

# **Modeling and Control of a Single-Phase, 10 kW Fuel Cell Inverter**

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in

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by  
Troy A. Nergaard  
Electrical Engineering

## **Abstract**

As the world's energy use continues to grow, the development of clean distributed generation becomes increasingly important. Fuel cells are an environmentally friendly renewable energy source that can be used in a wide range of applications and are ideal for distributed power applications. In this study, the power conversion element of a dual single-phase, 10 kW stand-alone fuel cell system is analyzed. The modular converter consists of a DC-DC front-end cascaded with a half-bridge inverter. The entire system is accurately modeled, to help determine any interactions that may arise. Control strategies based on simplicity, performance, and cost are evaluated. A simple voltage loop, with careful consideration to avoid transformer saturation, is employed for the phase-shifted DC-DC converter. Several experimental transfer functions were measured to confirm the modeling assumptions and verify the control design of the DC-DC converter. Two control options for the inverter are explored in detail, and experimental results confirm that the modulation index must be controlled to regulate the output voltage during various load conditions. The final system is implemented without the use of current sensors, thus keeping the inverter cost down. Experimental results using a power supply are given for resistive, inductive, and nonlinear loads and the performance is acceptable. Fuel cell test results, including transient response, are also displayed and analyzed.

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# Chapter 1 – Introduction

## **1.1 Motivation**

Have you ever heard of the phrase “sustainable development?” Sustainable development means to meet the needs of the present without compromising the ability of future generations to meet their own needs [1]. With energy use rising every year, oil production growing at alarming rates, and greenhouse gas emissions on the rise, unfortunately the world is a lot closer to the category of “unsustainable development.” As awareness grows and technology advances, the opportunity to move closer to the sustainable path becomes much easier. Alternative, renewable energy sources have been around for quite some time; however, for numerous reasons, they have not yet emerged as primary power sources. The recent revitalization of the fuel cell hopes to help discontinue this trend.

The introduction of fuel cell systems into the power generation market will not only supply clean renewable energy to millions of users, but it will help reduce the dependence on oil, which is of critical importance, especially to countries like the United States. The United States imports over 50% of its total oil consumption, which amounts to about 10.7 million barrels a day. [2]. Although fossil fuel reserves are still large, they are finite and the world’s crude oil production is projected to peak sometime in the early 21<sup>st</sup> century [1].

Furthermore, fuel cells are ideal for distributed power generation applications. The clean, quiet, and efficient nature of fuel cell systems would make them desirable for remote locations and in developing countries where electricity infrastructures do not exist and transmission is poor [3]. Fuel cell power generation technology has economical, political, and environmental advantages.

## **1.2 Application**

There are many promising applications for fuel cell systems including portable power, transportation, and distributed power generation [4]. The output power of these systems can range from a few watts to several hundred kilowatts. The area of interest here is distributed power generation, which could involve standby power for commercial customers, regional systems supplying neighborhoods or individual units that may or may

not be grid inter-tied. Approximately 40% of the total electric power consumed in the United States is due to the residential market [5]. Specifically targeted in the following analysis is stand-alone systems intended for household applications. In a stand-alone system, the user is isolated from the utility grid either by choice or by circumstance.

The average energy consumption of a household is influenced by many factors including climate, season, size of house, and size of family. Studies have been completed [6-8] and a typical 24-hour load profile for a medium size household is shown in Figure 1.1. This profile shows a peak of 6.6 kW and a daily average of about 3 kW. To accommodate for motor starts, such as air conditioning, and short durations of unusually heavy loads, such as hair dryers and microwaves, it is necessary to have a fuel cell system that is capable of around 10 kW.

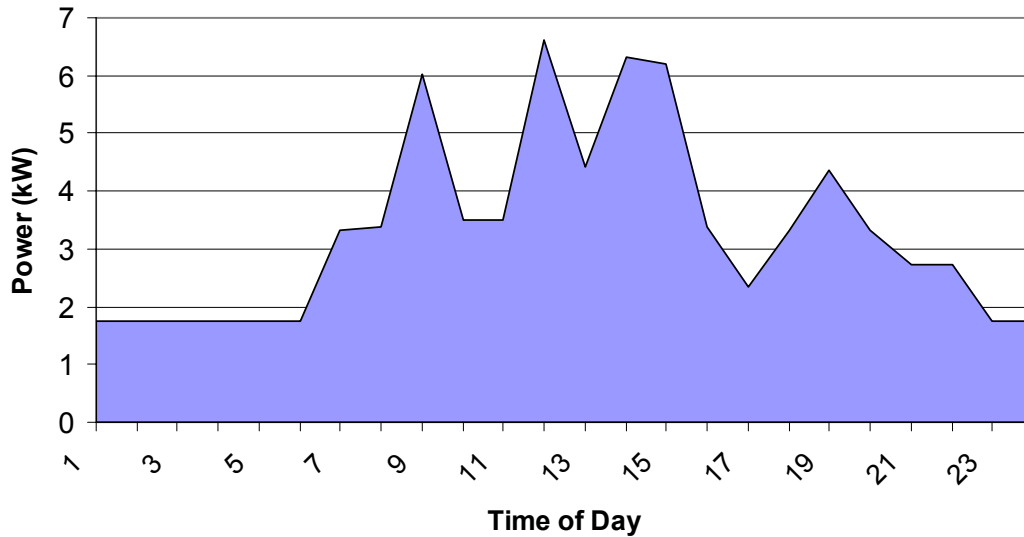


Figure 1.1 Average Daily Household Energy Use

### 1.3 Objective and Outline

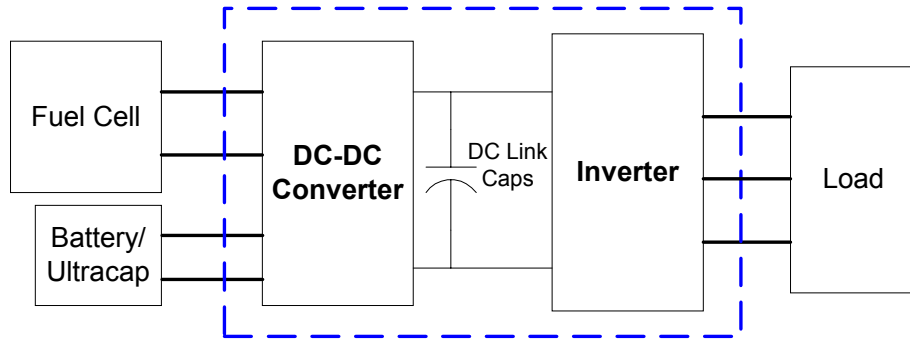
Fuel cell systems often consist of three main stages: the fuel processing, the chemical to electrical conversion, and the power processing. The power processing, which is typically made up of power electronics, is the focus of this work. The power converter specifications for the prototype system, originally given by the 2001 Future Energy Challenge (FEC), require a nominal 48 V<sub>dc</sub> input from the fuel cell and two 60 Hz, single-phase 120 V<sub>ac</sub> outputs and one 240 V<sub>ac</sub> output from the inverter [9]. The load

can range from 0 to 10 kW and will include nonlinear and non-unity power factor conditions. Furthermore, for fuel cell systems to be competitive with conventional power generation, a large emphasis must be placed on reducing the cost of the inverter. The complete specifications are given in Table 1.1 and follow standards such as IEEE-519 [10] and guidelines from the Information Technology Industry (ITI) council (formerly CBEMA). The ITI/CBEMA curve, which describes the AC input voltage envelope that most information technology equipment can tolerate, is shown in Appendix A. It will serve as the main standard for the transient analysis of the inverter.

**Table 1.1 Design Specifications**

| Parameter                 | Target Requirement                                                                                                                                                           |
|---------------------------|------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|
| Output Power Capability   | 10 kW continuous, Single-phase 120/240 V, 60 Hz output suitable for domestic applications                                                                                    |
| Input Source              | 48 V DC nominal source (tolerance range 42 V- 72 V) with slow transient characteristics                                                                                      |
| Manufacturing Cost        | \$500 maximum when scaled to a 10kW design in high volume production                                                                                                         |
| Package Size              | Volume less than 50L                                                                                                                                                         |
| Package Weight            | Mass less then 32 kg, not including energy sources or batteries                                                                                                              |
| Overall Efficiency        | Higher then 90% for 10 kW resistive load                                                                                                                                     |
| Total Harmonic Distortion | Output voltage: less than 5% when supplying a standard nonlinear test load                                                                                                   |
| Safety                    | Intended for safe, routine use in a home or small business by non-technical customers.                                                                                       |
| Voltage Regulation        | Output voltage tolerance no wider than $\pm 6\%$ over the entire line voltage and temperature range, from no-load to full-load                                               |
| Frequency Regulation      | 60 $\pm$ 0.1 Hz                                                                                                                                                              |
| Acoustic Noise            | No louder than conventional domestic refrigerator. Less than 50 dBA sound level measured 1.5 m from the unit.                                                                |
| Electrical Noise          | Able to meet FCC Class A--industrial requirements for conducted and radiated EMI.                                                                                            |
| Protection                | Self-protection against output short circuit, over current, over temperature, over voltage, and under voltage or loss of input source with no damage caused by any of these. |
| Environment               | Suitable for indoor installation in domestic applications, 10°C to 40°C possible ambient range.                                                                              |
| Lifetime                  | The system should function for at least ten years with routine maintenance when subjected to normal use in a 20°C to 30°C ambient environment.                               |

For economical reasons, the fuel cell stack produces a nominal voltage of 48 V; therefore, the input voltage must be boosted with a dc-dc converter before it can be inverted. Also required is some type of energy storage to supplement the fuel cell during start-up and load transients. The basic structure of the system is shown with the block diagram in Figure 1.2, with the power electronics highlighted in the blue dotted box.



**Figure 1.2 Block Diagram of Fuel Cell System**

As part of the Future Energy Challenge, an initial prototype of the power stage was designed, built, and partially tested. This inverter, although modified from the original, is the foundation for the work done throughout this thesis. The basic objective is to develop an accurate model and cost effective control scheme for the system. This involves studying the interactions of the fuel cell and converters and analyzing the performance of the inverter using two simple control methods. First, Chapter 2 will give an overview of the design options and some more details on the chosen inverter design, including information on the ultracapacitors used. Next, Chapter 3 will develop the models for each stage of the system and analyze the different control options. This will include the design of a closed loop DC-DC full-bridge converter, which uses only voltage feedback for control. Theoretical and experimental transfer functions are compared for the DC-DC converter. The half-bridge inverter is designed to run open loop and closed loop, and the performance is compared in simulation and experiment. Chapter 4 will provide experimental results over a wide range of loads, using a power supply and a fuel cell as the input source. Finally, a summary of the control design and corresponding results are presented and future work is considered in Chapter 5.

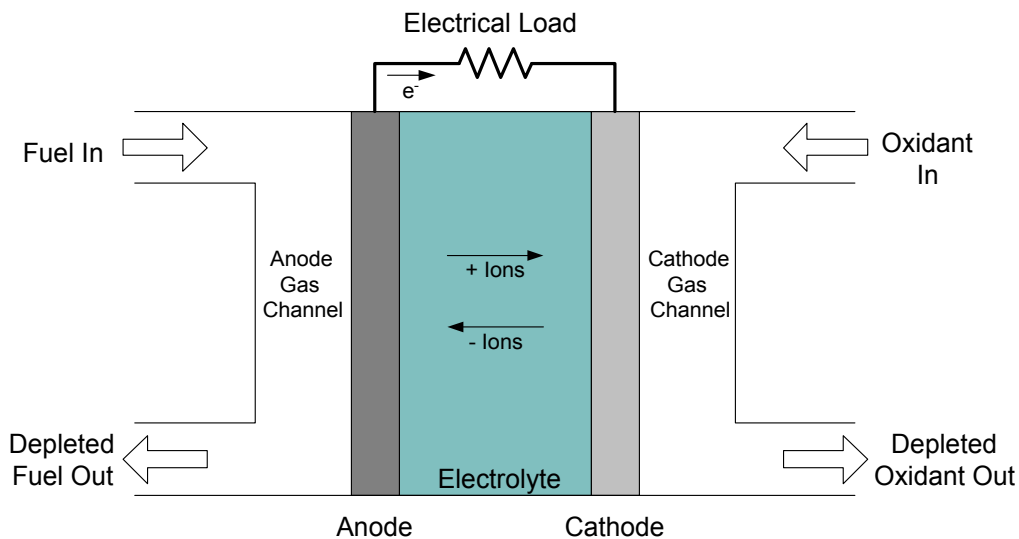
## **1.4 Background Information**

The following section provides a backdrop for remainder of the thesis. Information, including literature reviews, on the basic operation of fuel cells and on inverter technology is provided.

### 1.4.1 Fuel Cells

The history of fuel cells dates back to 1839 when Sir William Grove, a British Scientist, discovered the technology. However, it was not until the mid 1900's when fuel cells began to make a name for themselves in the space industry [1]. Shortly after that, several private companies became interested in fuel cell technology, but the economic and technological barriers were difficult to overcome. Recently, the fuel cell has been revived and shows tremendous promise in the transportation and utility sectors.

Fuel cells are electrochemical devices that convert chemical energy, typically from hydrogen, directly into electrical energy. Similar to a battery, a fuel cell consists of two electrodes (anode and cathode) and an electrolyte. The fuel is supplied to the anode where it is oxidized and produces positive ions that flow through the electrolyte. The electrons are directed outside creating current through the external circuit and then back to the cathode. At the cathode, oxygen (air) is supplied and reduced, consuming the electrons from the external circuit, and the only resultant byproduct is  $H_2O$  (water). More details on the chemical and electrical aspects of fuel cells can be found in [4] and [11]. A basic schematic of a fuel cell is shown in Figure 1.3.



**Figure 1.3 Individual Fuel Cell Schematic**

There are several different types of fuel cells, most often categorized by the type of electrolyte present. Four of the more common fuel cells are proton exchange membrane fuel cells (PEMFC), phosphoric acid fuel cells (PAFC), molten carbonate fuel cells

(MCFC), and solid oxide fuel cells (SOFC). The characteristics of these four are summarized in Table 1.2. The PEMFC is probably the most well known fuel cell and shows promise for applications in the medium power range. Because of its efficiency and relatively low operating temperature range, the PEMFC is ideal for residential applications and is the chosen fuel cell for the system under study.

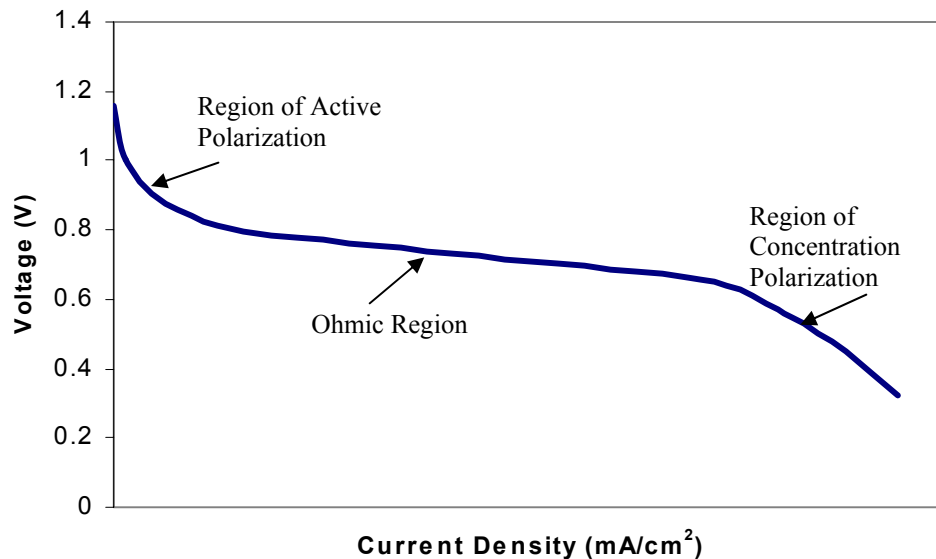
**Table 1.2 Characteristics of Types of Fuel Cells**

|                       | PEMFC                     | PAFC                            | MCFC                            | SOFC                              |
|-----------------------|---------------------------|---------------------------------|---------------------------------|-----------------------------------|
| Electrolyte           | Polymer Membrane          | Phosphoric Acid                 | Molten Mixture                  | Ceramic                           |
| Operating Temperature | $\leq 80^{\circ}\text{C}$ | $205^{\circ}\text{C}$           | $650^{\circ}\text{C}$           | $800\text{-}1000^{\circ}\text{C}$ |
| Power Density         | $> 0.6 \text{ W/cm}^2$    | $0.2\text{-}0.3 \text{ W/cm}^2$ | $0.1\text{-}0.2 \text{ W/cm}^2$ | $0.25\text{-}0.35 \text{ W/cm}^2$ |
| Charge Carrier        | $\text{H}^+$              | $\text{H}^+$                    | $\text{CO}_3^-$                 | $\text{O}^-$                      |
| Catalyst              | Platinum                  | Platinum                        | Nickel                          | Perovskites                       |

Each individual fuel cell produces a very small voltage; the theoretical potential of a cell at  $80^{\circ}\text{C}$ , 1 atm of pressure and under no load is  $V_0 = 1.16 \text{ V}$  [5]. As the current density is increased, the fuel cell output voltage drops. The “voltage” efficiency of a fuel cell is given in (1.1), where  $V$  is the actual terminal voltage.

$$\eta = \frac{V \cdot I}{V_0 \cdot I} = \frac{V}{V_0} \quad (1.1)$$

A typical polarization curve for an individual fuel cell is shown in Figure 1.4. The V-I curve shows three main regions, but the ohmic region is the normal operating area of the fuel cell. Since the individual cell voltage in this region is only about 0.7 V, they are often “stacked” together in series to produce a more useful voltage level. However, a large number of series cells can lead to reduced reliability since if one fails, the entire stacks stops functioning.



