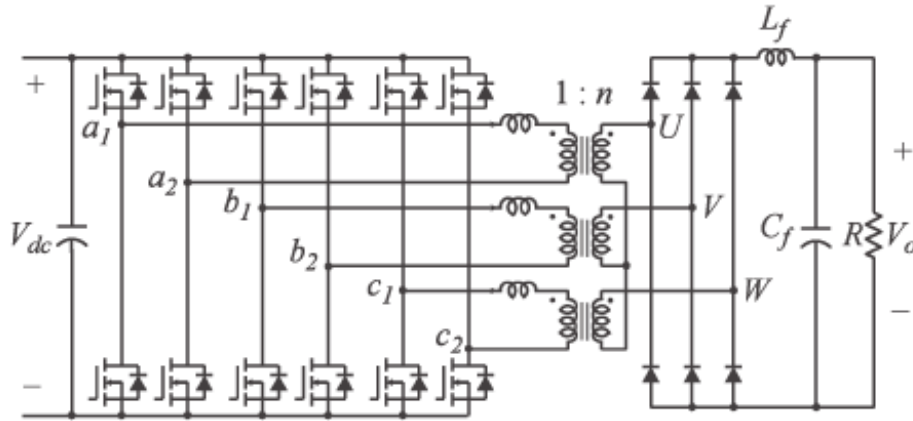


Fig. 33: Topology proposed by Virginia Tech [14].

The reported efficiency and cost are 96% and \$45/kW, respectively [15]. The high efficiency and low cost of this design is achieved in several ways. First, zero voltage switching is utilized to reduce the switching loss. Next, the topology uses a three phase transformer to improve the efficiency. The three phase transformer allows the input

current to be shared between several transistors on the primary side. This reduces the conduction loss in the switching FETs. The three phase transformer can be thought of as three single phase transformers connected in parallel on the primary side and in a combination of series and parallel on the secondary side (two in parallel connected in series with the other). This connection allows the turn ratios of the transformers to be reduced, therefore reducing the leakage inductance which improves the soft switching performance. The connection of the three phase transformer on the secondary side also contributes to a greater efficiency. The connection reduces the number of diodes on the secondary side, reducing the losses due to diode voltage drop. Finally, interleaving control is used to reduce the size and cost of the filtering inductor and capacitor. Due to the three phase transformer, this topology does not suffer efficiency loss at light loads. This is because the circulating current in the three phase transformer helps maintain soft switching over a wide load range.

Like the topology from the Seoul National University of Technology the design does have shortcomings despite its excellent efficiency at both heavy and light loads. First, the input current is not evenly shared between the MOSFETs on the primary side [14]. This is because only two of the three h-bridges are on at any time. This uneven current sharing creates a higher conduction loss in the FETs. Secondly, the secondary sides of the transformers are connected in such a way that two transformers are connected in parallel and the third in series. Although this configuration helps create the excellent efficiency it has two issues that can reduce the efficiency. First, the current sharing becomes a problem in the parallel connected transformers. Exact current sharing cannot

be guaranteed, the uneven sharing can cause undue stress in the MOSFETs and the rectifying diodes. Secondly, the parallel combination does not effectively minimize the turn ratio of the transformers creating possible stability issues and efficiency loss.

Both papers analyzed in this thesis have very high efficiencies and low cost. However, by utilizing the advantages and minimizing the disadvantages of both, a new DC/DC converter can be developed.

A New Converter for Residential Fuel Cell Applications

There are two major challenges in the design of a new converter. The first challenge is the extremely high input current. The high current causes very high conduction losses in the switching devices. Additionally, expensive, high current, switching devices must be used to cope with the large currents. A method of managing the input current must be utilized to reduce the conduction losses and cost. The second challenge is the low input voltage. To meet the output voltage specification the input voltage must be boosted significantly. The simplest way of boosting the voltage is to use a high turn ratio transformer. However, the high turn ratio can cause stability problems and create complications in soft switching cycle. Both topologies analyzed in the previous section find ways to overcome these challenges. The best parts of each of these converters can be used to create a new converter topology.

The topology utilizing these advantages is shown in Figure 34. This topology has three features to deal with the high input current, improve the efficiency, and reduce the cost. The topology can be thought of as three soft switched full bridge topologies with

their inputs connected in parallel and their outputs connected in series. The parallel input is similar to the technique used by Virginia Tech. The connection allows the input current to be split up between the three units, reducing the current through each MOSFET, reducing the conduction loss.

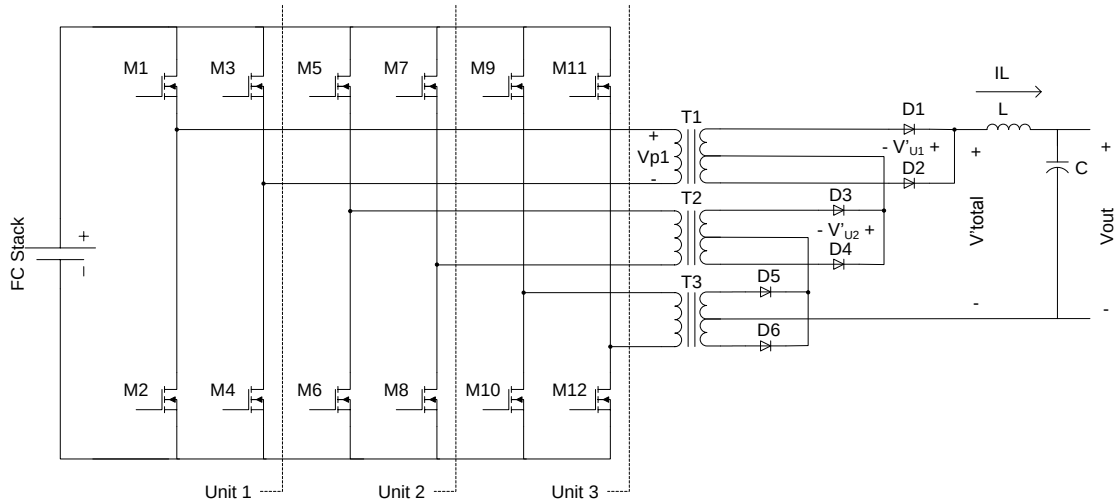


Fig. 34: Topology of new converter.

The second feature is that the switching cycles of the converters are interleaved by 120° . In interleaved control the turn on points of each are staggered so no two units turn on at the same time and one unit is always on at any time. This control triples the effective switching frequency on the series secondary side reducing the required filter size. The smaller filter requirement reduces the cost of the system and improves the quality of the output voltage.

The third feature is the series connection on the secondary side provides three benefits to the system. First, the series connection adds the output voltages of each converter unit. This technique allows the turn ratio of each transformer to be smaller because the output of each unit is $V_o/3$ versus V_o . A smaller turn ratio is beneficial

because it reduces the leakage inductance of the transformer improving the soft switching cycle, reducing the duty cycle loss, and improving the stability. The second benefit is the series connection ensures that the current in each unit is shared evenly, thereby eliminating the need for any external current sharing control. The third benefit is the most unique feature of the topology, the ability to turn off one unit at light load to maximize the efficiency of the converter at light loads.

The input to output voltage transfer function of this converter is based off the full bridge transfer function in equation 7. The transfer function is simply three times the full bridge transfer function due to the three unit's output stage in series. The new transfer function is

$$V_o = 3 * V_{in}(D) \frac{N_s}{N_p}. \quad (18)$$

Light Load Operation

During operation at light loads the output characteristics of the fuel cell stack must be analyzed. Figure 35 shows the output voltage-current characteristics of a typical fuel cell.

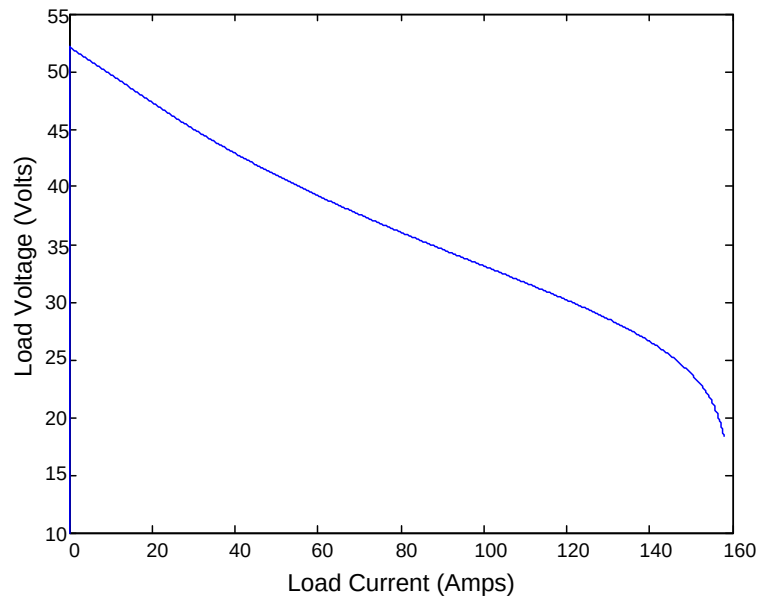


Fig. 35: SOFC output V-I curve.

Figure 35 shows that during large output loads the output current is high while the output voltage is small. At this point the converter must provide a large boost to maintain the 400V output voltage. At light loads the output current decreases while the output voltage increases. At this point the converter must provide a smaller boost to maintain the output voltage.

Two things happen to the converter at a light load. First, soft switching fails at light loads because the output current is not high enough to empty the resonant tank and create zero voltage switching. Secondly, the smaller boost requirement reduces the duty cycle of the converter. The smaller duty cycle increases conduction losses in the MOSFETs. Turning off a unit eliminates both problems with light load operation. Disabling a unit forces the two remaining units to provide more boost. Additionally, the disabled unit increases the power requirements of the two remaining units. The increase in power allows the remaining units to achieve soft switching even during light loads.