**Figure 1.4 Single Cell Polarization Curve**

There are numerous advantages with the use of fuel cells for power generation. The biggest advantage, perhaps, is the environmental impact. Pure hydrogen fuel cells produce no harmful emissions, and even fuel cells that require a reformer produce significantly less emissions than their internal combustion generator counterparts. The efficiency of a fuel cell is another advantage over internal combustion engines, which are limited by the Carnot cycle efficiency. Due to the direct conversion from chemical to electrical energy, the efficiency of a fuel cell is in the 40-50% range. Most of that power lost is in the form of heat, which can be captured relatively easily and put to use. Co-generation is a big advantage with PEMFCs and the residential market. The heat produced from the fuel cell stack can be used to partially heat the home and the water. Lastly, fuel cells systems are very quiet and can fit nicely in a basement or a garage.

Of course there are a few drawbacks, most of which will get better with time and exposure. The main disadvantages are cost and availability; both of which go hand-in-hand. Fuel cells have already decreased in price from just a few years ago, but they still have a long way to go in order to compete with the price of standard grid power. Not only are the stacks costly, but also are the sub systems including the power electronics. The other main drawback is the lack of a hydrogen infrastructure. Although hydrogen is virtually everywhere, it is not easily refined, stored, or available. Electrolysis using solar power is one very good option, and as alluded to before, hydrogen can be reformed from

almost any hydrocarbon fuel including methanol, propane, and natural gas. Just like any emerging technology, it takes time to work out the kinks; however, it is clear that fuel cell systems have the potential to positively impact the power generation market in years to come.

### 1.4.2 Inverters

Inverters that take DC and produce a constant amplitude sinusoidal output have been studied and designed for many years. Initially, most inverter technology used silicon-controlled rectifier (SCR) devices and a transformer coupling to approximate a sine wave via line commutation [13]. As power transistors became more feasible, most low to medium power inverter systems replaced the SCR with the MOSFET or the IGBT. These new transistors lead the way for force-commutated inverters that can be classified in terms of their output waveform. A summary of the basic types of force-commutated inverters is shown in Table 1.3 [14]. Significant research and development in the area of pulse width modulation (PWM) has been done in attempt to reduce the passive filter size and create a better sinusoidal output, thus reducing harmonics. Today the PWM inverter is still very popular, but there has also been much work on other modulation schemes such as hysteresis control and space vector modulation (SVM).

**Table 1.3 Types of Force-Commutated Inverters**

| Classification        | Description                                                                                                                                                           |
|-----------------------|-----------------------------------------------------------------------------------------------------------------------------------------------------------------------|
| Square Wave           | The output is alternately positive or negative (only two states) and is directly related to the dc input voltage.                                                     |
| Quasi Square Wave     | The output goes from positive to zero to negative and back to zero (three states) and the average output can be controlled by varying the duration of the zero state. |
| Pulse Width Modulated | The output consists of trains of high frequency pulses of varying width that alternate between positive and negative voltage.                                         |
| Step Wave             | The output consists of the sum or the difference of varying width pulses and the unfiltered output is close to the desired sinusoid.                                  |

Inverters can be found in a variety of forms, including full bridge or half bridge, single phase or three phase, current source (CSI) or voltage source (VSI), and two-level or multilevel. Historically, more emphasis has been placed on the design of three phase inverters, especially for applications like motor drives, active filters, and uninterruptible power supplies (UPS). These same applications can be applied to single-phase inverters as demonstrated in [15-17]. Single-phase inverters that require a DC-DC front-end,

similar to the one described in this thesis, have also been developed for photovoltaic energy conversion systems [18], but have not been analyzed in detail.

Control techniques for inverters are also quite abundant and include options such as analog versus digital, open loop versus closed loop, and stationary reference versus rotating reference. The control design for single-phase inverters can actually be more complicated than that for three phase inverters, because it is difficult to apply coordinate transformations to single-phase systems. Stationary frame controllers with AC systems are subject to steady-state errors, because of the gain limitations at the fundamental frequency. Several control methods for single-phase inverters are summarized in [19] and [20]. The methods discussed range from deadbeat to state feedback and some involve sensing both current and voltage. In [21], a synchronous frame controller, utilizing capacitor current feedback, is presented for a single-phase inverter. Although this method does allow for a conventional DC control method to be used, it requires complex computations and sensing of the filter capacitor current. An interesting alternative to the synchronous frame controller is to transform the compensator rather than transform the feedback signals [22]. There are numerous control schemes, varying in cost and complexity, available for single-phase inverters. The best method ultimately depends on the application and the resources available.

## Chapter 2 – Design Overview

### 2.1 Design Options

There are several possible topologies and energy storage options for a 48 V to 120 V inverter. Throughout this section, several DC-DC converter and inverter topologies are presented and compared based on their performance and cost. Energy storage options are also compared and contrasted. More information and design details can be found in [23] and [24].

#### 2.1.1 Topology

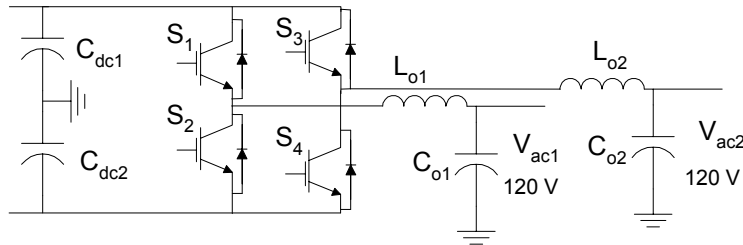
The simplest topology involves a low voltage inverter and then a 60 Hz transformer to step up the voltage. This option has the least number of components yet still provides isolation. However, a 10 kW, 60 Hz transformer is physically large and heavy, and will not meet the size or weight specifications given. The cost of a 60 Hz transformer can also be quite significant.

Assuming the 60 Hz transformer option is unreasonable, the overall topology of the system requires a front-end DC-DC converter to boost the fuel cell voltage and then an inverter to create the 120 V AC output. Since isolation is desirable, there are four main front-end topologies to consider and they are listed in Table 2.1. Neither the forward nor the push-pull circuits work well for higher power applications. The single switch forward converter is common in the 30 W to 100 W power range, but for low voltage device applications, the duty cycle is limited to less than 50% and an additional winding is required to reset the transformer when the device is off. The push-pull converter is a great option for low power systems (500 W to 1.5 kW). Its major problem is the center tap termination, which tends to cause saturation of the transformer at high power levels due to slightly unbalanced excitation. The half-bridge topology requires a split capacitor bus, twice the device current, and twice the transformer turn ratio. The full-bridge topology is a good option and can be easily scaled for higher power levels. Although the full-bridge requires the most switches, the switch's voltage stress can be less than 100 V; therefore, the low voltage power MOSFET, a more economical and efficient device, can be utilized.

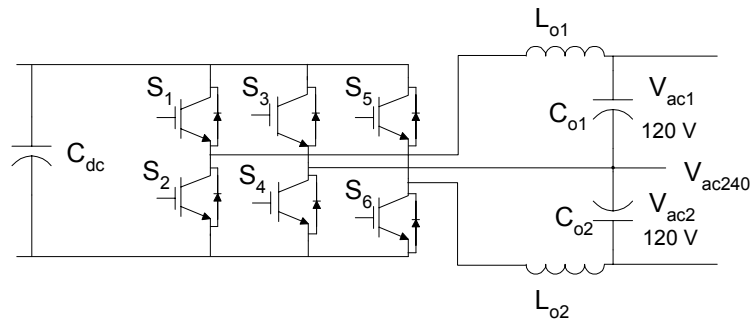
**Table 2.1 DC-DC Converter Options**

| Topology    | # of Switches | Device Voltage | Advantages                                                                     | Disadvantages                                                                                                             |
|-------------|---------------|----------------|--------------------------------------------------------------------------------|---------------------------------------------------------------------------------------------------------------------------|
| Forward     | 1             | $\geq 200$     | <ul style="list-style-type: none"> <li>• Minimum component count</li> </ul>    | <ul style="list-style-type: none"> <li>• Requires lossy resetting</li> <li>• Limited duty cycle</li> </ul>                |
| Push-Pull   | 2             | 200            | <ul style="list-style-type: none"> <li>• Low component count</li> </ul>        | <ul style="list-style-type: none"> <li>• Center tap is a problem at high power</li> <li>• Twice device voltage</li> </ul> |
| Half-Bridge | 2             | 100            | <ul style="list-style-type: none"> <li>• Low component count</li> </ul>        | <ul style="list-style-type: none"> <li>• Requires twice device current</li> </ul>                                         |
| Full-Bridge | 4             | 100            | <ul style="list-style-type: none"> <li>• Soft switching is possible</li> </ul> | <ul style="list-style-type: none"> <li>• High component count</li> </ul>                                                  |

From a cost and performance standpoint, it is unclear which inverter topology is the best. The single bus with two half-bridge inverters, as shown in Figure 2.1, seems simple and has a low component count. However, the single bus topology may result in the need for some very large capacitors due to large, low frequency neutral currents present in an unbalanced load situation. The six-switch, single bus topology in Figure 2.2 allows complementary SPWM switching and thus minimizes passive components, including the AC filter and the DC bus capacitors; however, it also introduces significantly more control effort into the system [25].

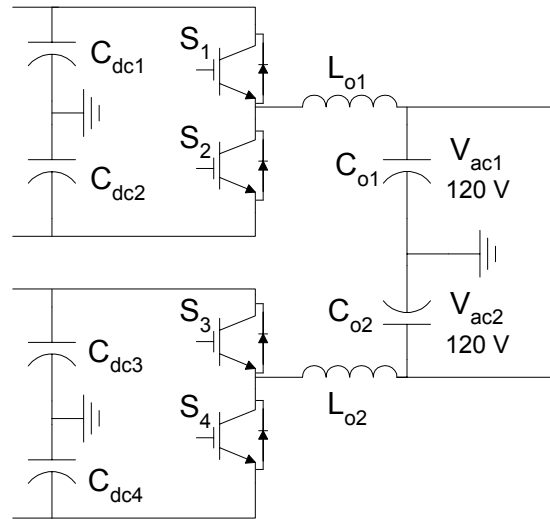


**Figure 2.1 Single DC Bus with Split Capacitor**



**Figure 2.2 Single Phase Three-wire**

If a modular design, as shown in Figure 2.3, consisting of a dual dc bus configuration is used the component count will increase. On the other hand, the circuit becomes more flexible and introduces some redundancy into the system. This can be very important if it is to be considered as a sole source of power for a home. It also makes the rectifier stage much easier to design, because of the reduced current through the diodes and the DC filter inductor.



**Figure 2.3 Dual DC Bus**

Table 2.2 shows a summary of the different inverter options, including the number of semiconductor switches needed. Each option has certain advantages and disadvantages, so the best option depends on the goals of the project and application of the system. It should be noted that for the specified voltage level, the most economical semiconductor device for the inverter is the 600 V IGBT.

**Table 2.2 Inverter Options**

| Topology                                     | # of Switches | Advantages                                                                      | Disadvantages                                                                  |
|----------------------------------------------|---------------|---------------------------------------------------------------------------------|--------------------------------------------------------------------------------|
| Single-bus with 2 paralleled half-bridges    | 4             | <ul style="list-style-type: none"> <li>• Minimum component count</li> </ul>     | <ul style="list-style-type: none"> <li>• Large DC filter components</li> </ul> |
| Dual-bus with 2 split half-bridges (modular) | 4             | <ul style="list-style-type: none"> <li>• Reliability and flexibility</li> </ul> | <ul style="list-style-type: none"> <li>• High component count</li> </ul>       |
| Single-bus with 3 wire single phase          | 6             | <ul style="list-style-type: none"> <li>• Small passive components</li> </ul>    | <ul style="list-style-type: none"> <li>• Complex control</li> </ul>            |

### **2.1.2 Energy Storage**

In order to operate the fuel cell efficiently, the flow rate of hydrogen must be adjusted with changes in electrical load. Depending on the type of fuel cell system, this flow change can be a very slow process and can have time constants in the 30-second range. Therefore, some type of supplementary energy storage is typically required.

Batteries are the common choice and can be a good one if a high voltage battery configuration is used. This high voltage string of batteries can be placed across the regulated DC bus with little or no additional circuitry. The high side placement may also allow for a de-rated front-end, depending on the battery size. However, large number of cells in series can be expensive and often lead to unbalance issues [26]. Also additional inductors may be needed in series with the battery to block the 120 Hz ripple caused by the inverter. Furthermore, battery placement on the high side may not provide the auxiliary power necessary to start-up the fuel cell ancillary system. Low voltage battery configurations are more desirable, but require additional circuitry to ensure safe operation. If the low voltage batteries are interfaced with the high side, a bi-directional DC-DC converter is needed. The low voltage battery string could be placed on the low side and controlled with semiconductor switches or a contactor. However, batteries in general have a short lifetime and if stacked in series, often require some type of complex voltage monitoring scheme. Battery management systems can be effective, but not without a significant cost added to the system.

An interesting alternative to batteries is the use of ultracapacitors, which have a wider voltage range than batteries and can be directly paralleled across the input bus. Ultracapacitors have a specific energy density less than that of a battery, but a specific power greater than a battery, making them ideal for short (up to several seconds) pulses of power. Certain ultracapacitors (unsymmetrical electrochemical) can hold charge over extended periods of time, so as to act somewhat like a battery. However, unlike batteries, these ultracapacitors have a short charge time and a much longer lifetime. The energy storage options are summarized in Table 2.3.

**Table 2.3 Energy Storage Options**

| Energy Storage | Voltage/ Location   | Advantages                                                                                                                                                                     | Disadvantages                                                                                                                                                                                                                                          |
|----------------|---------------------|--------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|--------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------------|
| Battery        | 12 V,<br>Any side   | <ul style="list-style-type: none"> <li>• Minimum # of batteries</li> </ul>                                                                                                     | <ul style="list-style-type: none"> <li>• Requires bi-directional DC-DC converter</li> </ul>                                                                                                                                                            |
| Battery        | 400 V,<br>High side | <ul style="list-style-type: none"> <li>• Allow smaller size front-end DC-DC converter</li> <li>• Stable high-side dc bus voltage</li> <li>• No additional circuitry</li> </ul> | <ul style="list-style-type: none"> <li>• Lack of start-up and support for fuel cell ancillary system</li> <li>• Large number of batteries, low life-expectancy due to unbalanced cell voltages</li> <li>• Additional inductor may be needed</li> </ul> |
| Battery        | 48 V,<br>Low side   | <ul style="list-style-type: none"> <li>• Reasonable # of batteries</li> </ul>                                                                                                  | <ul style="list-style-type: none"> <li>• Requires additional circuitry for protection</li> <li>• Full power DC-DC converter needed</li> </ul>                                                                                                          |
| Ultracapacitor | 48 V,<br>Low side   | <ul style="list-style-type: none"> <li>• No additional circuitry,</li> <li>• No maintenance, long life</li> </ul>                                                              | <ul style="list-style-type: none"> <li>• Availability and cost</li> <li>• Full power DC-DC converter needed</li> </ul>                                                                                                                                 |

## 2.2 Chosen Design

A prototype fuel cell inverter was designed and built, consisting of two modular 5 kW units. The basic power stage topology chosen was a full-bridge front-end and half-bridge inverter. A digital signal processor (DSP) was used for control purposes and to communicate with the fuel cell. Ultracapacitors were placed across the input bus and used as the energy storage device. Some of the basic design procedures and philosophies are summarized in the remainder of the chapter, while the main parameters for one modular unit are given in Table 2.4 where  $f_{\text{smos}}$  and  $f_{\text{sigbt}}$  are the switching frequencies.

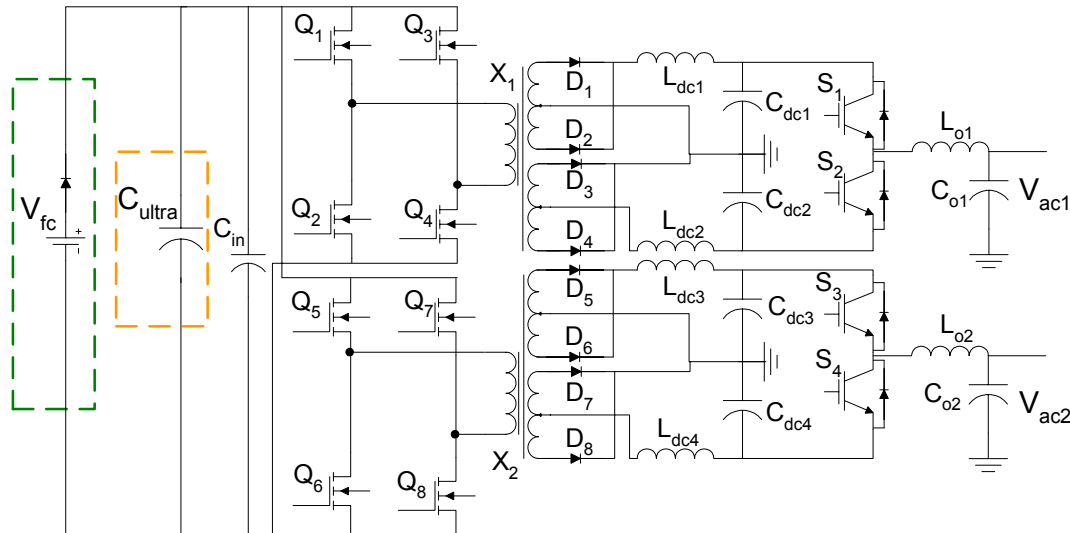
**Table 2.4 Design Parameters**

| Parameter          | Value                |
|--------------------|----------------------|
| $P_{\text{Omax}}$  | 5 kW                 |
| $V_{\text{INmin}}$ | 42 V                 |
| $V_{\text{IN}}$    | 48 V                 |
| $V_{\text{INmax}}$ | 65 V                 |
| $V_{\text{ac}}$    | 120 V <sub>rms</sub> |
| $I_{\text{INmax}}$ | 120 A                |
| $I_{\text{Omax}}$  | 41 A <sub>rms</sub>  |
| $f_{\text{smos}}$  | 28 kHz               |
| $f_{\text{sigbt}}$ | 24 kHz               |

### 2.2.1 Power Stage

It is critical to look at the big picture when designing the power stages of the system. The chosen topology, device ratings, and passive component size will affect system

efficiency, performance, and cost. It is impossible to minimize both the number of components and the component ratings. Therefore some trade-offs had to be made based on the application. For a stand-alone power generation system, it was decided that a redundant system would be desired by the user and would allow for lower component ratings. The final topology, consisting of two modular units, is shown in Figure 2.4. Although this configuration has a larger component count than a single DC bus topology, the device ratings can be significantly reduced. In comparison to the single DC bus topology (with paralleled MOSFETs), no additional devices are necessary for the dual full-bridge circuits. For the rectification stage, it is possible to use a common cathode discrete package and still get away with only four devices. Inductors can be costly, so the addition of  $L_{dc3}$  and  $L_{dc4}$  to the circuit presents a clear disadvantage. However, the current through each inductor is cut in half, which can greatly reduce the size of the core needed. Although not implemented, there is also the option of interleaving  $L_{dc1}$  with  $L_{dc3}$  and  $L_{dc2}$  with  $L_{dc4}$ . Coupling of inductors can reduce the current ripple and significantly reduce the size of the inductors [27]. Two dc bus capacitors are also added, but the capacitance can be reduced to half compared to a single dc bus system. The remaining components are needed regardless of the topology configuration.



**Figure 2.4 10 kW Inverter Topology**

For reasons explained earlier in this chapter, a full bridge configuration was chosen for the DC-DC converter. The input side involves relatively low voltage and relatively high current, which makes the power MOSFET the device of choice. It is very important to find a device with significant voltage margin, but still very low on-resistance. To keep cost at a minimum, it is also desirable to use discrete devices and eliminate the use of snubber and clamping circuits. Although MOSFETs in the 75-100 V range have limited room for voltage overshoot, their lower on-resistance make them quite attractive for this application. The low device voltage rating requires that the circuit parasitics and transformer leakage inductance be minimized. A complete list of components used in the converter is given in Appendix B.

The gate drive circuit for the MOSFETs uses a very functional integrated circuit, the HCPL316J, which provides sufficient driving current and optical isolation for both input and fault signals. The high current output capability of the chip can directly drive the devices without an extra driving stage. The HCPL316J gate driver chip also provides de-saturation (de-sat) protection when the device is over current or short-circuited. The de-sat function reduces the gate drive power supply to zero so that the device turns off when the device voltage exceeds a specified limit, which is typically caused by over-current or short circuit conditions. The turn-off action does not need to go through additional logic or through the DSP, and as a result, trips the device on a microsecond time scale.

The full-bridge topology also allows for a phase-shift modulation scheme to be used in order to reduce switching loss of the front-end converter at large loads. A phase-shifted full-bridge operates by complementary switching of the devices in each leg and then shifting one leg with respect to the other. This results in a quasi-square wave voltage across the primary of the transformer. The basic operation during two switching cycles is shown in Figure 2.5. More details on the operation of a full-bridge phase-shifted PWM converter can be found in [28] and [29].

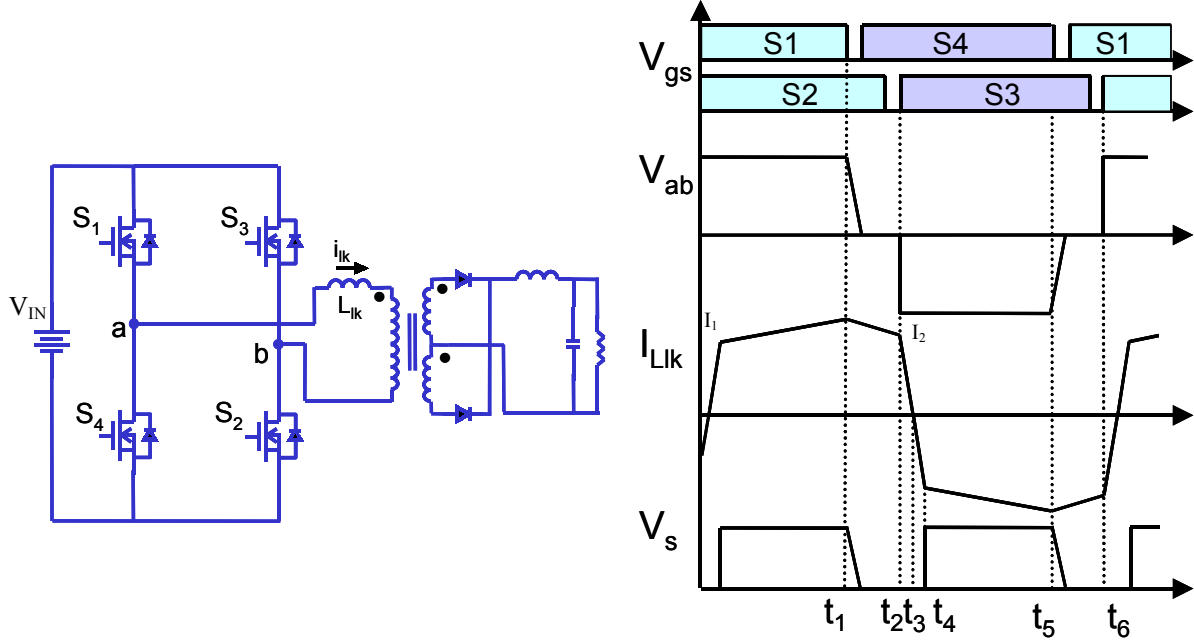


Figure 2.5 Phase-Shift Operation

The magnitude of the transformer leakage inductance is a crucial design variable with this topology. It helps the lagging leg switches achieve zero voltage switching (ZVS) by discharging the capacitance across the device before it turns on. However, the leakage inductance also causes a loss of duty cycle as shown in (2.1), where  $\Delta D$  is the change in duty cycle,  $I_1$  and  $I_2$  are related to the peak current at different charging intervals, and  $I_o$  is the output current of the converter [28]. In addition, the leakage inductance together with the parasitic capacitance of the circuit can result in large parasitic ringing in the secondary side diodes. Considering the input voltage of the system, it was more desirable to minimize the leakage inductance and consider the ZVS at large loads as an added benefit.

$$\Delta D = \frac{\frac{I_1 + I_2}{V_{IN}} \frac{T_s}{2}}{\frac{L_{lk}}{2}} \approx \frac{\frac{2I_o}{V_{IN}} \frac{T_s}{2}}{\frac{L_{lk}}{2}} \quad (2.1)$$

The transformer design was a challenging task considering the high power and multiple secondary outputs required. The size of magnetic components is directly related to power and inversely related to the switching frequency. The large input current required at full load makes it difficult to push the switching frequency of the dc-dc

converter much past 50 kHz. Multiple secondary outputs were required in attempt to provide individual voltages to the DC bus capacitors and thus eliminate unbalanced conditions. The center-tapped structure also allowed for full-bridge rectification using only two diodes. Given a minimum input voltage,  $V_{INmin}$ , of 42 volts, the required output of 120 volts rms for each half-bridge inverter, and an ideal modulation index,  $m$ , to be around 85%, then the turns ratio of the transformer can be calculated from (2.2) and (2.3). These equations involve  $D$  as the primary duty cycle,  $N_p$  as the number of turns on the primary and  $N_s$  as the number of turns on the secondary of the transformer.

$$V_{DCbus} = \frac{V_{ac} \sqrt{2} \cdot 2}{m} = 400 V \quad (2.2)$$

$$\frac{1}{n} = \frac{N_s}{N_p} = \frac{\frac{V_{DCbus}}{2}}{V_{INmin} (D - \Delta D)} = 5.9 \quad (2.3)$$

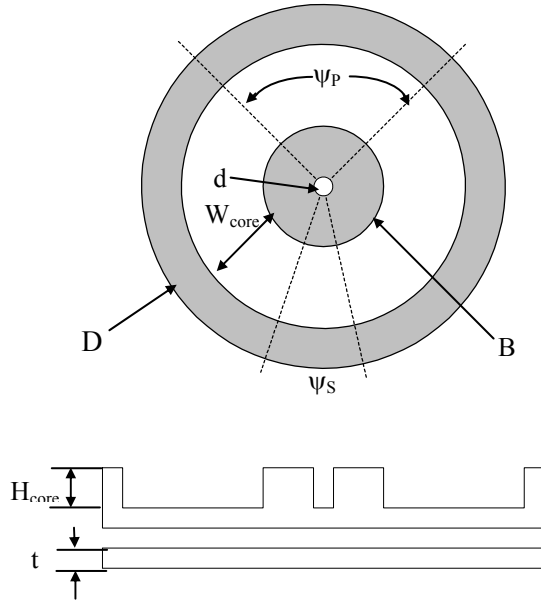
To allow for losses and for a slightly higher DC bus voltage ( $V_{DCbus}$ ), the final turns ratio,  $n$ , used was 1:6.5. From the turns ratio and dc bus voltage calculations above and assuming the efficiency of the converter is 90%, the maximum duty cycle can be approximated as in (2.4).

$$D_{max} = \frac{\frac{V_{DCbus}}{2}}{\frac{V_{INmin}}{n} \cdot \eta} = 0.81 \quad (2.4)$$

When designing for a ferrite core, a reasonable maximum flux density is 2500 Gauss. For a fuel cell the maximum power will occur at the minimum voltage so  $V_{INmin}$  and  $D_{max}$  can be used in a version of Faraday's Law to find the turn-area product,  $NA$ , shown in (2.5).

$$NA = \frac{V_{INmin} T_s D_{max}}{2 \Delta B} = 1221 mm^2 \quad (2.5)$$

The large turns ratio, the multiple secondary outputs, and the 5 kW power level all make utilizing a planar structure very beneficial. Due to availability, the custom planar pot core structure shown in Figure 2.6 was used.



**Figure 2.6 Planar Core Structure**

**Table 2.5 Core Dimensions**

| Description         | Symbol     | Dimension           |
|---------------------|------------|---------------------|
| Outer Diameter      | D          | 96.3 mm             |
| Center Leg Diameter | B          | 28 mm               |
| Window Width        | $W_{core}$ | 32 mm               |
| Window Height       | $H_{core}$ | 7.1 mm              |
| Shell Thickness     | t          | 6.7 mm              |
| Centering Hole      | d          | 3 mm                |
| Area                | A          | 615 mm <sup>2</sup> |
| Primary Gap Angle   | $\Psi_P$   | 89°                 |

Material = MN60LL

Since the cross sectional area of the core is only 615 mm<sup>2</sup>, according to (2.5) two turns are required on the primary. A multi-layer printed circuit board (PCB) was designed for the windings of the transformer. Using a PCB for the windings of the transformer can help minimize leakage inductance and create an easy assembly process. The cross sectional area of copper required is calculated for the primary and secondary using (2.6) and (2.7), which takes into account rms current and temperature rise [30].

$$Area_p = \left( \frac{I_p}{k \cdot t^{0.44}} \right)^{1/0.725} = 20530 \text{ mils}^2 \quad (2.6)$$

$$Area_s = \left( \frac{I_s}{k \cdot t^{0.44}} \right)^{1/0.725} = 1166 \text{ mils}^2 \quad (2.7)$$

In (2.6) and (2.7) the *Area* is determined in mils<sup>2</sup>, where *I* is the current in amps, *t* is the temperature rise of the copper in °C (20° is used), and *k* is a constant which depends on whether it is an outer layer or an inner layer. Although skin effect is not a big problem at this switching frequency, the skin depth calculation is shown in (2.8) and (2.9) is used to determine the limitation of copper thickness in ounces. Here  $\rho$  is the resistivity of copper in  $\Omega\text{m}$ ,  $\mu_r$  is the relative permeability of copper, and *oz* is the standard thickness of 1.4 mils referring to a square foot of copper.

$$\delta = \frac{10^5}{2.54} \sqrt{\frac{\rho}{\pi \cdot \mu_r \cdot \mu_o \cdot f_{smos}}} = 15.32 \text{ mils} \quad (2.8)$$

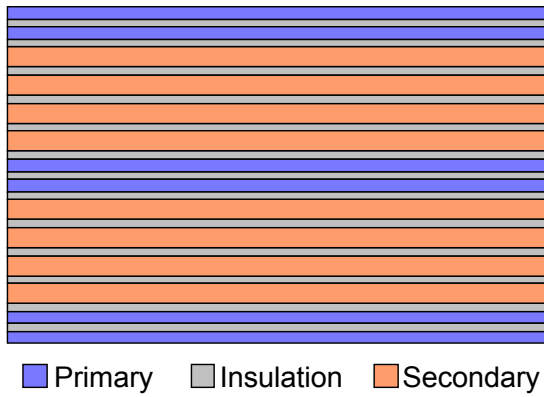
$$Cu_{\max} = \frac{\delta}{oz} = 10.7 \text{ ounces} \quad (2.9)$$

The cross sectional area calculated above can then be used to determine the ounce of copper required for the primary and the secondary:

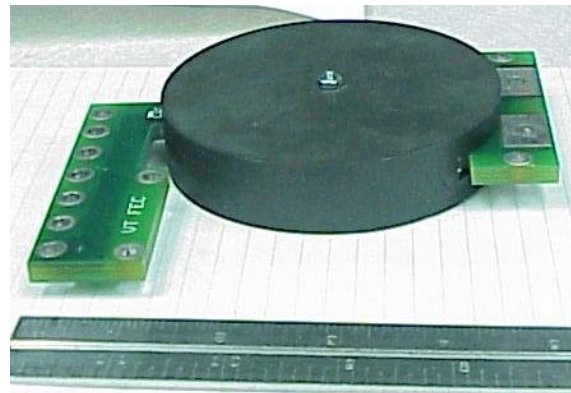
$$Cu_p = \frac{\frac{Area_p}{(W_{core} - tol)}}{\frac{Turnsperlayer_p}{oz}} = 14.1 \text{ ounces} \quad (2.10)$$

$$Cu_s = \frac{\frac{Area_s}{(W_{core} - tol)}}{\frac{Turnsperlayer_s}{oz}} = 5.6 \text{ ounces} \quad (2.11)$$

where *tol* is a factor taking into account the tolerance of the PCB and the core. Since it is desired to minimize the leakage inductance, an interleaved winding structure was chosen. The primary consists of six layers of three oz copper in parallel, and the secondary consists of eight layers of six oz copper resulting in a 14-layer PCB structure, shown in Figure 2.7. The measured leakage inductance of the complete transformer, which is shown in Figure 2.8, is 100 nH.