A possible problem with deactivating a unit is that the output of each unit is no longer $V_o/3$, but $V_o/2$ requiring a larger duty cycle reducing the conduction loss. Since the duty cycle is already close to 100% the 400V output voltage cannot be maintained. The remaining two converters are able to provide the necessary boost due to the doubling of the input voltage from the stack. The equation for the output voltage is now

$$V_o = 2 * V_{in} * N. \quad (19)$$

Assuming the input voltage remains constant and using equation 19, the output voltage would be two thirds the required output voltage. However since the input voltage effectively doubles at a light load the output voltage can be thought of as

$$V_o = 4 * V_{in}(\text{heavy load}) * N. \quad (20)$$

Due to equation 20 the two remaining units have enough boost to maintain the required 400V. If the input voltage does not increase with a light load it would not be possible to turn off a converter.

Finally, the phase shift of the two remaining converters is changed to 180° during light loads. With one converter deactivated the switching becomes unbalanced using 120° shift. Changing the phase shift to 180° creates a balanced system and improves the quality of the output voltage.

Switching Operation

Figure 36 shows the switching operation of the new converter. For simplicity the hard switched cycles will be analyzed using the nomenclature defined in Figure 34. The soft switching transitions are small compared to the longer hard switch cycles.

At time $t=0$, switch pair M1,M4 in unit 1, M6,7 in unit 2 and M10,11 in unit 3 are on. This places V_{in} across the primary windings of T1 (V_{p1}) and $-V_{in}$ across the primaries of T2 and T3. These voltages are passed to the secondary sides of the transformers. D1 conducts in unit 1, while D4 and D6 conduct in units 2 and 3 respectively. The output voltage of the rectifiers in each unit ($V'_{U1,2,3}$) is the input voltage times the turn ratio or $V_{in} \cdot N$. Due to the series connection of the rectifiers the voltage at the inductor (V'_{total}) is the $V'_{U1} + V'_{U2} + V'_{U3}$ or $3 \cdot V_{in} \cdot N$.

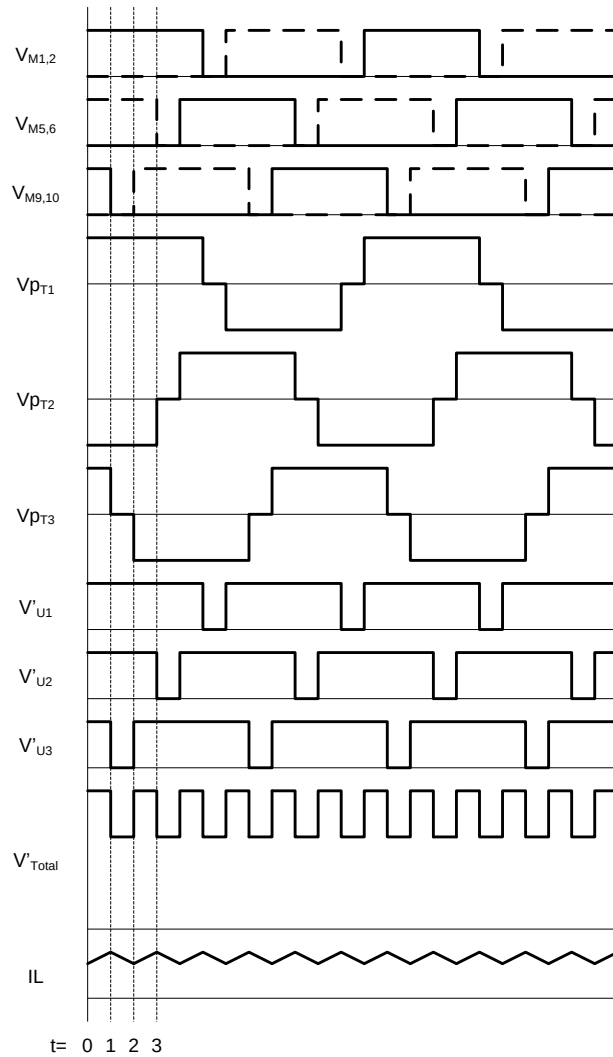


Fig. 36: Switching waveforms for the new topology.

At time $t=1$, switch pair M1,4 and M6,7 remain on while pair M10,11 are turned off. The output of the rectifiers in units 1 and 2 remain at $V_{in} \cdot N$ while the rectifier of unit 3 is zero. At this point the rectifier acts like a short to the rest of the circuit and currently freely flows through both diodes in unit 3. The voltage at the inductor is now the sum of unit 1 and 2 or $2 \cdot V_{in} \cdot N$ with unit 3 providing no voltage.

At time $t=2$, switch pair M9,12 in unit 3 is turned on while M1,4 in unit 1 and M6,7 in unit 2 remain on. This state is similar to time $t=0$ where the rectifier outputs of each unit are $V_{in} \cdot N$. The inductor voltage is the series combination of the three units and is equal to $3 \cdot V_{in} \cdot N$.

At time $t=3$, switch pair M6,7 is turned off while M1,4 and M9,12 remain on. This state is the same as $t=1$ where one unit is turned off. Again, the inductor voltage is $2 \cdot V_{in} \cdot N$.

Figure 36 shows that the switching frequency at the inductor is triple the switching frequency at the rectifier and six times the switching frequency at the MOSFETs. This allows the MOSFETs to be switched at a low, more efficient frequency, while creating a large switching frequency at the inductor creating a lower cost inductor.

Design of the Converter

Power Circuit

Like the full bridge converter for industry the design of the converter uses the input and output specifications to design the components. The design starts with the transformers, then the MOSFETs and then followed by the diodes and output filter.

The transformer begins the same way as before. The turn ratio, input, output voltage and current must be accounted for. According to the specifications, the input voltage is 22-41V while the output voltage is 400V for the entire converter. Using equation 7 the turn ratio can be calculated. Since there are three converters connected in series the output voltage is one third the total output power. Using the minimum input voltage and assuming the maximum duty cycle of 95% the turn ratio is calculated using

$$400/3 = 22(D) \frac{N_s}{N_p}. \quad (21)$$

According to equation 21 the turn ratio needed is 6.06. Knowing the output voltage and the power requirement the output current is 12.5A and the input current is 75.75A.

These specifications were given to a transformer company, American Magnetics, for construction. The turn ratio was modified to be 6.66 to compensate for the soft switching duty cycle loss. The data sheet is shown in the appendix. The total cost of the transformer for large quantities is \$25 each.

The MOSFETs are designed according to the total input voltage and the current of each unit. The input current of 75A for each unit was found designing the transformer. The transistor used in [4] is used in this design. This transistor (IXYS IXFN340N07) is a very high quality, high current, low Rds device. At \$6 each, it is a very reasonable price.

The diodes are selected using the voltage at the secondary side and the output current. The peak voltage at the secondary side using the maximum input voltage of 41V is 492V. The necessary current is the same as the output current or 12.5A.

The 600V 16A FEP16JT by Fairchild Semi is selected. It has a 1.5V drop at full load which reduces losses and costs \$0.35 in large quantities.

The inductor is calculated using equation 11. Due to the interleaved switching the frequency is three times the switch frequency. Like the industrial converter a switching frequency of 50 kHz was chosen for optimal efficiency and cost. This means the inductor frequency is 300 kHz. The desired current ripple and transformer voltages are entered into equation 11

$$L = \frac{(419-400)0.5}{0.2(12.5)300000} = 13\mu H. \quad (22)$$

The 15.6μH, 15.6A inductor 2306-V-RC by Bourns was chosen. It has a cost of \$1.62 in large quantities.

Finally the capacitors are chosen. Six Panasonic (EEU-EB2W470) 47uF, 450V capacitors are used. They were chosen as a good balance between cost, capacitance, and small ESR. Their cost is \$0.87 in large quantities.

Control Circuit

The most important design goal of the control circuit is to maintain a 120° phase shift between the units as well as provide the delays necessary to maintain soft switching. To accomplish this design goal the control circuit consists of three UC3875s. Like the industrial converter the 3875s provide all the necessary controls to maintain soft switching. The table calculating the necessary transition times is provided in the appendix.

To interleave the three controllers a microcontroller is used. The microcontroller sends sync pulses, delayed by 120° to the oscillators of the UC3875s. An analog solution is possible; however, a microcontroller allows easy adjustment of the delays and

frequency of the controllers. The microcontroller also allows a unit to be turned off while changing the phase shift to 180° at the same time.

The converter will be run open-loop for verification of the design. No control loops are needed. The converter is simply fed a constant voltage to its error amplifier.

Preliminary Results

The Prototype

Figure 37 shows a picture of the prototype converter. On the left side of the prototype are the switching MOSFETs and the control circuitry. The copper rings in the middle are the transformers and finally on the right is the secondary side consisting of the diodes, inductor and capacitors.

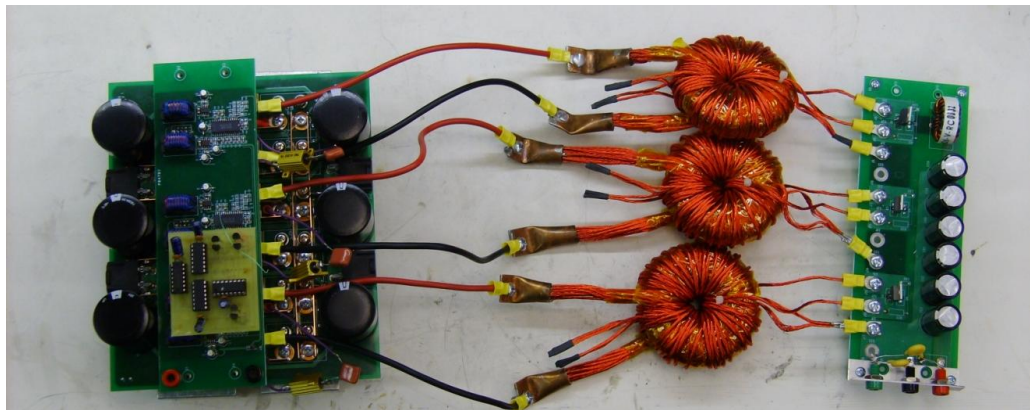


Fig. 37: Prototype converter.

Figure 38 shows a closeup of the primary side including the MOSFETs and the control circuit. The control circuit was placed in this location to minimize the noise on the control signals. The control board contains three UC3875s and has the necessary gate drive components to drive the MOSFETs. The control board also consists of a separate clock board that is attached separately. Figure 39 shows the primary board

without the control circuit. The connectors between the capacitors are the high current input connectors. Copper bus bars were used to carry the high current to the switches. Fence style bus bars were used to easily shed heat from the copper and reduce losses due to the skin effect.