**Figure 2.7 PCB Structure**



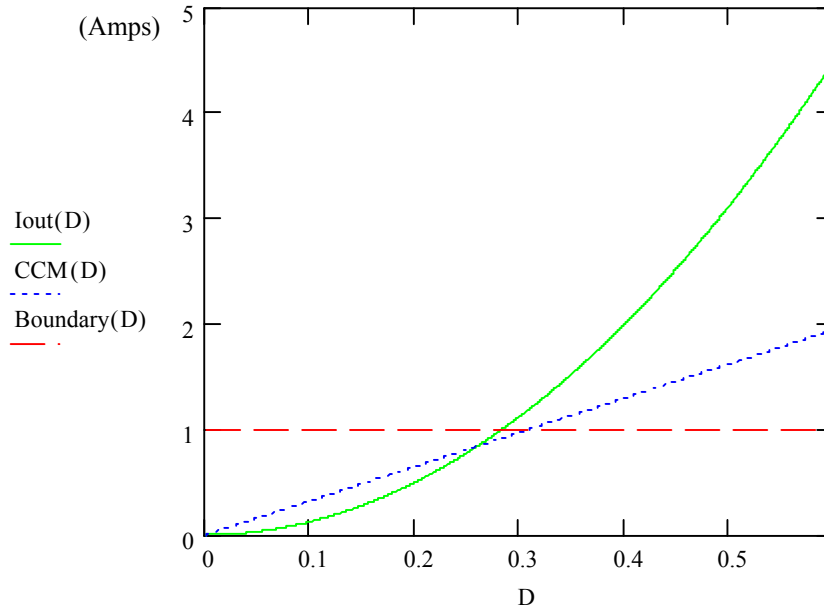
**Figure 2.8 Planar Transformer**

The center-tapped transformer allows a minimum number of rectifier diodes to be used. These devices must be fast recovery and have a very high voltage rating to compensate for the ringing in the secondary described earlier. After the full wave rectification, a LC filter is required to create two 200 V references and thus a 400 V DC

bus. The desired cut-off frequency, the DC bus voltage ripple, and the boundary between discontinuous conduction mode (DCM) and continuous conduction mode (CCM) all affect the design of the inductor and the capacitor. The response of the full-bridge phase-shifted DC-DC converter varies depending on the current waveform of the filter inductor. More details on the response related to CCM and DCM are given in Chapter 3. It is desirable to have the converter work in CCM over most load conditions, so a relatively large inductance is required. According to [34], the condition for the converter to reach CCM is when (2.12) is equal to one.

$$Boundary(D) = D + \frac{2L_{dc} \cdot I_O \cdot (2f_{smos}) \cdot n}{D \cdot V_{in}} \quad (2.12)$$

Clearly, (2.12) is directly related to the load and the inductance value,  $L_{dc}$ . Assuming a filter inductance value of about 500  $\mu\text{H}$ , the graph shown in Figure 2.9 can help determine at what load condition the converter will enter CCM. From the graph below it is shown that the converter will enter CCM at an output current of slightly more than 1 A, which is about 450 W (<10% of rated power).



**Figure 2.9 DC-DC Load Current and CCM condition vs. Duty Cycle**

Given the inductance value, the DC bus capacitance can be determined by (2.13) assuming a desired cutoff frequency,  $f_c$ , of about 200 Hz. The final capacitance value

was increased to 2.2 mF to help absorb more of the 120 Hz ripple expected from the inverter.

$$C_{dc} = \frac{1}{(2\pi f_c)^2 \cdot L_{dc}} = 1.3mF \quad (2.13)$$

The inverters are half-bridge topologies utilizing a split capacitor bus. As mentioned previously, however, this split bus is not like the typical half-bridge configuration powered by one source. Because of the way the transformer was designed, the capacitors do not divide the source in half. This prevents any imbalance issues and helps reduce the number of components needed.

The voltage is larger on the secondary side, so IGBTs rated at 600 V were chosen to meet this criterion. These IGBTs are moderately fast and capable of 60 amps of continuous current. The gate drive circuit for the IGBTs is very similar to that used for the MOSFETs in the DC-DC converter. It also protects the devices during an output short circuit condition.

The PWM output from the IGBT half-bridge is then filtered using an inductor and a capacitor to get a clean sinusoidal output. It is desired to have the cutoff frequency,  $f_{co}$ , of the filter around 1 kHz to eliminate high frequency switching ripple, yet still maintain reasonably sized components. It is also very important to have the characteristic impedance of the filter match the maximum load impedance, so that the inrush current to the filter capacitor is limited. Using (2.14), solving for  $L_o$  and substituting into (2.15), the filter component values can be selected. In the inductor design, careful consideration was taken to avoid core saturation and limit the core loss. The determined inductance value was 330  $\mu$ H and the final capacitance value was chosen to be 50  $\mu$ F.

$$Z_o = \sqrt{\frac{L_o}{C_o}} = R_{L\_5kW} = 2.88\Omega \quad (2.14)$$

$$f_{co} = \frac{1}{2\pi \sqrt{L_o C_o}} \approx 1 \text{ kHz} \quad (2.15)$$

### 2.2.2 Ultracapacitor

In this particular system, the ultracapacitors are used to initially power the auxiliary supplies of the inverter and to supplement the fuel cell during load transients. The ultracapacitors can be sized according to the worse case transient condition and the time constant of the fuel cell. If the fuel cell had a time constant,  $\tau$ , of 30 seconds and there was a load step of 10 kW, the energy required from the ultracapacitors is found in (2.16).

$$E_{10kW\_transient} = \Delta P_{10kW} \int_0^{5\tau} e^{\frac{-t}{\tau}} dt = 90 \text{ Wh} = 324 \text{ kJ} \quad (2.16)$$

A 10 kW step in load in a house is extremely unlikely, so ultracapacitors sized at about half of the energy calculated in (2.16) are probably sufficient. Table 2.6 illustrates the characteristics of two different ultracapacitors used with the prototype system.

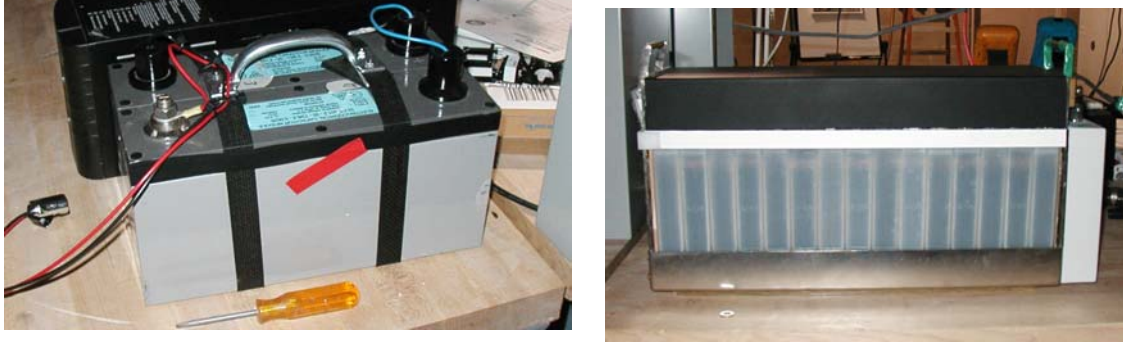
**Table 2.6 Ultracapacitor Specifications**

| Type                                   | Characteristic       | Value      |
|----------------------------------------|----------------------|------------|
| Unsymmetrical<br>(used four in series) | Capacitance          | 600 F      |
|                                        | ESR (at 25° C)       | 3.5 mΩ     |
|                                        | Useful Voltage Range | 6.5 – 13 V |
|                                        | Absolute Max Voltage | 14.5 V     |
|                                        | Energy               | 35 kJ      |
| Symmetrical<br>(used two in parallel)  | Capacitance          | 1.65 F     |
|                                        | ESR (at 25° C)       | 140 mΩ     |
|                                        | Useful Voltage Range | 0 – 110 V  |
|                                        | Absolute Max Voltage | 110 V      |
|                                        | Energy               | 20 kJ      |

The unsymmetrical ultracapacitors will maintain a charge over long periods of time; however, their useful energy range is limited to roughly half of the rated voltage. Yet the unsymmetrical type in Table 2.6 would still meet the energy requirement as illustrated by (2.17).

$$E_{Unsymmetrical} = \frac{1}{2} C \left( V^2 - \left( \frac{V}{2} \right)^2 \right) = 152 \text{ kJ} \quad (2.17)$$

Two different types of the unsymmetrical ultracapacitors are shown in Figure 2.10. The symmetrical ultracapacitors (not shown) can be used until the voltage is zero and there is no energy left. Conversely, these types of ultracapacitors tend to have a large equivalent series resistance (ESR) and therefore discharge much faster when dormant.



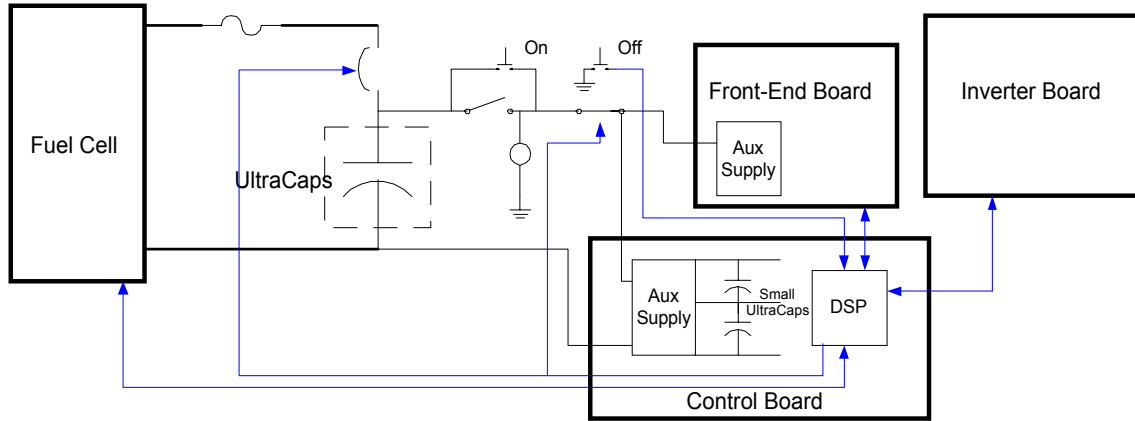
**Figure 2.10 Ultracapacitors**

### **2.2.3 Control**

The control for the entire inverter system is done with the Analog Devices ADMC401 DSP. The ADMC401 chip is a 26 MIPS, fixed-point processor. This DSP board has six PWM signals available, 12 general-purpose I/O pins, 8 analog-to-digital (A/D) inputs, and 8 digital-to-analog (D/A) outputs.

Due to the limited PWM signals, the phase shifted gate signals for the DC-DC converters are performed by the UC3895 chip from Texas Instruments. A control signal from the DSP via the D/A is sent to the UC3895, which then determines the required phase shift for the front-end converter. If a DSP with more PWM signals were available, it would be feasible to eliminate the UC3895 chip and create the phase-shifted signals with the DSP. Eliminating the UC3895 would not only reduce the cost, but it would also eliminate some irritating noise problems. The two half-bridge inverters are controlled entirely by the DSP, which generates two sets of complementary PWM signals and sends them to the gate drivers.

The A/D inputs are used to read in the feedback signals from the system, including input voltage, DC bus voltage, and output voltage. From a system standpoint, the DSP board is in complete control. A small PCB interfaces the DSP with the rest of the system, filters the A/D inputs, and buffers the digital outputs. All of the sensing parameters are monitored for control and for fault conditions. A simple interface of LEDs and on/off buttons communicate information to the user. If necessary, communication between the fuel cell is also done with the DSP. A control diagram is shown in Figure 2.11, where the blue lines represent control signals to and from the DSP.



**Figure 2.11 Control Block Diagram**

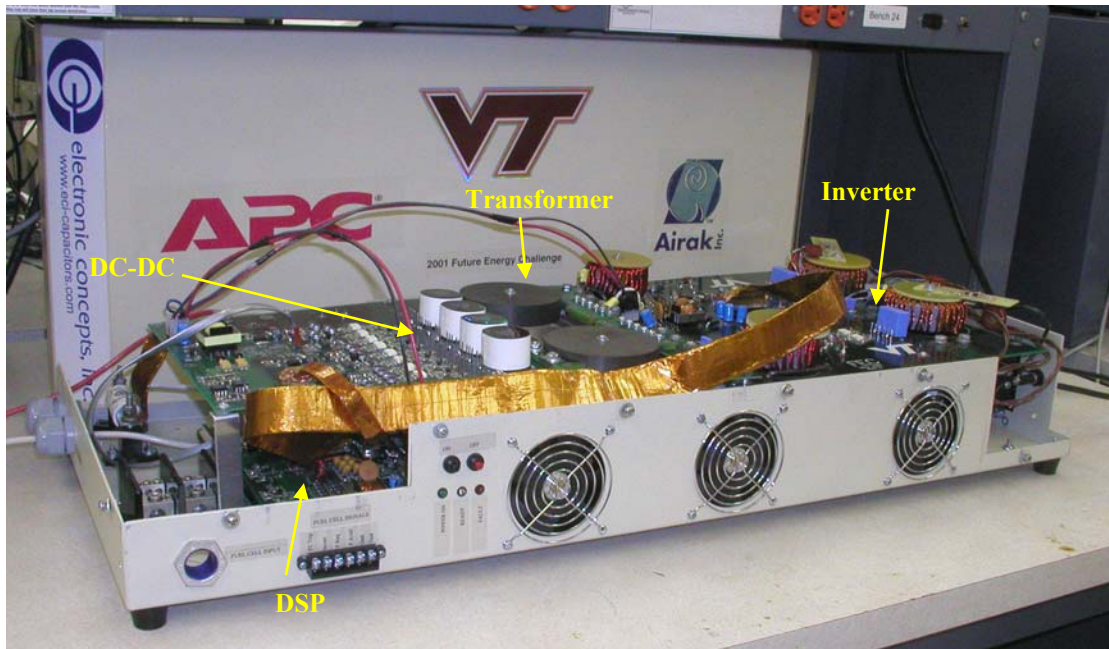
### 2.2.4 System

The emphasis on integration, isolation, protection, and manufacturability all lead to a low cost, high performance system. There are three PCBs that make up the entire system. The only connections that are necessary, besides input and output, are from front end to the transformer, from transformer to the inverter, and then a few plug-in cables from the control board to each of the power boards. The auxiliary power supplies, gate drives, sensing circuits, and filters are completely integrated into the power stage boards. The main transformers are planar and consist of PCBs that drop directly into the core. This provides complete isolation from input to output and from power to control.