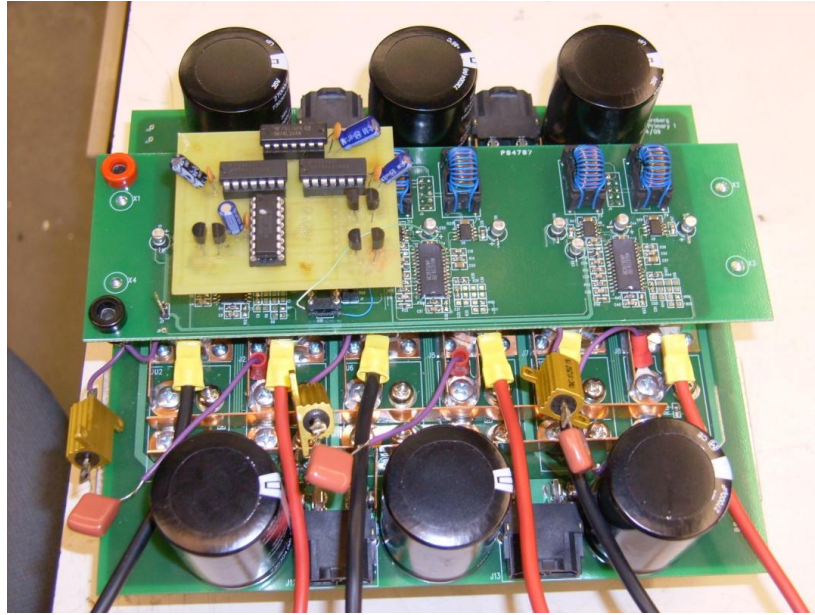


Fig. 38: Primary side.

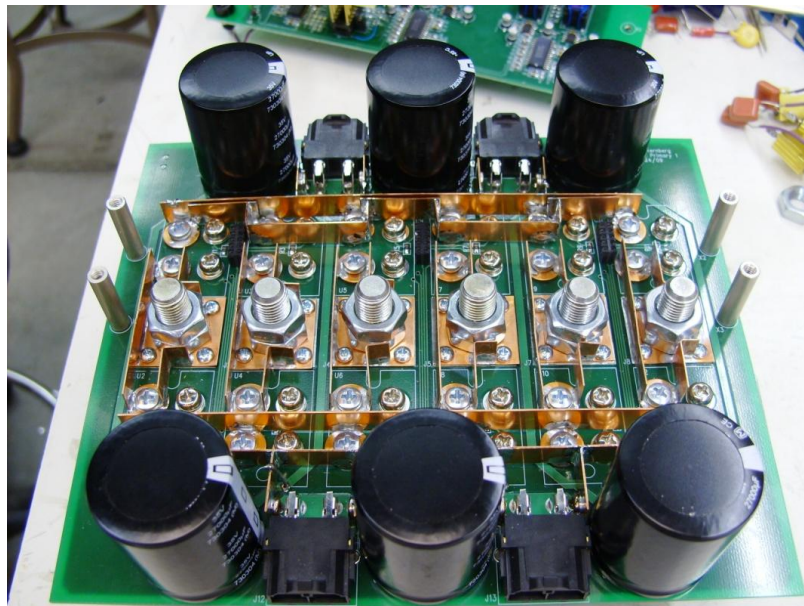


Fig. 39: High current bus bars.

Figure 40 shows the secondary side containing the diodes, inductor and capacitors. The connectors on the right are the connections for the load.

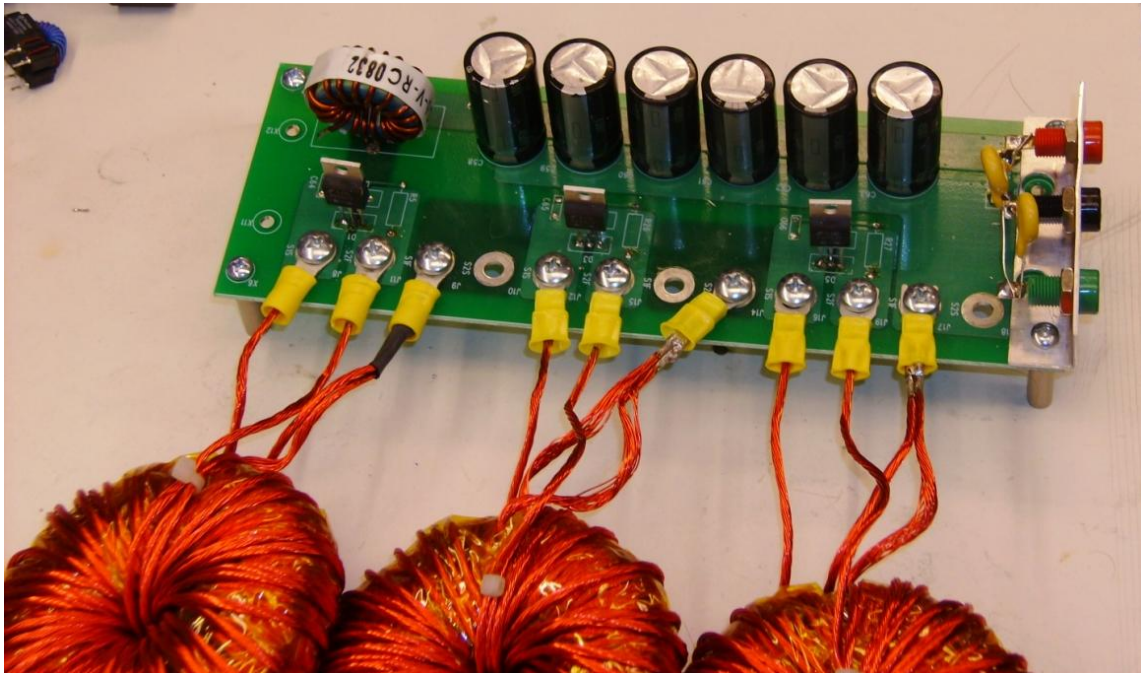


Fig. 40: Secondary side.

Operation

Due to time constraints, the converter was tested at a light load with all three units operating to verify proper switching operation. Several voltages and currents throughout the converter were tested to verify operation. The 120° gating signals, the primary and secondary voltages, the output voltage of each unit, the series combination of each unit's output, the inductor current, the output current ripple, and the output voltage.

The first voltage to be measured was the primary voltages of each transformer, these voltages show two properties of the converter, the 120° phase shift and the correct operation of each unit's H-bridge. Figure 41 shows this measurement.

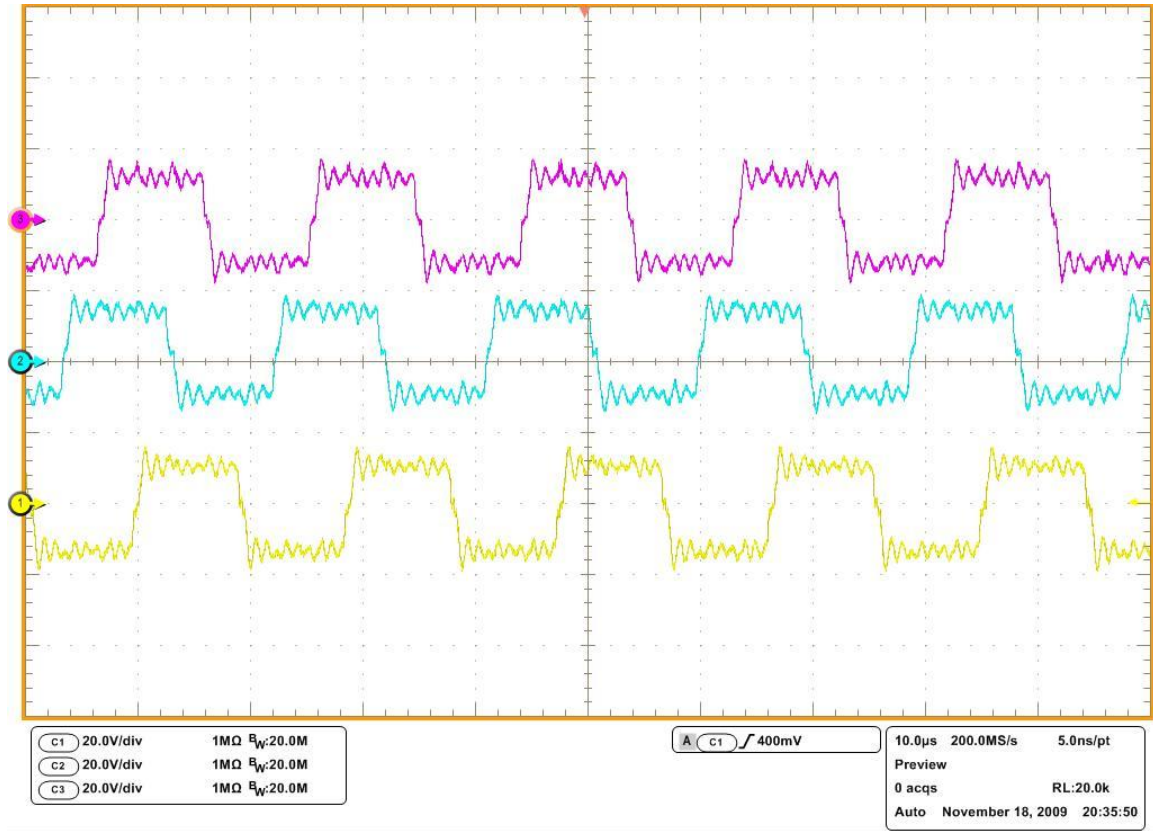


Fig. 41: Primary voltages with 120° phase shift.

Figure 42 shows the secondary voltages. The voltage at the secondary side is significantly higher than the primary. Additionally the secondary voltages maintain the 120° phase shift.

Figure 43 shows the output voltage of each unit ($V'_{U1,2,3}$ from Figure 34) as well as the inductor voltage (V'_{total} from Figure 34). Figure 43 shows the series addition of the output voltages and the resulting inductor voltage. Figure 44 shows a MOSFET gating signal and the inductor voltage. This shows that the series combination of units multiplies the FET gating frequency by six. The zero spots in the output voltages can be

correlated to the flat spots in the inductor voltage. This can be seen in the switching signals in Figure 36.

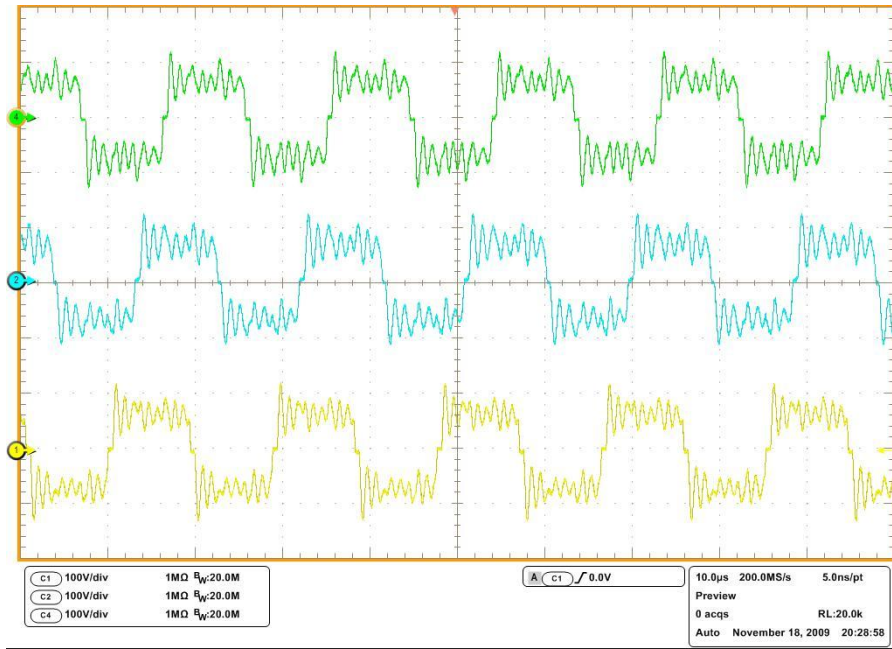


Fig. 42: Secondary voltages.

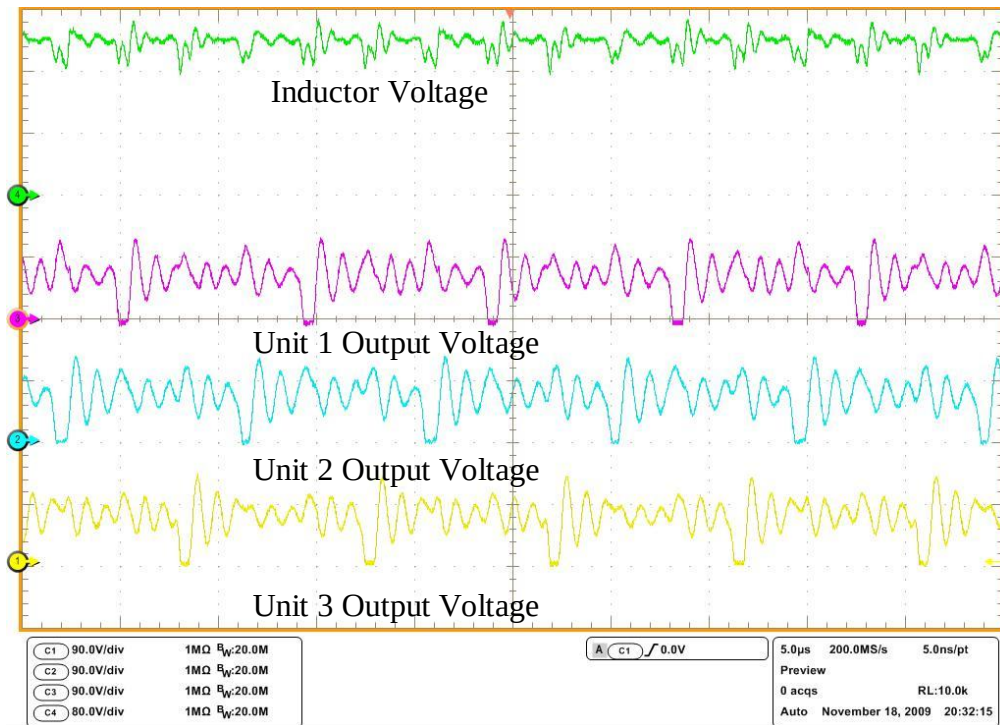


Fig. 43: Unit output voltages and inductor voltage.

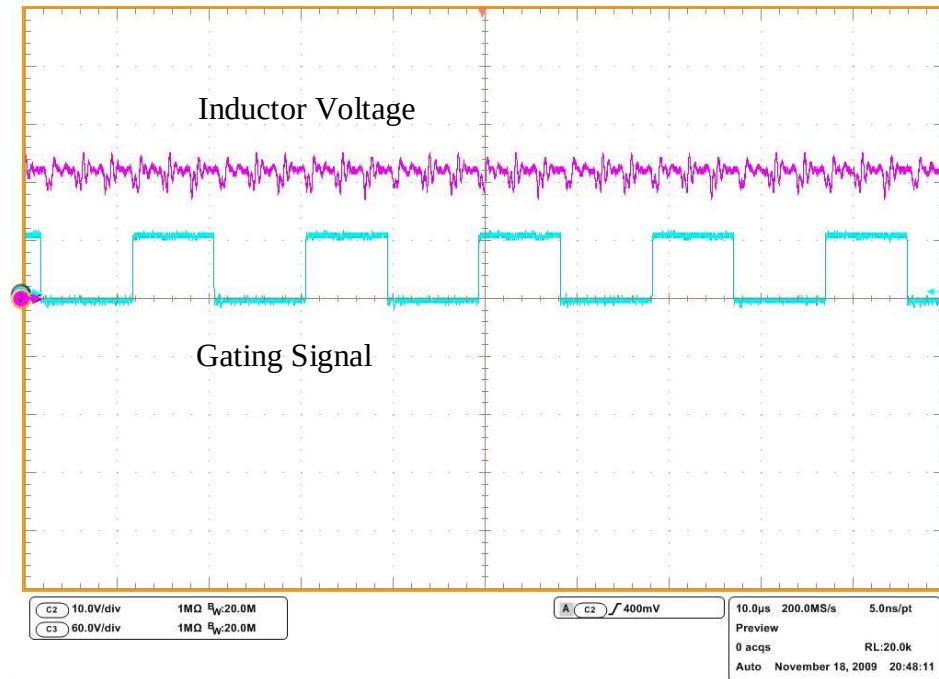


Fig. 44: Gating frequency versus inductor frequency.

The inductor current, and output voltage is shown in Figure 45. The inductor current matches the triangular waveform shown in Figure 36. The current rises when the voltage across the inductor is positive, and decreases when the voltage across it is negative. The output voltage shown in the graph is very clean. The voltage also abides by the input to output voltage transfer function from equation 18.

Finally, the output current was measured. The output current is shown in Figure 46. The output current ripple can be seen in the figure. When measured, the current ripple is exactly 20% the average current value. This verifies the inductor calculation and selection.

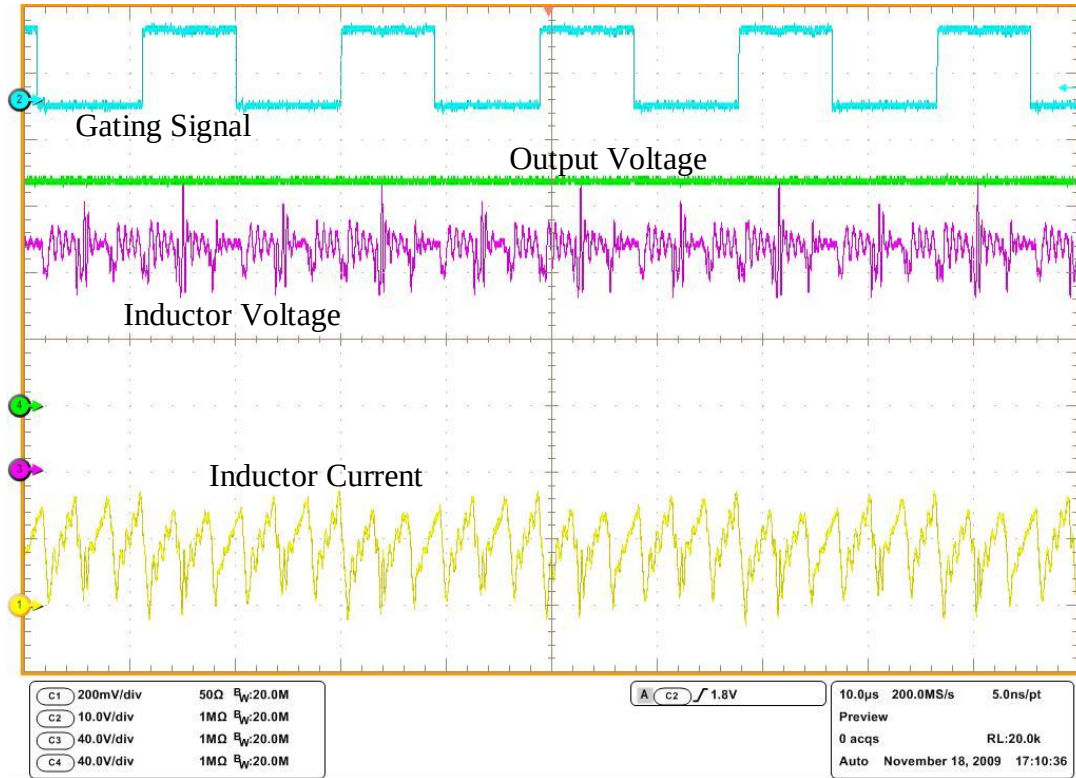


Fig. 45: Inductor current and output voltage.