Protection for under voltage, over voltage, and short circuit are all performed each interrupt cycle and appropriate action is taken if a fault occurs. For instance, if a de-sat fault is detected by the DSP from the MOSFET or IGBT gate drivers, it will try to restart once more to make sure it wasn't noise, and then it will immediately shut down the system and light up the fault LED. Input and output fuses are also installed for added protection.

Packaging of power electronic circuits is a feature that can easily be overlooked. It can be a major source of cost and complication in a system if it is not addressed correctly. Cooling and layout are two areas that must be examined to obtain an effective design. Three heat sinks are necessary to keep the device temperatures at an acceptable level. The heat sinks are sized assuming 90% overall efficiency at 10 kW, so each heat sink should dissipate approximately 300 watts. Three small AC fans are added to keep air flowing throughout the enclosure and to provide a minimum load to the system. A

picture of the packaged hardware prototype is shown in Figure 2.12. Although the packaging of the prototype was far from optimized, it did fit together nicely and allow for easy debugging and component replacement.



**Figure 2.12 Fuel Cell Inverter**

## Chapter 3 – Modeling and Control

### 3.1 Fuel Cell

Although often assumed to be, the fuel cell is not an ideal voltage source. Because of the sloping V-I curve that fuel cells exhibit, their electrical characteristics cannot be completely ignored during the modeling process. Figure 3.1 shows how the output voltage of a 1.8 kW PEM fuel cell varies with load and with fuel flow rate.

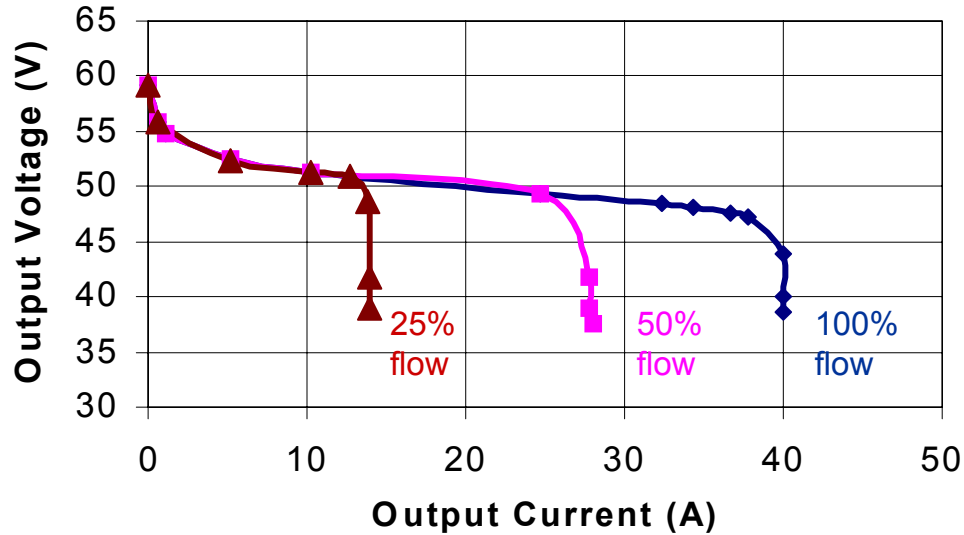


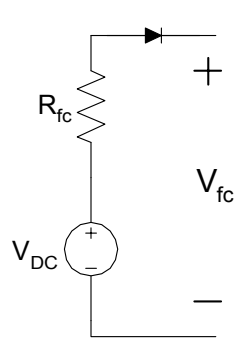
Figure 3.1 PEM Fuel Cell Stack (1.8 kW) Output Characteristics

The output voltage variation of the fuel cell results in an input voltage variation to the inverter. This means that the inverter should not only have good audio susceptibility, but also be resilient to large signal changes in input voltage. Depending on the fuel cell stack, the hydrogen flow rate may change according to the load in attempt to operate the fuel cell in its most efficient region. This change in hydrogen flow may involve mechanical valves and other relative slow processes, which are controlled on the fuel cell side. Therefore the response of the fuel cell may be rather slow to changes in inverter load, and it may require some communication from the inverter.

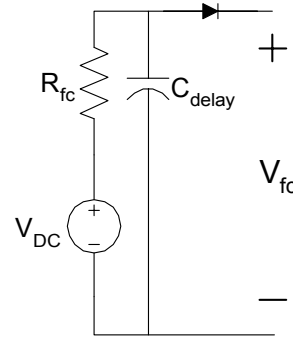
#### 3.1.1 Modeling

From the electrical point of view, the simplest way to model the fuel cell is to assume its characteristics are completely linear. Then the fuel cell can be viewed as an ideal voltage source with a series resistance proportional to the slope of the middle region

of the fuel cell V-I curve. An ideal diode can be added to prevent current from flowing back to the stack, as shown in Figure 3.2. This model is accurate through the majority of the operating region of the fuel cell and is a very good approximation. If the fuel cell has a slow response time, a small capacitor can be added to give the voltage source a time constant as shown in Figure 3.3.

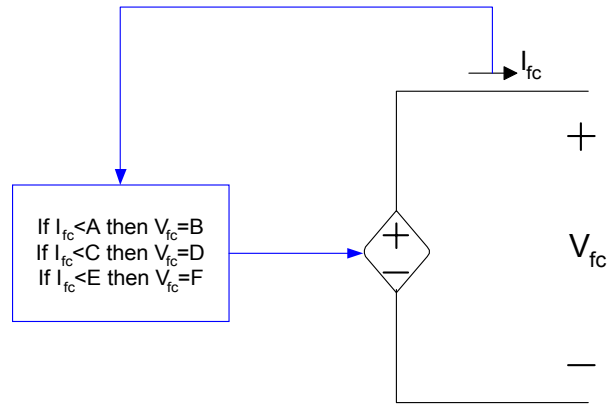


**Figure 3.2 First Order Model of Fuel Cell**



**Figure 3.3 First Order Model with Delay**

A more accurate electrical model of a fuel cell can be developed if needed. The nonlinear model requires current feedback and uses conditional equations to determine the output voltage. A block diagram version of this model is shown in Figure 3.4, and an example of code, written in MAST is shown in Appendix D. Other parameters, such as temperature and hydration levels, can be incorporated into the fuel cell model if necessary [35].



**Figure 3.4 Higher Order Model of Fuel Cell**

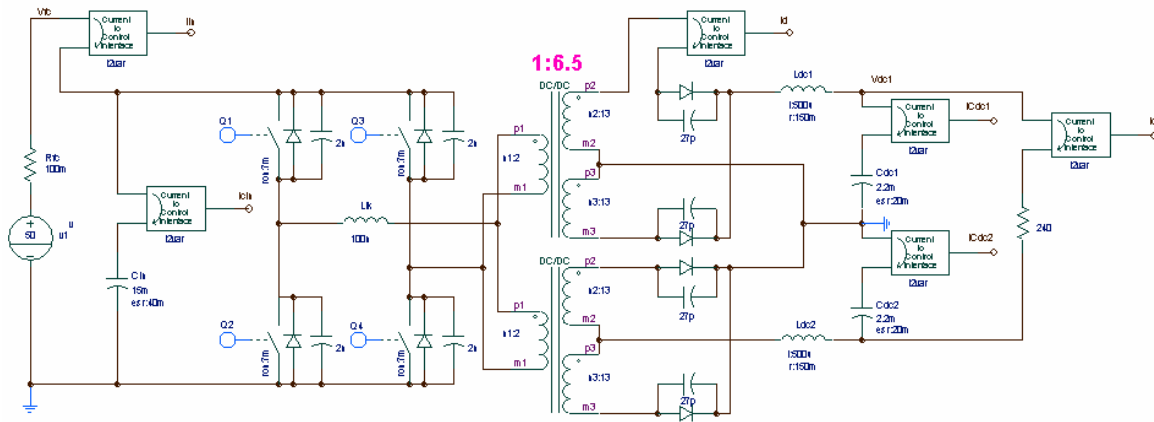
The linear fuel cell model (Figure 3.2 or Figure 3.3) was used for the majority of the control design and simulations in the following sections.

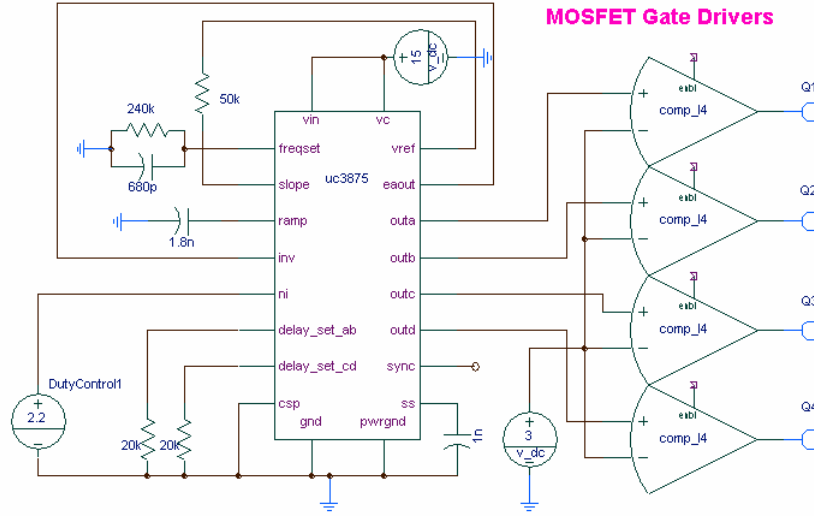
## 3.2 DC-DC Converter

### 3.2.1 Modeling

In order to accurately design the control for the DC-DC converter, it must be modeled and simulated. For simplification, only one of the 5 kW modules was used in the following analysis.

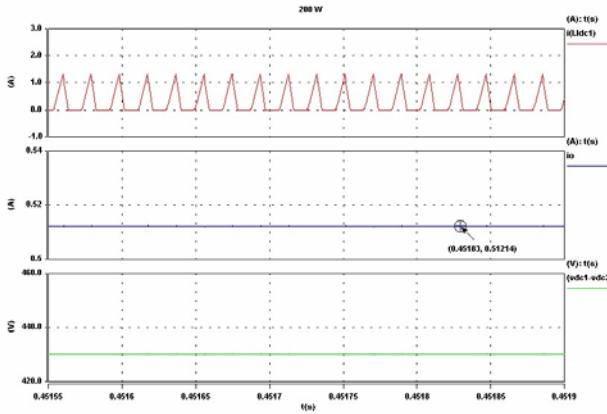
To verify the basic operation and the CCM/DCM boundary, an open loop switching model was developed in Saber as shown in Figure 3.5. The power stage uses ideal switches with an on-resistance and output capacitance for the MOSFETS and an ideal transformer with a separate leakage inductance on the primary side. The diodes are ideal with a junction capacitance associated with each one and the filter components have series resistances with values according to measurements of the physical components.



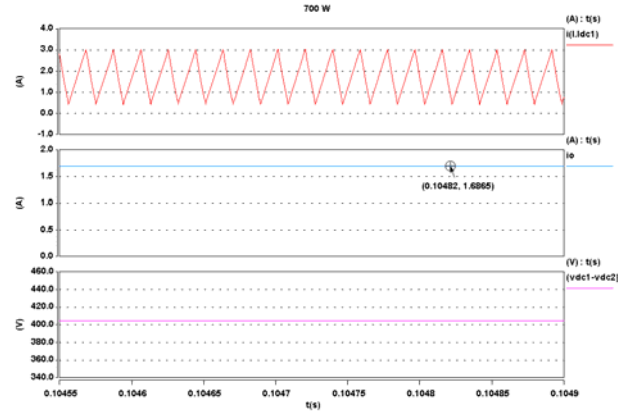


**Figure 3.6 Saber Model of UC3875**

Simulations were run at various loads to check the approximate boundary condition of the inductor and the soft switching of the devices. Figure 3.7 shows the inductor current, output current, and output voltage at light load condition of 200 W and Figure 3.8 shows the same circuit variables at 700 W.



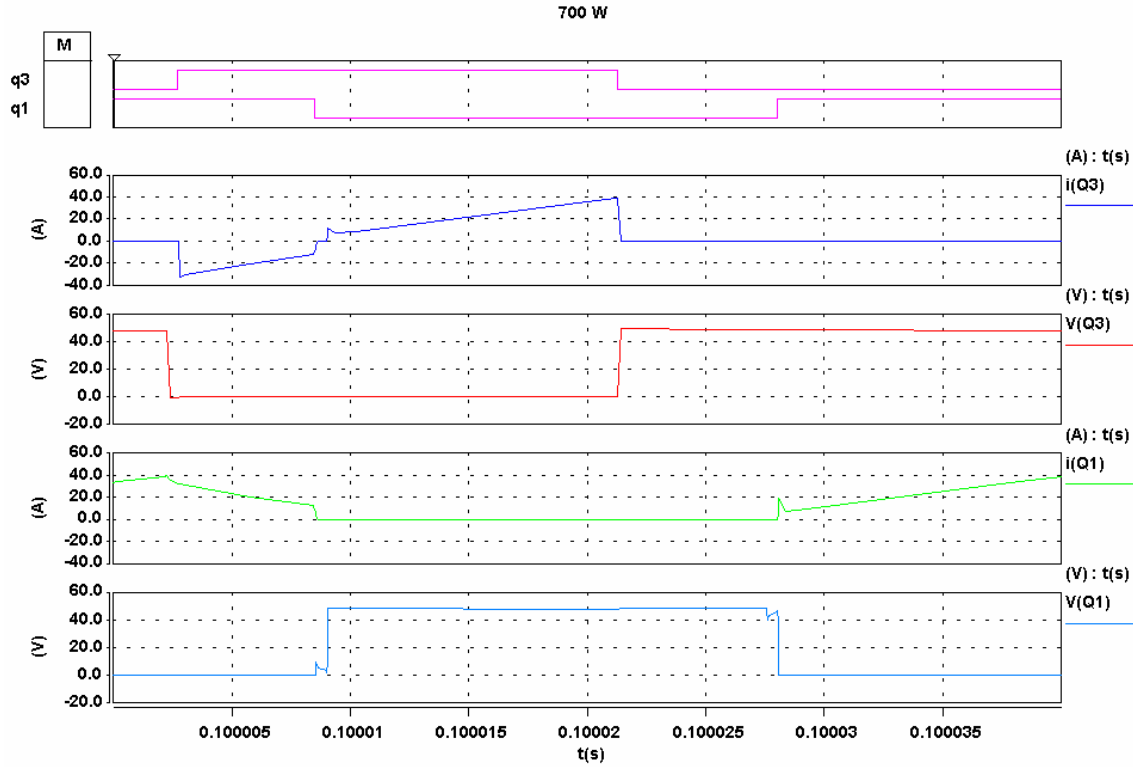
**Figure 3.7 Simulated DCM Operation of Front-End**



**Figure 3.8 Simulated CCM Operation of Front-End**

Figure 3.7 and Figure 3.8 clearly define the two different operating regions of the converter, with the transition occurring somewhere between 200 W and 700 W. It can be shown through simulation that the critical point occurs just below 400 W, which is very close to the predicted value calculated in Chapter 2. Figure 3.9 shows the gate signals, voltages, and currents of the upper two MOSFETs at a load of 700 W. The voltage across Q3 goes to zero before the gate turns on and before the current changes, thus

achieving zero voltage switching. The negative current through the device represents the conduction of the body diode. At this load condition (14% of rated load) only one leg achieves soft switching which is evident by the Q1 waveforms.



**Figure 3.9 Simulated ZVS of Leading Leg Device**

In order to design the control loop for the full-bridge phase-shifted converter, it is necessary to find a small signal model of the converter. Fortunately, Vlatkovic has already developed a small signal model in [33]. The conventional full-bridge topology is derived from the buck converter, and therefore has a very similar small signal model. The phase-shifted version is also similar to the buck, but slightly more complex. The main difference comes from the effective duty cycle that the phase-shifted converter sees, which is shown in (3.1). This duty cycle is now a function of input voltage, leakage inductance, output filter inductor current, and switching frequency. These parameters can affect the response of the converter and should be included in the small signal model.

$$D_{eff} = D - \Delta D = D - \left( \frac{\frac{N_s}{N_p}}{\frac{V_{in}}{L_{lk}} \frac{T_s}{2}} \left( 2I_L - \frac{V_o}{L_f} D \frac{T_s}{2} \right) \right) \quad (3.1)$$

The small signal model will depend on perturbations of filter inductor current, input voltage, and duty cycle. According to [33], the effect of duty cycle modulation,  $di_L$ , due to the change of filter inductor current,  $i_L$ , is shown in (3.2), where  $n$  is the turns ratio of the transformer defined as  $N_p/N_s$ . This effect is similar to a current feedback and will introduce additional damping into the system. The duty cycle modulation,  $dv_i$ , due to the change in input voltage is shown in (3.3), and it acts as a voltage feed forward increasing the audio susceptibility of the converter. Now the small signal duty cycle,  $d_{eff}$ , can be determined in (3.4). The prototype system was designed to have a small leakage inductance, which would normally make the  $di_L$  and  $dv_i$  terms negligible; however, since this is a boost application the turns ratio is large, so the effects of the phase-shift operation are still prevalent in the converter performance.

$$\hat{di}_L = -\frac{4L_{lk}f_s}{nV_{in}} \hat{i}_L \quad (3.2)$$

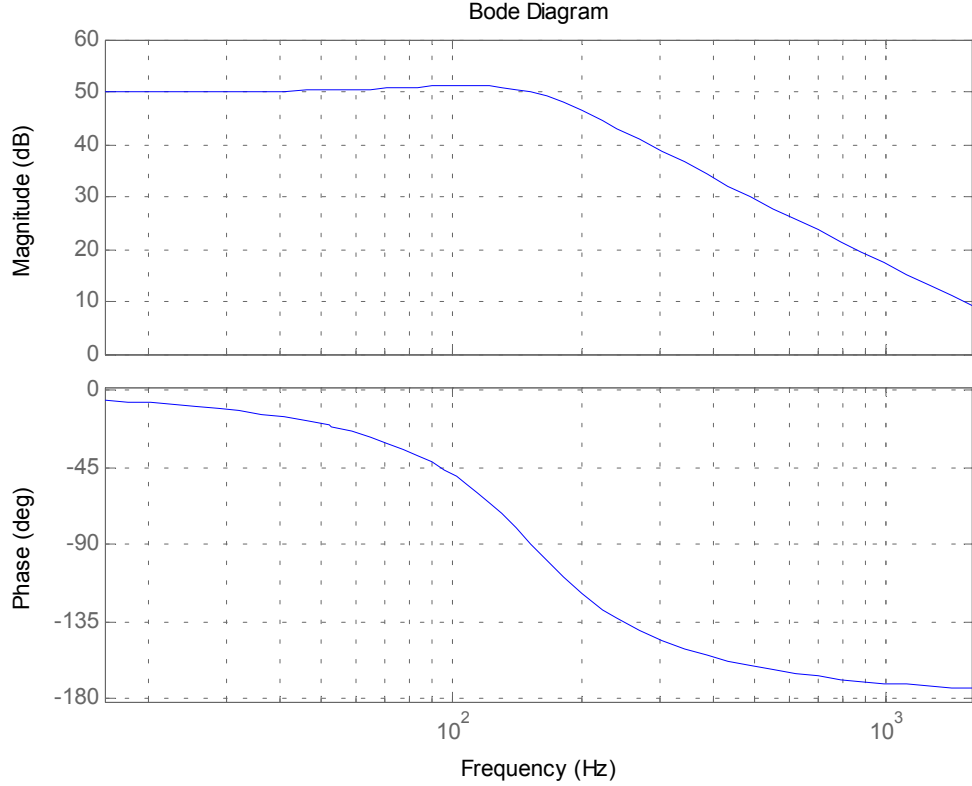
$$\hat{dv}_i = \frac{4L_{lk}f_s I_L}{nV_{in}^2} \hat{v}_{in} \quad (3.3)$$

$$\hat{d}_{eff} = \hat{d} + \hat{di}_L + \hat{dv}_i \quad (3.4)$$

Using the equations above and the knowledge of the standard buck converter small signal circuit, the small signal circuit of the phase-shifted converter in CCM can be constructed and is shown in Appendix C. Using the small signal circuit, transfer functions can be derived that will aid in the development of the control design. The control to output transfer function is given in (3.5), and the output impedance is given in (3.6). These equations are for CCM operation only and do not include ESR of the capacitors. The open loop plant transfer function of (3.5), for  $R=240 \Omega$ , is shown in Figure 3.10.

$$G_{vd} = \frac{n^{-1}V_{in}}{s^2LC + s\left(\frac{L}{R} + 4n^{-2}L_{lk}f_sC\right) + \frac{4n^{-2}L_{lk}f_s}{R} + 1} \quad (3.5)$$

$$Z_o = \frac{sL}{s^2LC + s\frac{L}{R} + 1} + \frac{4n^{-2}L_{lk}f_sR}{(1 + sRC)\left(s^2LC + s\frac{L}{R} + 1\right)4n^{-2}L_{lk}f_s + R\left(s^2LC + s\frac{L}{R} + 1\right)^2} \quad (3.6)$$



**Figure 3.10 Theoretical Control to Output Transfer Function in CCM**

To effectively and efficiently design the control loop for the converter, an average switch model of the phase-shifted PWM converter was developed. The average switch model can help predict both small and large signal behavior and significantly reduce simulation time. Tsai has already developed the basic three-terminal switch for the phase-shifted converter in [34]. This model is valid for both CCM and DCM and the averaged equations, corresponding to Figure 3.11, can be found in [34]. The model in DCM is actually the same as the conventional PWM converter.

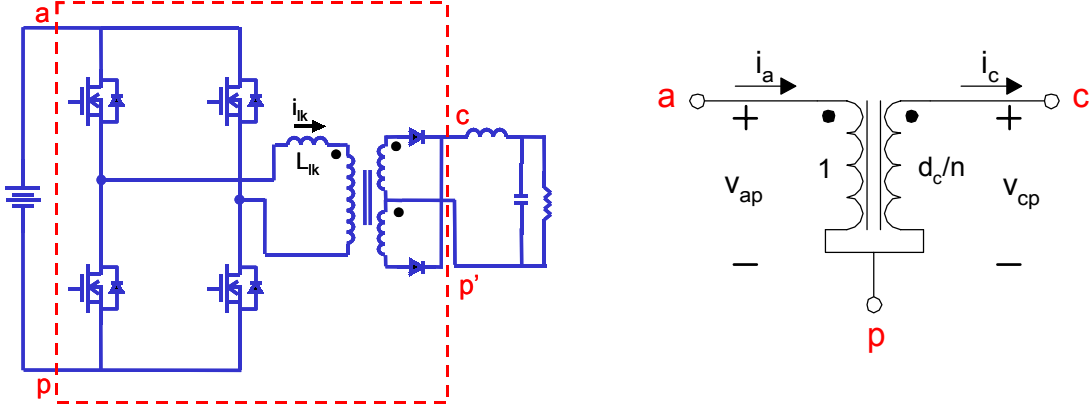


Figure 3.11 Single Output Three-Terminal Averaged PS-PWM Switch

The model for the single secondary output is shown in Figure 3.11, where  $p$  and  $p'$  are tied together [34]. This model can be modified to give the resultant dual secondary phase-shifted switch model as shown in Figure 3.12. Now the model has four terminals with two outputs ( $c$  and  $c-$ ), and the equations are given in (3.7) and (3.8). They are similar to the three-terminal model except that the input current,  $i_a$ , is doubled.

$$\left. \begin{aligned} v_{cp} &= \frac{v_{ap}(d - d_L)}{n} \\ i_a &= \frac{2i_c(d - d_L)}{n} \end{aligned} \right\} \text{in CCM where } d_L = \frac{4L_{lk}i_c f_s}{nv_{ap}} \quad (3.7)$$

$$\left. \begin{aligned} v_{cp} &= \frac{v_{ap}d}{n(d + d_2)} \\ i_a &= \frac{2i_c d}{n(d + d_2)} \end{aligned} \right\} \text{in DCM where } d_2 = \frac{4L_f i_c f_s n}{dv_{ap}} \quad (3.8)$$

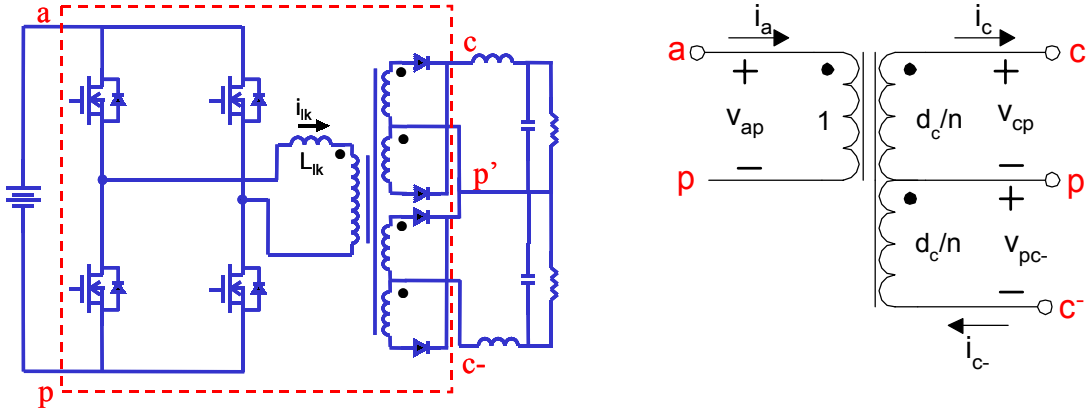
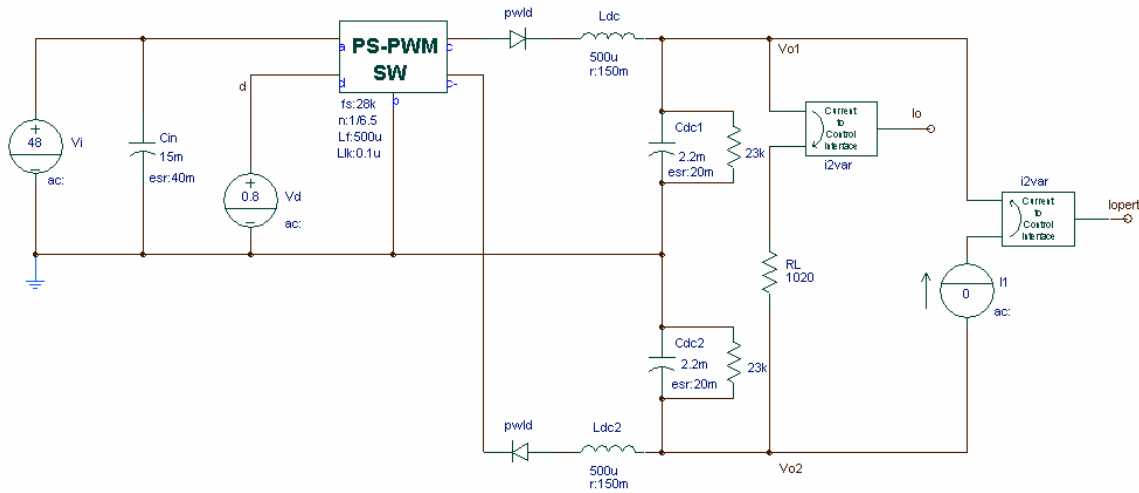


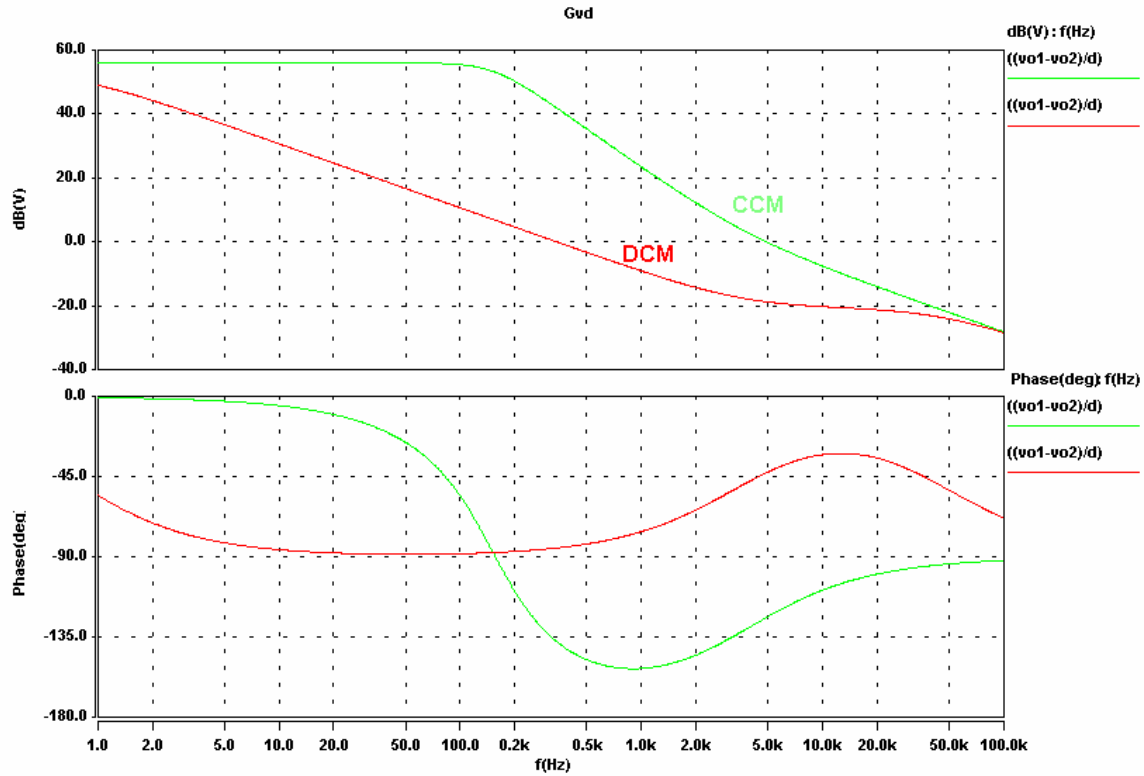
Figure 3.12 Dual Output Four-Terminal Averaged PS-PWM Switch

Using the new equations, the four-terminal average switch can be implemented in Saber and used in simulations. The detailed Saber implementation can be found in Appendix D. The PS-PWM switch model requires that the values of the switching frequency, the transformer turns ratio, the filter inductance, and the leakage inductance be passed to it. The duty cycle is the control input and the other terminals are connected to the circuit, as shown with the Saber simulation model in Figure 3.13. The ideal diodes are not necessary, but are added to account for their forward voltage drop. Despite the dual filter output, the inductance passed to the PS-PWM model is merely the value of a single filter inductor.



**Figure 3.13 Saber Open Loop Average Circuit of Front-End**

To verify the operation of the average model, several simulations were run and compared to the switching model. Frequency analysis was also performed to verify the transfer functions of the converter. Open loop control to output plots for DCM and CCM are shown in Figure 3.14. The CCM bode plot is very close to the theoretical plot shown in Figure 3.10, except that the simulation exhibits a zero around 5 kHz from the ESR of the capacitor. As expected, the phase-shift damping term eliminates the resonant peaking of the output filter.



**Figure 3.14 Simulated Control to Output Transfer Functions**

The open loop plots of audio susceptibility and output impedance for DCM and CCM conditions are shown in Figure 3.15 and Figure 3.16 respectively. Compared to the standard buck converter, the audio susceptibility of the phase-shifted converter is actually improved, because there is no resonant peaking. The output impedance, however, is larger than the standard PWM buck converter at low frequencies, which is mainly due to the negative sign in (3.2).