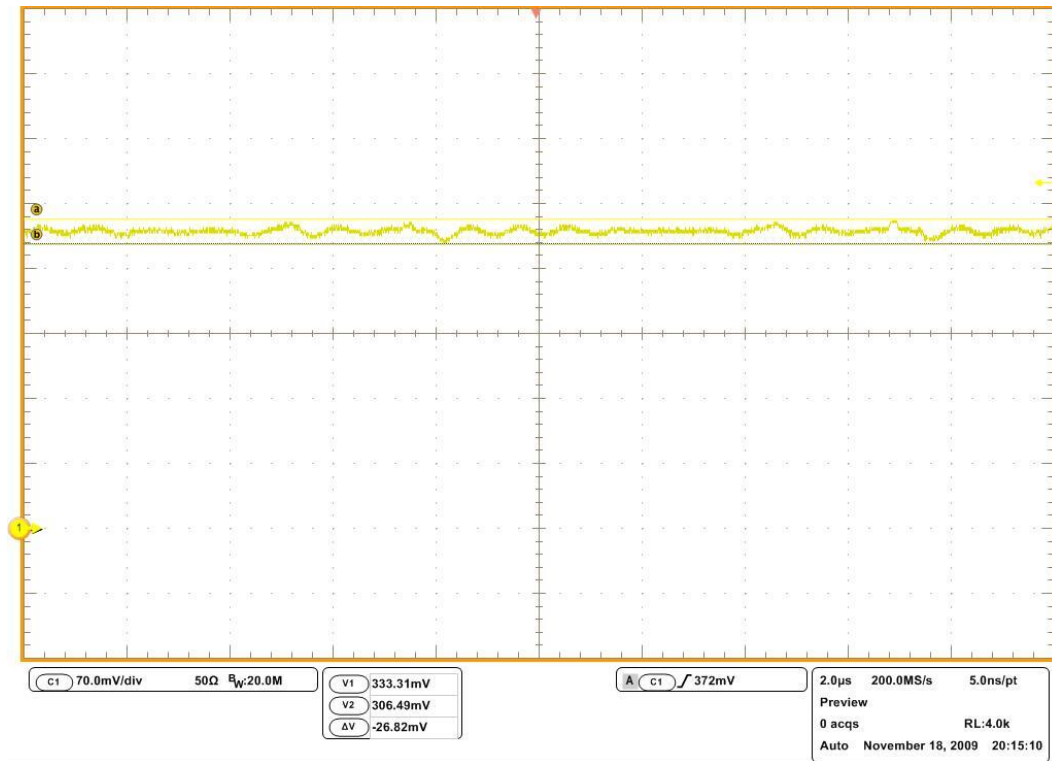


Fig. 46: Output current.

Efficiency

The efficiency was calculated the same way as the converter for industry. At the light load the efficiency was calculated at 65%. Although this value is small, the efficiency was measured at an output power of 75 watts. This is about 1% of the rated output power. Full bridge converters, soft switched or not, have very poor efficiency at extremely light loads.

Without running the converter at large loads, the efficiency can be estimated based off component ratings. The majority of the losses occur in the semiconductors and the transformer. The losses in the FETs come from two sources, the conduction loss and the switching loss. The conduction loss can be calculated from the R_{ds} of the FET. Using the input current calculated in the design section and an R_{ds} of $4\text{m}\Omega$, the conduction loss is 102 watts. The switching loss cannot easily be calculated directly from circuit parameters. An estimate of the switching loss can be found from other work with similar power level converters using soft switching [4].

The transformer losses include conduction loss and eddy losses. Like the switching loss in the FETs these losses are difficult to calculate. An estimate of the based off previous work is about 6 watts.

Finally, the losses in the diodes come from the voltage drop of the PN junction and the current flowing through the diodes. The forward voltage drop of the diodes is 1.5 volts. Using ohms law the losses in the diodes is 54 watts.

Table 4 outlines the efficiency. The total heat loss in the converter is 182W. Given an input current of 5000 watts the estimated efficiency is 96%.

Table 4: Losses summary.

MOSFETs	Conduction Loss	102W
	Switching Loss	20W
Transformer	Core and Winding Loss	6W
Diodes	Conduction Loss	54W
Total		182W

Cost

Like the converter for industry the cost of the residential converter can be broken down into three sections, the power circuit, the control circuit and ancillary components. The power circuit is the majority of the cost and is broken down into its constituent components. The control circuit consists of many small components whose cost is summed up into a lump sum. The ancillary components consist of heatsinks, fans, and non electrical components. Table 5 outlines the cost of each section as well as the total cost.

Table 5: Residential converter bill of materials.

<i>Item</i>	<i>Manufacturer</i>	<i>Model</i>	<i>Quantity</i>	<i>Price Each</i>	<i>Extended Price</i>
Power Circuit					
MOSFETs	IXYS	IXFN340N07	12	\$6	\$72
Transformer	American Magnetics	AM-19488	1	\$25	\$25
Diodes	Fairchild Semi	FEP16JT	6	\$0.35	\$2
Inductor	Bourns	2306-V-RC	1	\$1.62	\$2
Capacitor	Panasonic	EEU- EB2W470	6	\$0.87	\$5
PCB	Various		1	\$10	\$10
Hardware	Various		1	\$20	\$20
Power Circuit Subtotal					\$136
Control Circuit	Various		1	\$20	\$20
Ancillary Components	Various		1	\$40	\$40
Total Cost					\$196
Price Per Kilowatt					\$39.19

Conclusion

In this section a DC/DC converter for residential fuel cell applications was designed built and tested. This converter was designed to meet the 2003 DOE/IEEE Future Energy Challenge criteria for SOFC powered residential power generation systems. These requirements pose several challenges for DC/DC design. A low voltage, high current input converter with a large output voltage requirement is a challenging requirement. The design becomes more challenging when low cost and high efficiency

requirements are added. A new converter topology was constructed to manage these challenges.

A converter utilizing several technologies is designed to manage these challenges. These technologies include, phase-shifted soft switching full bridge converter, multi DC/DC converter units with a unique parallel input / series output connection, interleaving control, and the capability to modify the topology at a light load to improve efficiency.

Preliminary testing shows the new topology performs as expected providing an estimated 96% efficiency at a cost of \$39 per kilowatt. Further testing is required to obtain exact efficiency requirements at full load and light loads. The topology has the ability to contend with leading converters in fuel cell power management.

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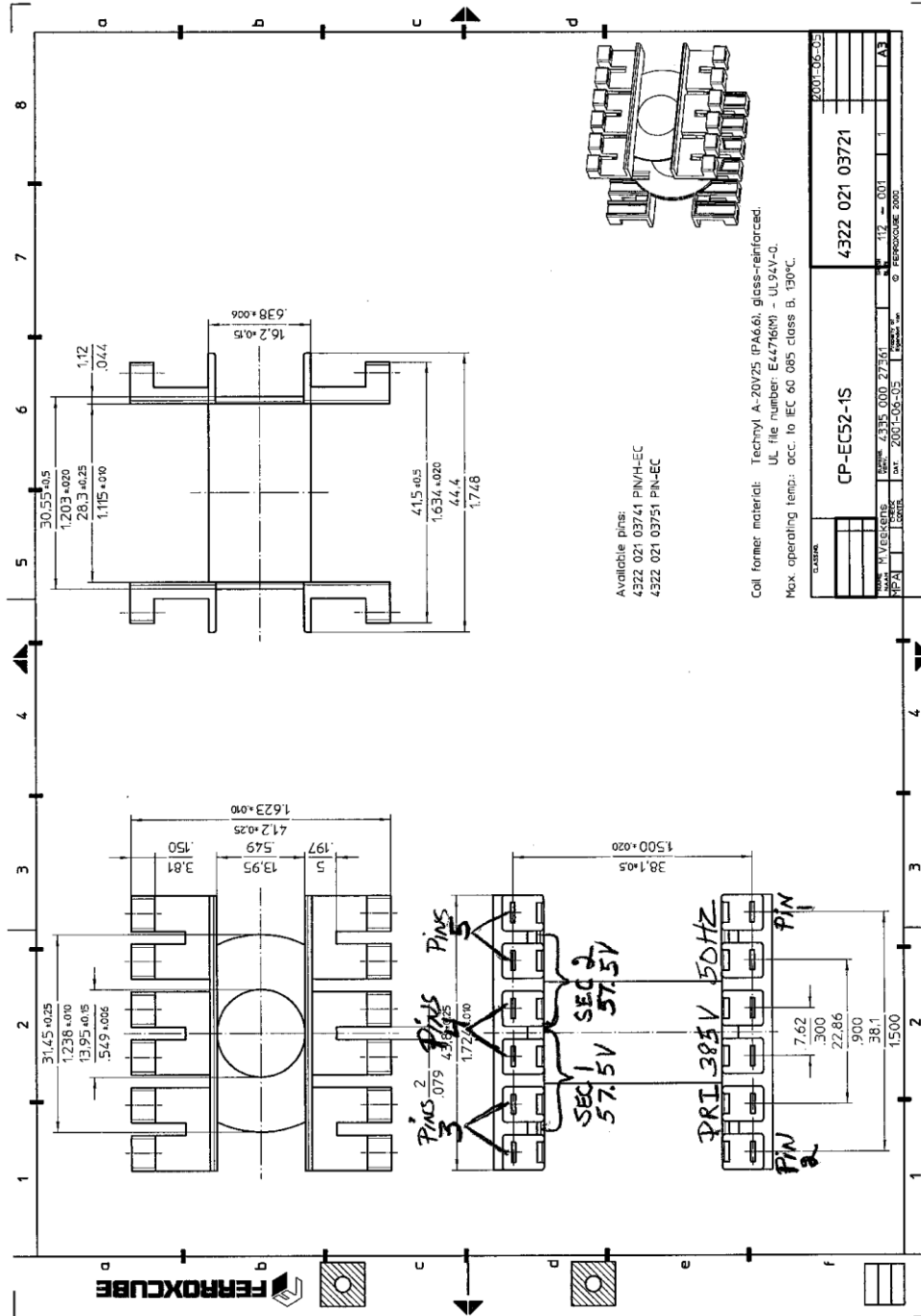
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APPENDICES

APPENDIX A:

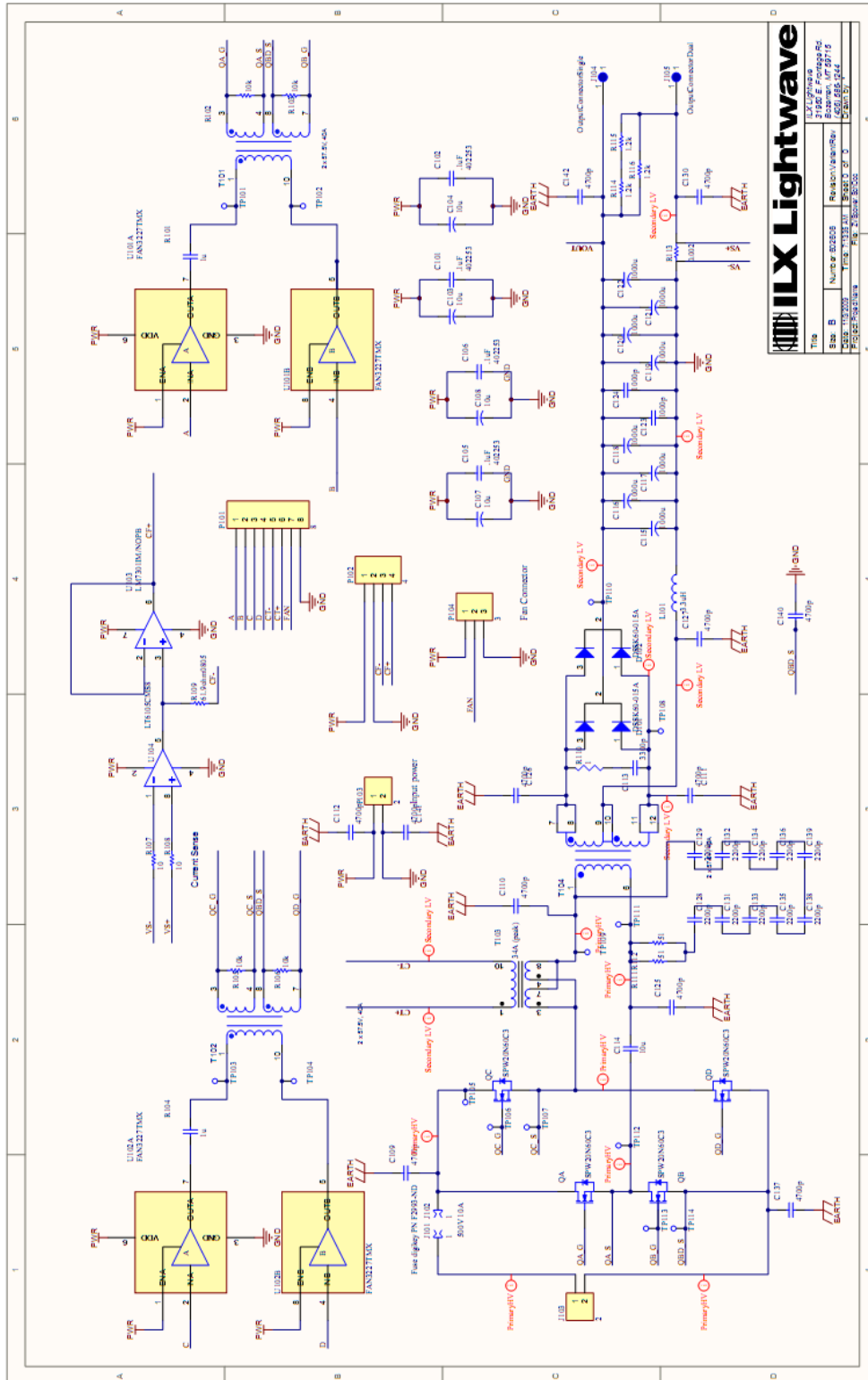
ILX TRANSFORMER DATA SHEET.



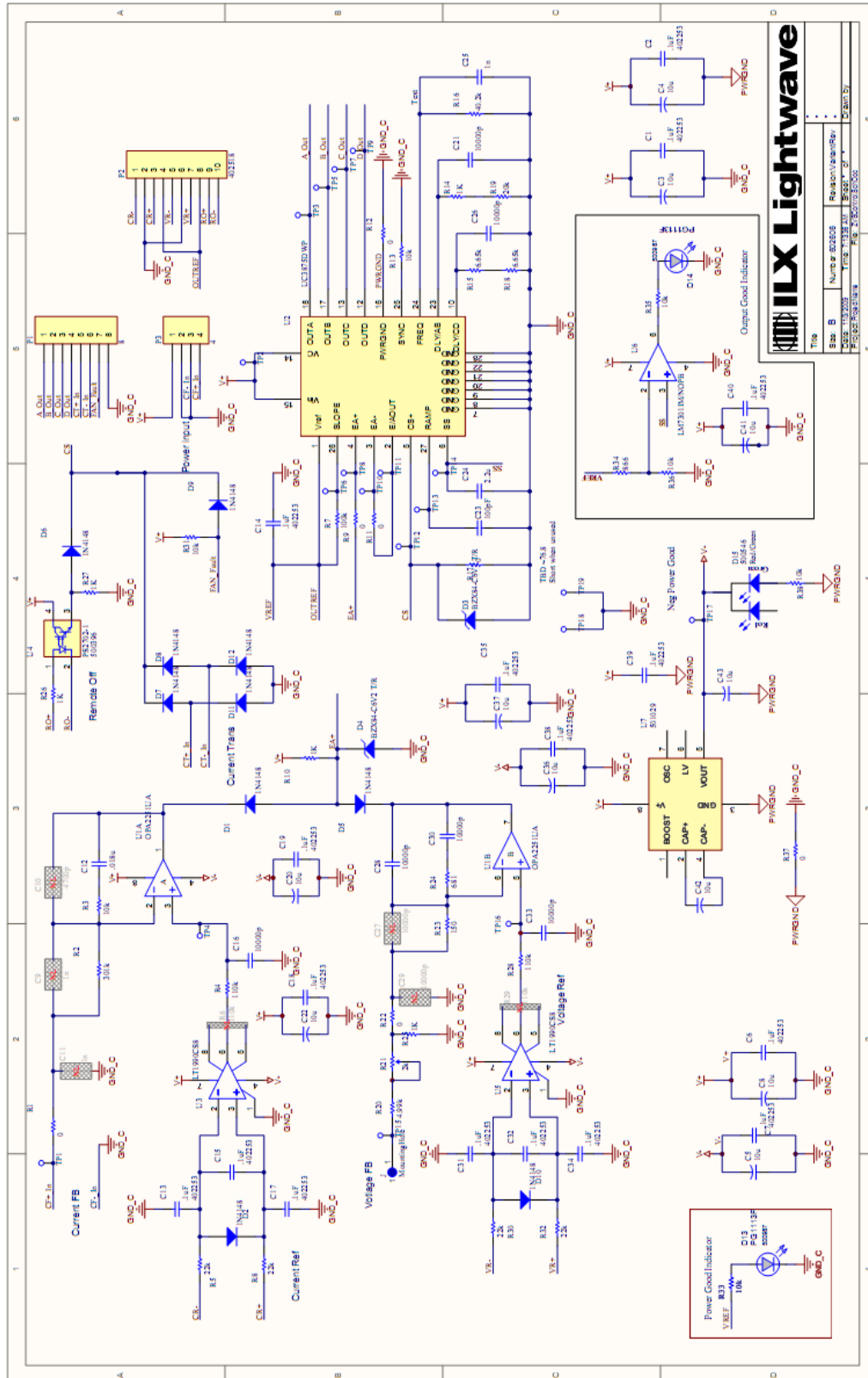
APPENDIX B:

ILX SCHEMATICS.

Power Circuit:



Control Circuit



APPENDIX C:

ILX ZVS TRANSITION TIMES.

Phase Shift ZVT Converter Design	transformer 1
Input Supply Voltage, V_{in} , (V)	385
Output Power, P_{out} , (W)	1500
Output Voltage, V_{out} , (V)	36
Output Current, I_{out} , (A)	40
Efficiency (estimate)	0.9
Input Power (estimate), P_{in} , (W)	1666.666667
Primary Current (estimate), I_{pri} , (A)	4.81000481
Switching Frequency, (kHz)	50
Maximum Duty Cycle (estimate)	0.9
Transformer Turn Ratio	6.33
Power Switches	
Mosfet Type	ixfn 340n07
Mosfet Output Capacitance, C_{oss} , (pF)	780
Mosfet $R_{ds(on)}$ @25C (ohms)	0.19
Power Loss Estimates	
Mosfet Total Conduction Losses (W)	18.65008755
Output Rectifier Loss (W)	36.76955262
Magnetic Core Losses (W)	15
Copper Losses (W)	8
Total Power Loss (W)	78.41964017
Full Load Efficiency (%)	95.03176227
Primary ZVT Resonant Tank	
Xfmr Magnetizing Inductance (μH)	2000
Xfmr Leakage Inductance (μH)	27
Xfmr Primary Capacitance (pF)	15
Resonant Tank	
Resonant Capacitance, C_r , (pF)	2095
Resonant Inductance, L_r , (μH) *series	100
Resonant Tank Frequency (mHz)	0.347895125
Resonant Tank Circuit Period, T_{res} , (nS)	2874.429474
Resonant Tank Frequency W_R , (mHz) radians	2.184781383
Tank Transition Time, t_{res} , (μS)	718.6073685
Resonant Tank Impedance, Z_R , (ohms)	218.4781383
Tank Energy	
Capacitive Energy Required, $W(C_r)$, (μC)	155.2656875
Inductive Energy Available, $W(L_r)$, (μC)	155.2656875
Critical Primary Current for ZVT (A)	1.762190044
Critical Secondary Current for ZVT (A)	11.15466298
Minimum Load for ZVT (% of full load)	27.88665744
Transition Time Summary	
Right Leg Transition Time, C/D (nsec)	457.7116997
Left Leg Transition Time, A/B, (nsec)	718.9718497
Primary di/dt Slew Time, (nS)	915.4233993
Total Transition Times, t_{tran} , (nS)	2092.106949
Power Transfer Time, $(1-t_{tran})$, (μS)	17.90789305
Maximum Duty Cycle (actual%)	89.53946526
Delay Resistor C/D	17576.12927
Delay Resistor A/B	27608.51903

APPENDIX D:

ILX BILL OF MATERIALS.

ZVS BOM		CREATED 8/26/09					(purple for >\$5)
PN	DESCRIPTION	QTY	COST	EXT. COST	VENDOR	QUOTED QTY.	
201473	CHASSIS, ZVS BOTTOM	1	\$14.5100	\$14.51	TRU-FAB	15	
201474	CHASSIS, ZVS TOP	1	\$13.0000	\$13.00	TRU-FAB	15	
201475	ZVS REAR PANEL	1	\$10.0000	\$10.00	TRU-FAB	15	
201476	INSULATOR, ZVS CHASSIS BOTTOM	1	\$2.0500	\$2.05	PRESSCUT	100	
201477	INSULATOR, ZVS CHASSIS TOP	1	\$2.0500	\$2.05	PRESSCUT	100	
201478	BRACKET, ZVS OUTPUT BUSS BAR	3	\$10.2100	\$30.63	KEN-RAY	15	
201479	BUSS BAR, ZVS OUTPUT POS	1	\$8.0000	\$8.00		15	
201480	BUSS BAR, ZVS OUTPUT NEG	1	\$6.0000	\$6.00		15	
201481	HEATSINK, TO-247, AAVID 78020	4	\$8.0000	\$32.00	AAVID	32ft MIN ORDER (96PCS)	
				\$0.00			
				\$0.00			
300165	SCR,6-32 X1-1/4" FHP SS	8	\$0.0370	\$0.30	MAX		
300232	SCR,4-40 X 1/4" FHP 100Z	12	\$0.0635	\$0.76	MAX		
300905	WSHR, #8 .375"OD, .093"THK, NYL	8	\$0.0373	\$0.30	MSC	100	
300906	CLIP, TO-247 AAVID MAX03-HG	8	\$0.2900	\$2.32	MOUSER	100	
400024	HDR,10 PIN SINGLE ROW GLD	1	\$0.9100	\$0.91	MAX		
400056	CONN,MALE,3PN,.100,GOLD	1	\$0.2599	\$0.26	MAX		
400201	RES,CF,47K,1/4W,5%	2	\$0.0100	\$0.02	MAX		
400322	CAP,CER,3300PF,50V,COG	1	\$0.4898	\$0.49	MAX		
400336	CAP,POLYP,.33,250V,VDE,X2	3	\$1.0885	\$3.27	MAX		
400350	CAP,ALUM,47UF,50V,RADIAL	1	\$0.0400	\$0.04	MAX		
400352	CAP,ALUM,100UF,50V,RADIAL	3	\$0.0720	\$0.22	MAX		
400410	HDR,2PN .1 CENTER	2	\$0.1710	\$0.34	MAX		
400498	CAP,ALUM, 1500UF,30V,RAD	2	\$0.4079	\$0.82	MAX		
400586	HDR, 10 PIN RIGHT ANGLE	1	\$0.5100	\$0.51	MAX		
400625	RES,PWR,1 OHM TO-220	2	\$1.1900	\$2.38	MAX		
400672	CAP,POLY 1000PF 50V RAD	2	\$0.0481	\$0.10	MAX		
	RES,SMT,1206,5%,0.125W,10K OHM	4	\$0.0090	\$0.04	MAX		
400794	CAP,4700PF,VDE TO GND,Y2	15	\$0.4100	\$6.15	MAX		
400887	HDR, 4 PN SINGLE M	2	\$0.2626	\$0.53	MAX		
400888	RES,SMT,ZERO OHM,.1W,0805	6	\$0.0033	\$0.02	MAX		
400895	RES,SMT,10K OHM, 0805	4	\$0.0189	\$0.08	MAX		
400939	RES,SMT,40.2K,25PPM,0805	1	\$0.1823	\$0.18	MAX		
400946	RES,SMT,20K,1%,0805	2	\$0.0488	\$0.10	MAX		
401014	CAP,SMT,CER,10PF,0805,NPO	11	\$0.0169	\$0.19	MAX		
401016	CAP,SMT,CER,33PF,0805,NPO	1	\$0.0300	\$0.03	MAX		
401018	FERRITE,SMT,120 OHM,1812	2	\$0.2705	\$0.54	MAX		
401063	RES,SMT,33K,.1W,0805,5%	2	\$0.0017	\$0.00	MAX		
401095	HDR, 1X3.1 SPACE, LOCKING	1	\$0.0700	\$0.07	MAX		
401100	PIN,TESTPOINT, SMT	2	\$0.1469	\$0.29	MAX		
401107	RES,SMT,0 OHM,0805,1%	3	\$0.0058	\$0.02	MAX		
401140	RES,SMT,1.00K,0805,1%	5	\$0.0037	\$0.02	MAX		
401144	RES,SMT,10.0 OHM,0805,1%	5	\$0.0281	\$0.14	MAX		
401148	RES,SMT,100K,0805,1%	2	\$0.0025	\$0.01	MAX		
401149	RES,SMT,110K,0805,1%	2	\$0.0279	\$0.06	MAX		
401157	RES,SMT,20.0K,0805,1%	1	\$0.0031	\$0.00	MAX		
401166	RES,SMT,4.99K,0805,1%	1	\$0.0026	\$0.00	MAX		
401183	CAP,SMT,CER,.1UF,50V	5	\$0.0078	\$0.04	MAX		
401197	RES,SMT,1.91K,0805,1%	1	\$0.0900	\$0.09	MAX		
401198	RES,SMT,8.06K,0805,1%	1	\$0.0900	\$0.09	MAX		

401199	RES,SMT,681K,0805,1%	1	\$0.0300	\$0.03	MAX	
401205	RES,SMT,3.48K,0805,1%	1	\$0.0300	\$0.03	MAX	
401300	RES,SMT,7.87K,1%,.1W,0805	1	\$0.0025	\$0.00	MAX	
401337	RES,SMT,10,.5%,1/8W,1206	5	\$0.0025	\$0.01	MAX	
401338	RES,SMT,270,5%,.1W,0805	1	\$0.0018	\$0.00	MAX	
401352	RES,SMT,150,5%,.1W,0805	1	\$0.0038	\$0.00	MAX	
401362	CAP,SMT,CER,2200P,0805	10	\$0.0054	\$0.05	MAX	
401387	RES,SMT,100K,5%,.1W,0805	1	\$0.0017	\$0.00	MAX	
401399	RES,SMT,10K,5%,.1W,0805	5	\$0.0017	\$0.01	MAX	
401417	RES,SMT,6.65K,1%,1W,0805	1	\$0.0344	\$0.03	MAX	
401466	CAP,SMT,2.2UF16V,CER,0805	2	\$0.1046	\$0.21	MAX	
401481	CAP,SMT,TANT,1UF,16V 3216	1	\$0.0100	\$0.01	MAX	
401508	CAP,SMT,10000PF,50V,0805	7	\$0.0292	\$0.20	MAX	
401539	CAP,SMT,1UF,25V,10%	4	\$0.8094	\$3.24	MAX	
401554	CAP,SMT,1NF,50V,5%,0805	3	\$0.0405	\$0.12	MAX	
401643	RES,SMT,22K,5%,0805	5	\$0.0473	\$0.24	MAX	
401653	RES,SMT,49.9K,1%,0805	1	\$0.0105	\$0.01	MAX	
401654	RES,SMT,33.2K,1%,0805	1	\$0.0272	\$0.03	MAX	
401669	RES,SMT,2.2K,5%,0805	1	\$0.0200	\$0.02	MAX	
401736	RES,SMT,169K,1%,.1W,0805	1	\$0.0105	\$0.01	MAX	
401737	RES,SMT,6.49K,1%,.1W,0805	1	\$0.0900	\$0.09	MAX	
401871	RES,SMT,24.9K,1%,0805	1	\$0.0104	\$0.01	MAX	
401895	RES,SMT,5 OHM 1% .1W 0805	2	\$0.0200	\$0.04	MAX	
402000	HDR, 8 PIN M-M SINGLE	1	\$2.2464	\$2.25	MAX	
402008	CAP,SMT,CER,.47UF,50V,10%,1206	1	\$0.0844	\$0.08	MAX	
402011	CAP,560PF,50V,0805	1	\$0.0540	\$0.05	MAX	
402012	CAP,SMT ALUM 10UF 35V	14	\$0.1322	\$1.85	MAX	
402021	RES, 51.1,1/8W,1%,0805	2	\$0.0247	\$0.05	MAX	
402023	RES,SMT 51.0 OHM 1% 1206	2	\$0.0273	\$0.05	MAX	
402049	CAP,SMT,CER,.22uF,25V	1	\$0.1789	\$0.18	MAX	
402072	RES,SMT,2K,1%,0805	1	\$0.0288	\$0.03	MAX	
402110	CONTACT, SOLDER SOCKET 20A	1	\$2.8835	\$2.88	MAX	
402151	CONN HEADER 2POS 4.2MM	2	\$0.9072	\$1.81	MAX	
402208	RES, SMT, 430K, 1%, 0805	1	\$0.0380	\$0.04	MAX	
402221	CAP, .22uF, 25V, 1206	1	\$0.1070	\$0.11	MAX	
402224	RES, 22, 1/10W, 1%, 0805	1	\$0.1000	\$0.10		
402253	CAP, SMT, .1UF, 100V, 0805	20	\$0.1755	\$3.51	MAX	
402373	RES, SMT, 80.6K, 1%, 1/8W, 0805	1	\$0.0200	\$0.02	MAX	
402378	PCB, ZVS36V40A	1	\$15.0000	\$15.00		
402468	IND, THD, HIPWR, 150uH	1	\$26.9600	\$26.96	WHITESHEET	1
402469	IND, TOROID, CM, 100uH	2	\$4.5000	\$9.00	WHITESHEET	1
402470	XFRMR, THD, 200:1	3	\$2.4400	\$7.32	WHITESHEET	
402471	RES, SMT, 909K, 1%, 2010	4	\$0.0900	\$0.36	WHITESHEET	100
402472	CAP, SMT, 470PF, 0805	3	\$0.1100	\$0.33	WHITESHEET	10
402473	CONN, SCREW TERM, 25A, 4POS	1	\$1.7600	\$1.76	WHITESHEET	25
402474	RELAY, NO, 30A, 12V COIL	1	\$1.7600	\$1.76	WHITESHEET	1
402475	RES, WW, 5W, 40 Ohm	0	\$0.7400	\$0.00	WHITESHEET	1
402478	CAP, RAD, 330uF, 420V	3	\$7.4500	\$22.35	WHITESHEET	1
402481	CONN, FUSE CLIP 1/4", PCB MOUNT	4	\$0.4800	\$1.92	WHITESHEET	80
402482	RELAY, SPST-NO, 30A, 12V	0		\$0.00		
402488	PCB, ZVS LOAD BOARD	1	\$15.0000	\$15.00		
402489	CAP, SMT, CER, 0.012UF, 805	0	\$0.1100	\$0.00	WHITESHEET	10
402490	CAP, SMT, CER, 330PF 10%, 50V, 805	0	\$0.0580	\$0.00	WHITESHEET	10
402491	CAP, CER, DISK, 2700PF, 1000V	1	\$0.3440	\$0.34	WHITESHEET	10
402492	CAP, ALUM, 470UF, 450VDC	0	\$6.5870	\$0.00	WHITESHEET	10
402493	XFRMR, CURRENT SENSE, 1:200, 25A	0		\$0.00		
402494	IND, CHOKE, HICURRENT, 200nH	0	\$2.3640	\$0.00	WHITESHEET	10
402495	XFRMR, SM Gate Driver, 1:1:1, 3300UH	2	\$3.2300	\$6.46	WHITESHEET	10
402496	RES, SMT, 8.87K, 1%, 0805	0	\$0.0380	\$0.00	WHITESHEET	10
402497	POT, TRIM, 2K, SIDE ADJ	1	\$0.7500	\$0.75	WHITESHEET	10