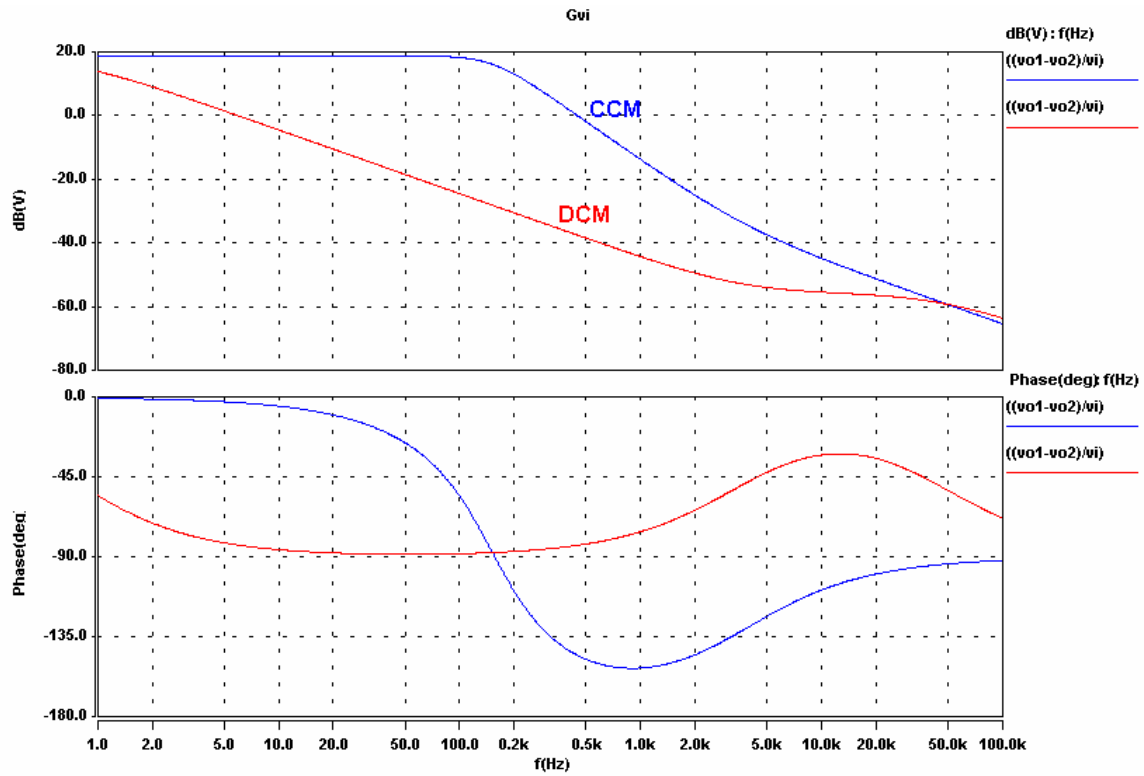


Figure 3.15 Simulated Audio Susceptibility

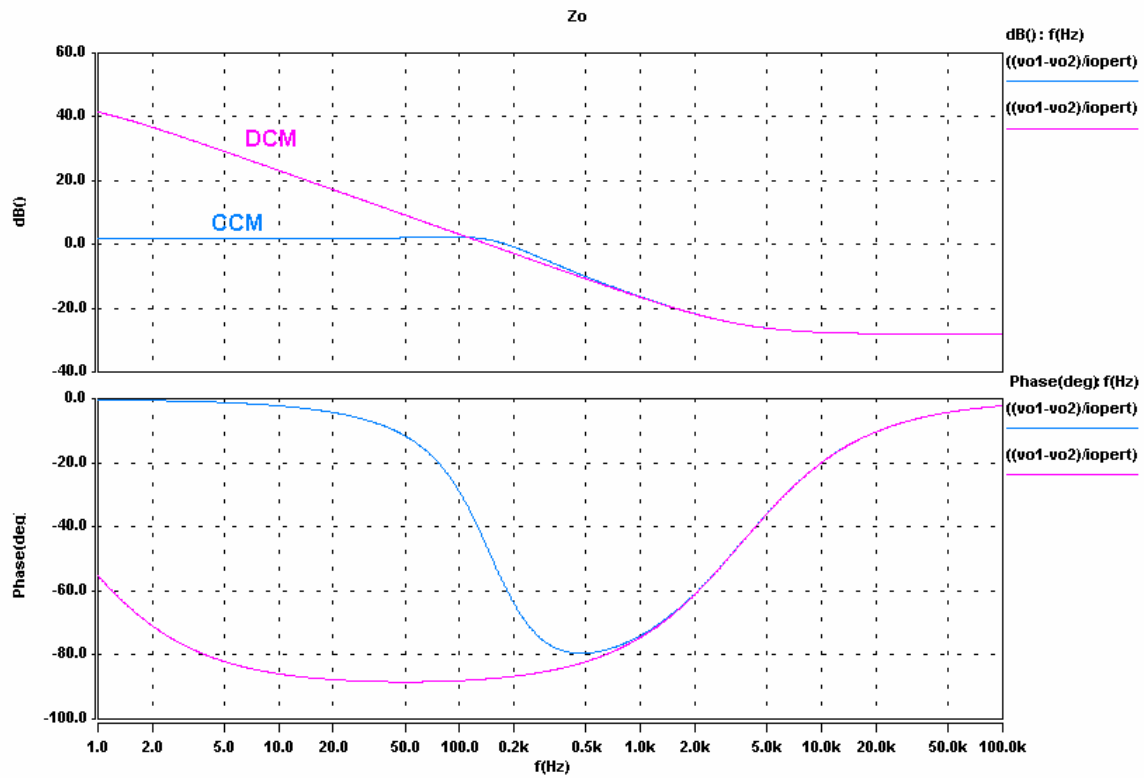


Figure 3.16 Simulated Output Impedance

### 3.2.2 Control Design

The majority of phase-shifted full-bridge converters are used to transition from high voltage to low voltage. One common example is their use in the telecom industry, where the input is usually a rectified utility resulting in a DC voltage of around 400 V and the output is 48 V. For applications like telecom, it is often necessary to have fast transient response and significant attenuation of the low frequency input noise, therefore requiring some type of current mode control. A common configuration is to have a fast inner current loop and a slower outer voltage loop. The current loop not only allows for fast response and low frequency rejection, but it also prevents transformer saturation that can arise from unbalanced excitation.

The front-end of the fuel cell inverter is basically the opposite of the telecom system, requiring the 48 V be boosted to 400 V. Furthermore, it is not necessary to have extremely fast transient response or to worry about the low frequency ripple on the output. In fact, a 120 Hz ripple will be created on the high voltage output once the inverter is running anyway. From this point of view, the current loop is not required for this application. However, the transformer saturation is still a concern, and therefore the current loop option is explored further.

Current sensing of a 5 kW DC-DC converter is not a trivial task and can involve significant added cost to the system. According to [36], there are many options for sensing the current of a phase-shifted converter. Current sensing resistors are not an option for this application, because of the high power levels. Due to the high input current, a hall sensor would be required for any primary side current sensing scheme. Although accurate, hall effect sensors are extremely expensive and often require a significant amount of power to operate. A current transformer (CT) on the secondary side is one possible way to sense the current. Although less expensive than the hall sensor, the CT is still quite costly at a 5 kW power level. For this application, it is desirable to avoid current sensing if possible.

Without current sensing, it is still possible to eliminate the DC component from the transformer by inserting a large capacitor in series with it [29]. However, with more than 100 A on the primary at full load, the size of that capacitor would be enormous and that option becomes unreasonable for this application. Therefore it is necessary to keep the

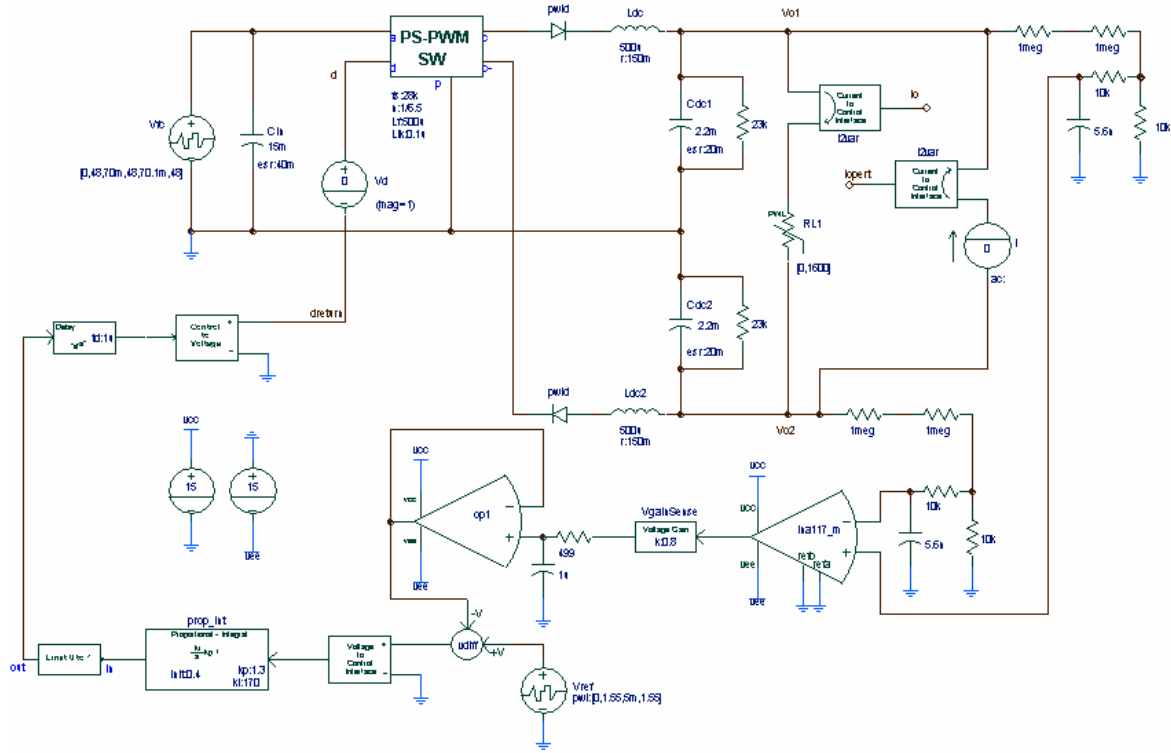
timing of the gate drives balanced and the layout of the circuit symmetrical to avoid transformer saturation. It is also advantageous to avoid full duty cycle operation at full load. This will assure that the transformer voltage always spends some time at zero voltage and can help evade the unbalanced excitation problem by allowing the transformer to reset.

There are many different control schemes for a converter of this nature, including state-space [37] and error compensation. The state-space methods result in great performance, but require knowledge of the inductor current and therefore involve current sensing. Moreover, state-feedback can involve extensive calculations and consume valuable processor time. Since high performance is not extremely important, it was determined that a simple voltage compensator should suffice as the control scheme. Although the implementation of the voltage compensator was done in the DSP, it was initially designed in the analog domain so that the transfer functions could be easily obtained through simulation. A proportional plus integral (PI) compensator of the form shown in (3.9) was used to close the voltage loop.

$$C(s) = K_p + \frac{K_i}{s} \quad (3.9)$$

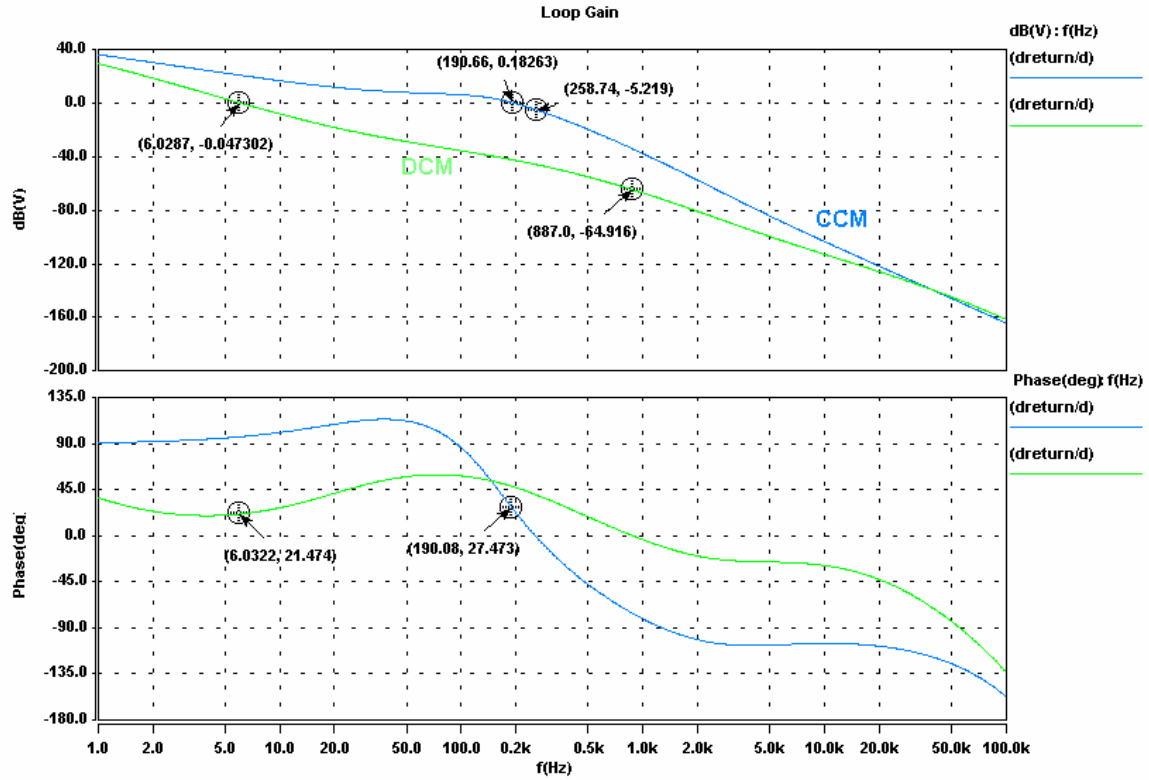
The integrator in the PI is used to generate zero steady-state error, while the zero ( $\omega_z = K_i/K_p$ ) is used to increase the phase at crossover and therefore increase the bandwidth. The open loop transfer functions in Figure 3.14 illustrate that the converter has an extremely different response in DCM versus CCM, which makes it challenging to design a compensator with good performance and yet stable in both operating regions. As explained in Chapter 2, the filter inductor was designed to make the CCM operating region as large as possible. However, due to the undetermined load, the converter may still operate in the DCM condition, therefore it is important to consider stability in both regions. Also, when designing the compensator, it is important to correctly model the sensing circuit and anti-aliasing filter so accurate parameter values can be chosen. The closed loop simulation circuit is shown in Figure 3.17. The optocoupler isolation is the only part not modeled correctly in Figure 3.17, as it is simply represented by a voltage gain block. A short delay is included in the feedback path to account for DSP

computation time and digital delay, and a limit is used to keep the duty cycle value less than one.



**Figure 3.17 Saber Closed Loop Analog Circuit of Front-End**

After several iterations, the ideal compensator parameters were determined to be  $K_p=1.3$  and  $K_i=170$ . The loop gain node plots for DCM and CCM are shown in Figure 3.18. In DCM, the bandwidth is extremely low at 6 Hz with a phase margin of  $21.5^\circ$  and a gain margin of 54.9 dB. This slow response in DCM is unavoidable when designing for a fast and stable response in CCM. The bandwidth in CCM is 190 Hz with a phase margin of  $27.5^\circ$  and a gain margin of 5.2 dB. These control characteristics are pushing the limits, but since it is a buck-type converter, it should not be a problem. The crossover frequency in CCM is not extremely fast, so a more complex compensator could be implemented. However, for this application 190 Hz should be adequate and a PI controller is easily realized in the DSP, so there is no need to make the compensator more complex.



**Figure 3.18 Simulated Closed Loop Gain of Front-End**

The closed loop audio susceptibility is shown in Figure 3.19, and the closed loop output impedance is shown in Figure 3.20. Both of these plots exhibit low frequency attenuation if compared to their open loop counterparts. However, the DCM output impedance plot is still significantly above 0 dB and may result in some variations in output voltage with varying load.