402498	FLTR, CHOKE, COMMON MODE, 75A	0	\$1.4960	\$0.00	WHITESHEET	10
402499	CAP, ALUM, 1000uF, 50V, Elect, Radial	8	\$0.4000	\$3.20		
402500	RES, SMT, 40.0K Ohms, 0805	0	\$0.4000	\$0.00	WHITESHEET	10
402501	XFRMR, ZVS, 1500VA, 6.69:1	1	\$27.1000	\$27.10	WHITESHEET	
402502	RES, SMT, 25.5K, 0805	0	\$0.0740	\$0.00	WHITESHEET	10
402503	RES, WW Non-Ind, 25 Ohms, 0.5W	0	\$0.7876	\$0.00	WHITESHEET	25
402504	CAP, Radial, 10UF, 630V, Metal Poly	2	\$7.7300	\$15.46	WHITESHEET	25
402505	IND, SMT, Power, 3.3UH, 42A	1	\$12.6800	\$12.68	WHITESHEET	10
402506	XFRMR, SM with bias, 8.375:1:1:1	0		\$0.00		
402513	FAN, 60MMX25MM, 28.5CFM, 3.24W	2	\$5.3780	\$10.76	WHITESHEET	10
				\$0.00		
				\$0.00		
500002	DIO, 1N4148, SIGNAL DIODE	1	\$0.0064	\$0.01	MAX	
500006	DIO, 1N4005, 1A, 500V	1	\$0.0200	\$0.02	MAX	
500193	DIO, 1N4740A, 10V 1W, ZEN	1	\$0.0510	\$0.05	MAX	
500301	DIO, SMT 1N4148 SOT23	10	\$0.0222	\$0.22	MAX	
500317	DIODE, SCHOTTKY SMT	2	\$0.0500	\$0.10	MAX	
500341	IC, SMT, NPN TRANSISTOR	1	\$0.0150	\$0.02	MAX	
500396	ICD, SMT, OTP, BJT NPN	3	\$0.5158	\$1.55	MAX	
500486	DIODE, 5A SCHOTTKY SMT	2	\$0.5238	\$1.05	MAX	
500504	DIO, SMT, SCHOTTKY 2A 20V	1	\$0.0898	\$0.09	MAX	
500515	IC, SMT, MOSFET P-CHAN POWR	1	\$0.2987	\$0.30	MAX	
500546	IC, SMT, LED RED/GREEN	2	\$0.3969	\$0.79	MAX	
500720	DIO, SMT, BAT54, SCHOTTKY	2	\$0.0400	\$0.08	MAX	
500816	TRAN, SMT, NPN, POWER, DPAK	1	\$0.3033	\$0.30	MAX	
500817	TRAN, SMT, PNP, POWER, DPAK	4	\$0.3538	\$1.42	MAX	
500850	FET, SMT, N-CH, 2.5V, 3A	2	\$0.3800	\$0.76	MAX	
500987	LED, GRN, SMT	1	\$0.1000	\$0.10	MAX	
501006	DIODE, 6.2V, 5MA, 0.5W, ZENER, SM	2	\$0.1200	\$0.24	MAX	
501011	DIODE, TO247, 600V, 60A	1	\$5.5400	\$5.54	WHITESHEET	1
501012	TRANS, NFET, 500V, 45A	0	\$12.6500	\$0.00	WHITESHEET	10
501013	RECT, BRIDGE, GBJ, 400V, 25A	1	\$1.5600	\$1.56	WHITESHEET	10
501014	IC, PFC, UC3854BDW	1	\$2.7000	\$2.70	WHITESHEET	25
501015	NFET, DPAK, 800V, 1A	1	\$0.9700	\$0.97	WHITESHEET	1
501018	DIODE, SCHOTTKY, DUAL, 150V, 30A	2	\$6.5000	\$13.00	WHITESHEET	20
501019	IC, GATE DRIVE, DUAL, 2A, SMT	2	\$0.8252	\$1.65	WHITESHEET	20
501020	IC, OPAMP, DIFFAMP, SMT	2	\$1.7928	\$3.59	WHITESHEET	20
501021	IC, OPAMP, CURRENT SENSE, LT6105	1	\$1.3944	\$1.39	WHITESHEET	25
501022	IC, OPAMP, GP Rail-Rail Dual, OPA2251	1	\$3.7800	\$3.78	WHITESHEET	25
501023	TRANS, NCH, 650V 20.7A, 20.8W	8	\$3.1500	\$25.20	WHITESHEET	100
501024	IC, SWITCHER, TNY267	1	\$1.0300	\$1.03	WHITESHEET	1-100
501025	VREG, SMT, PWM, QUAD, 5V OUT	1	\$8.0252	\$8.03	WHITESHEET	25
501026	DIO, TRANSORB, 1500W, 456V BKDN	1	\$0.4600	\$0.46	WHITESHEET	10
				\$0.00		
	COSEL CONNECTOR	1	\$0.3630	\$0.36		10
	ICL7660S-CBA	1	\$1.2400	\$1.24		25
	LM7301-NOPB	1	\$1.8500	\$1.85		25
				\$0.00		
				\$0.00		
				\$0.00		
				\$0.00		

TOTAL

\$415.31

GOAL: <\$375

TOTAL FOR PARTS OVER \$5

\$344.17

82.87%

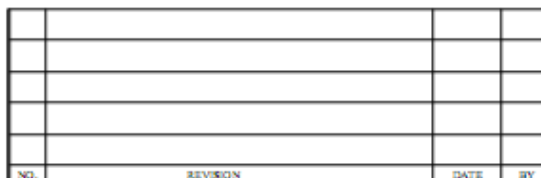
PRICE PER WATT

\$0.28

APPENDIX E:

DOE TRANSFORMER DATA SHEET.

TRANSFORMER OUTLINE & SCHEMATIC



CUSTOMER:		CUST. P/N	
Mont. St. Univ.		--	
DRAWN BY:	DATE:	AM P/N	
AMB	8/07/08	AM-19488	

APPENDIX F:

DOE ZVS TRANSITION TIMES.

Phase Shift ZVT Converter Design	transformer 1
Input Supply Voltage, Vin, (V)	29
Output Power, Pout, (W)	1662.5
Output Voltage, Vout, (V)	133
Output Current, Iout, (A)	12.5
Efficiency (estimate)	0.95
Input Power (estimate), Pin, (W)	1750
Primary Current (estimate), Ipri, (A)	63.52087114
Switching Frequency, (kHz)	50
Maximum Duty Cycle (estimate)	0.95
Transformer Turn Ratio	0.167
Power Switches	
Mosfet Type	ixfn 340n07
Mosfet Output Capacitance, Coss, (pF)	7100
Mosfet Rds(on) @25C (ohms)	0.004
Power Loss Estimates	
Mosfet Total Conduction Losses (W)	68.47454181
Output Rectifier Loss (W)	11.49048519
Magnetic Core Losses (W)	15
Copper Losses (W)	8
Total Power Loss (W)	102.965027
Full Load Efficiency (%)	94.16782403
Primary ZVT Resonant Tank	
Xfmr Magnetizing Inductance (uH)	7500
Xfmr Leakage Inductance (uH)	7
Xfmr Primary Capacitance (pF)	83000
Resonant Tank	
Resonant Capacitance, Cr, (pF)	101933.3333
Resonant Inductance, Lr, (uH) *series	1
Resonant Tank Frequency (mHz)	0.498749236
Resonant Tank Circuit Period, Tres, (nS)	2005.015604
Resonant Tank Frequency WR, (mHz) radians	3.1321452
Tank Transition Time, tres, (uS)	501.2539011
Resonant Tank Impedance, ZR, (ohms)	3.1321452
Tank Energy	
Capacitive Energy Required, W(Cr), (uC)	42.86296667
Inductive Energy Available, W(Lr), (uC)	42.86296667
Critical Primary Current for ZVT (A)	9.25883002
Critical Secondary Current for ZVT (A)	1.546224613
Minimum Load for ZVT (% of full load)	12.36979691
Transition Time Summary	
Right Leg Transition Time, C/D (nsec)	319.2700007
Left Leg Transition Time, A/B, (nsec)	501.5081376
Primary di/dt Slew Time, (nS)	638.5400014
Total Transition Times, ttran, (nS)	1459.31814
Power Transfer Time, (1-ttran), (uS)	18.54068186
Maximum Duty Cycle (actual%)	92.7034093