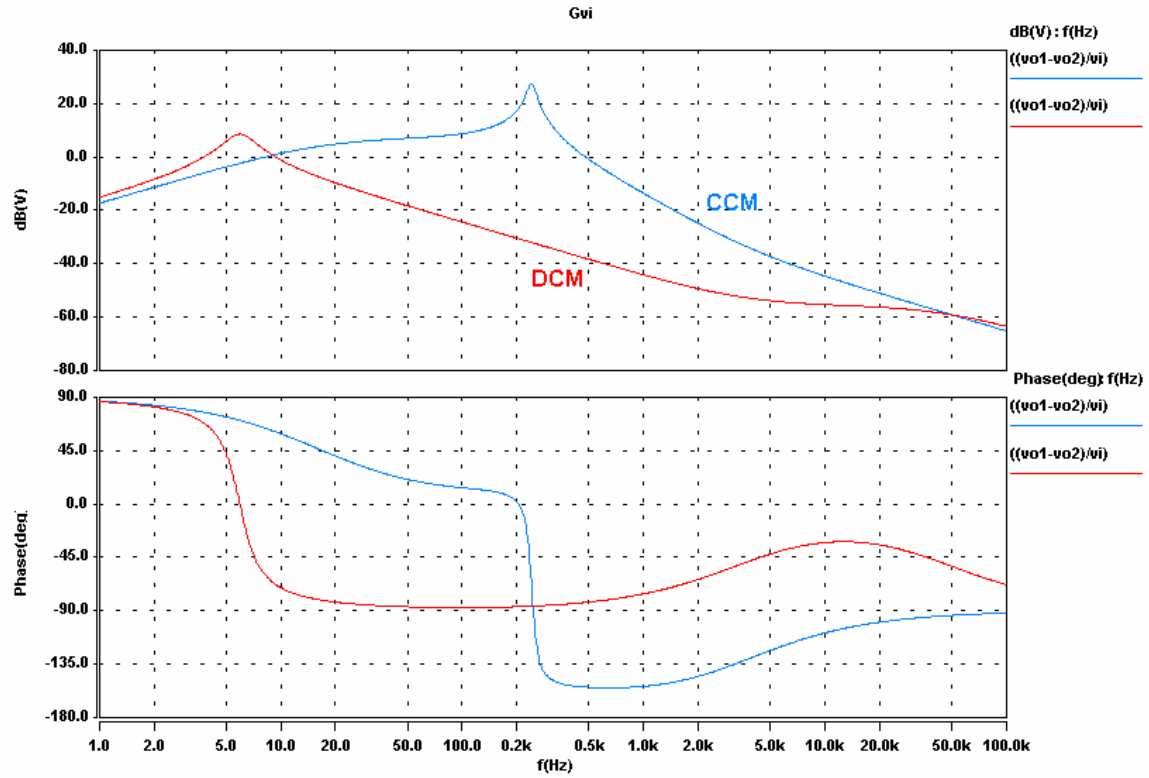


Figure 3.19 Simulated Closed Loop Audio Susceptibility

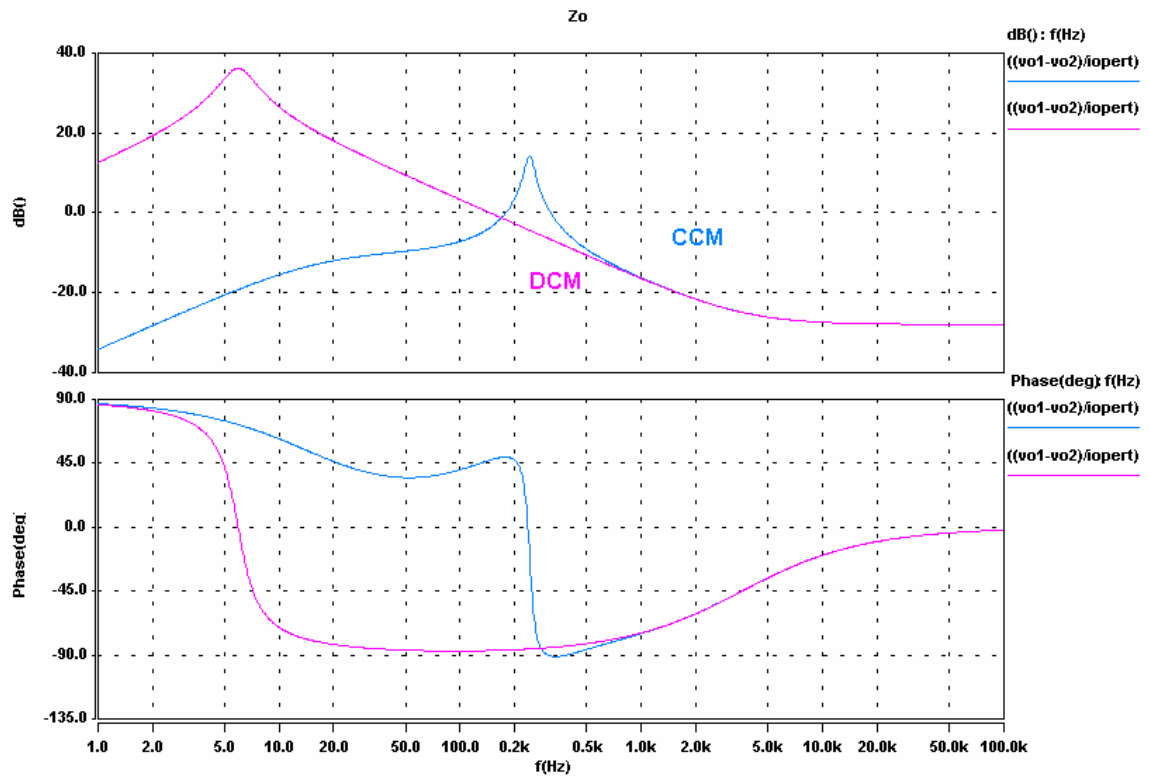
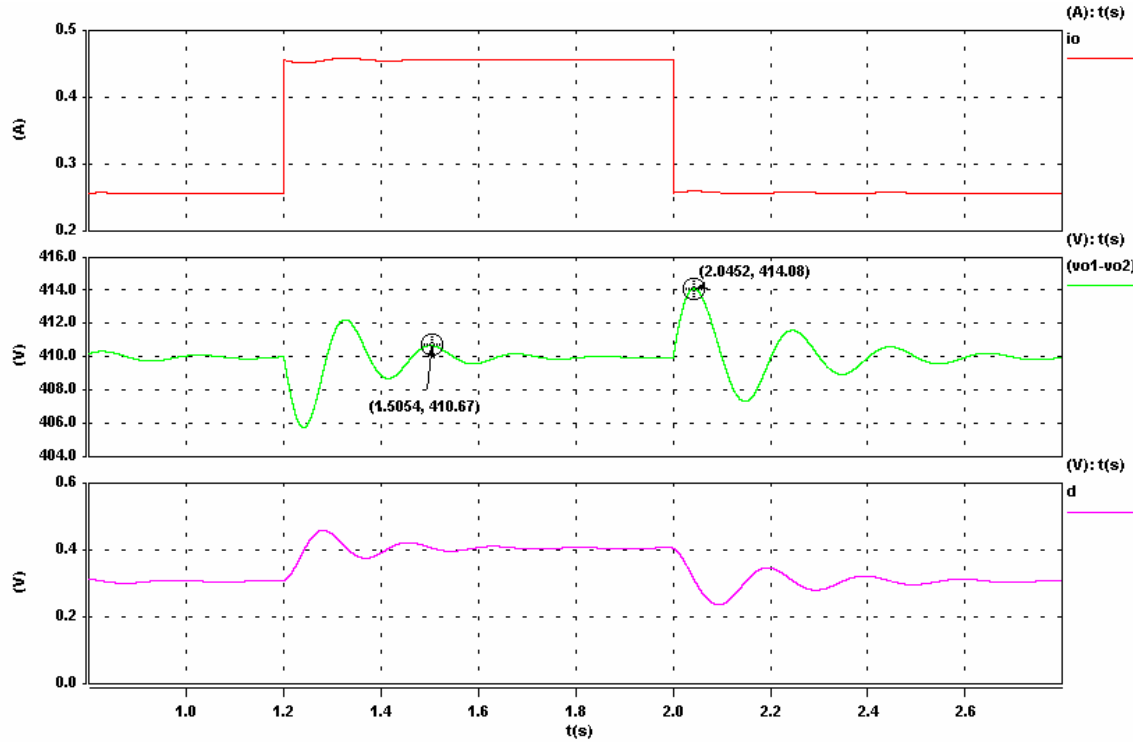


Figure 3.20 Simulated Closed Loop Output Impedance

To verify the response of the converter, load steps were performed in DCM and CCM conditions. The load step from 100 W to 200 W and back to 100 W is shown in Figure 3.21. As expected, this DCM transient is rather slow with a settling time of about 300 ms. Some oscillation does occur, but the overshoot is only 4 V or 1%, which is basically negligible.



**Figure 3.21 Simulated Load Transient in DCM Operation**

Figure 3.22 shows a load step from 2 kW to 4 kW and back to 2 kW. The CCM load transient is much faster, with less oscillation. The settling time is about 16 ms and the overshoot is the same as the DCM case at 4 V. An input voltage step of four volts, similar to what a fuel cell might produce during a transient, was simulated and the results are shown in Figure 3.23. The output voltage overshoot is still minimal at about 15 V or 4 %, and the settling time is approximately 25 ms.

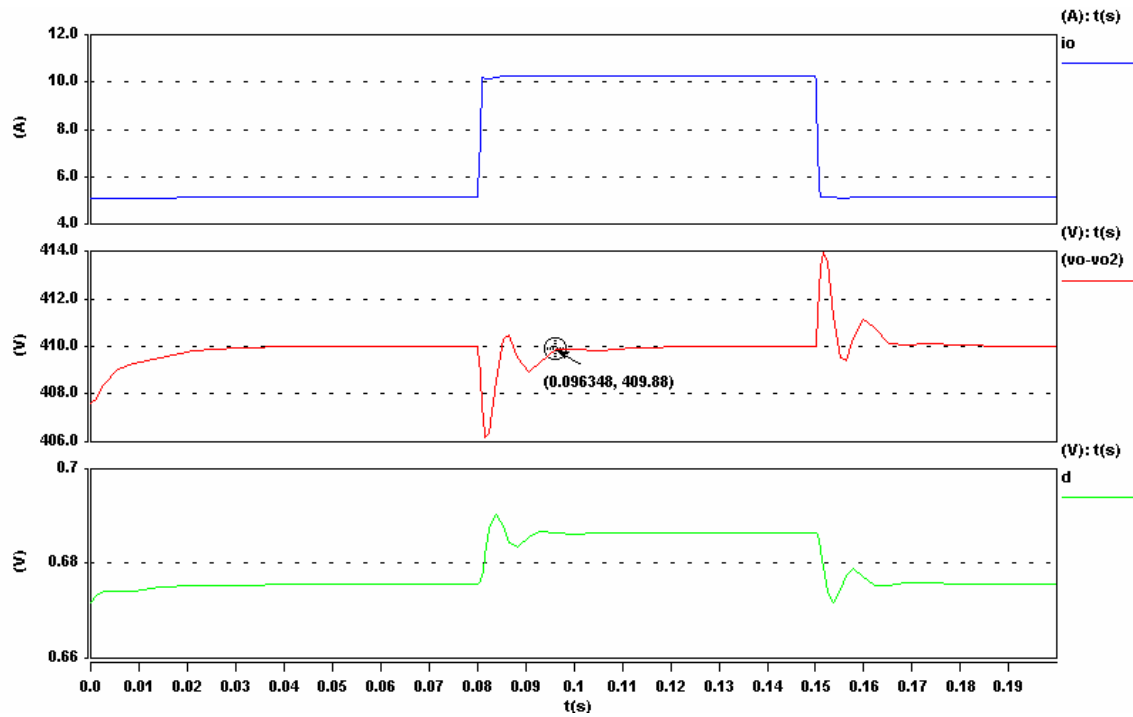


Figure 3.22 Simulated Load Transient in CCM Operation

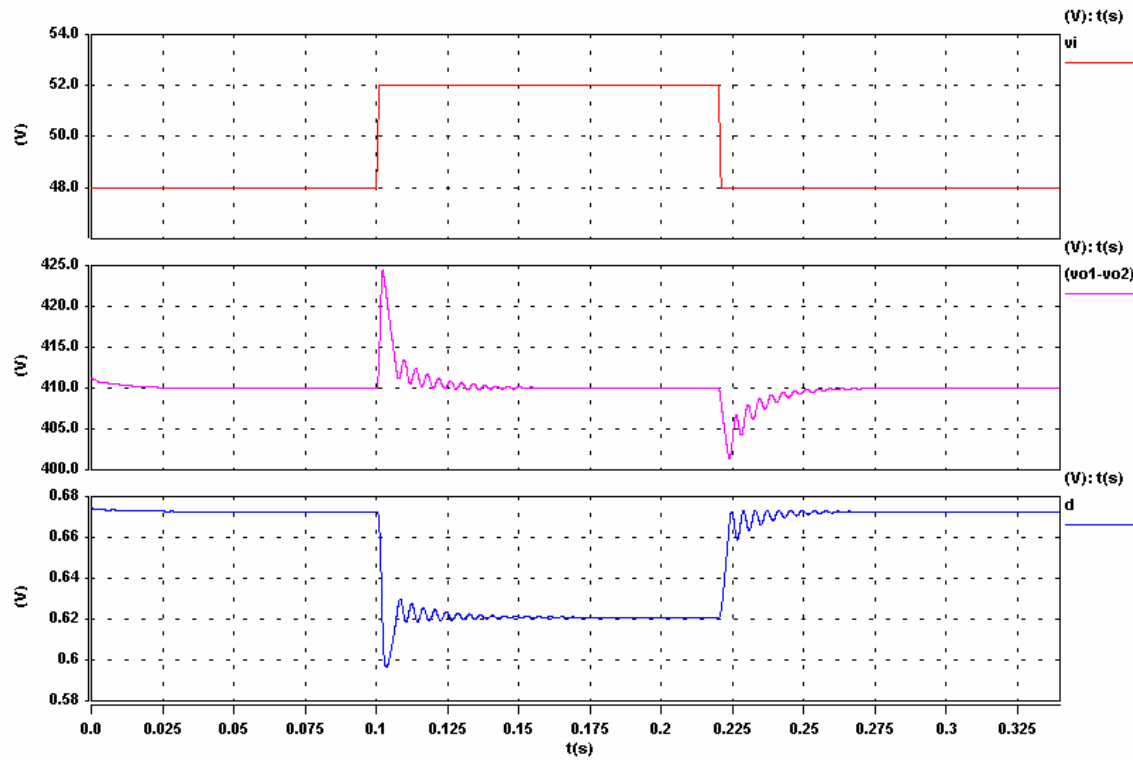
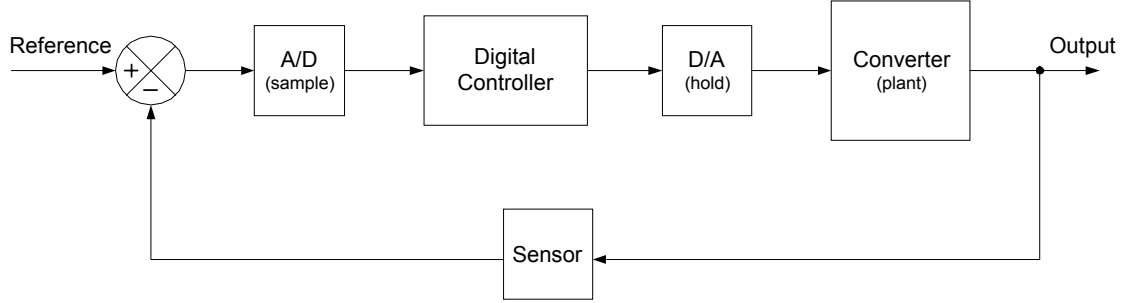


Figure 3.23 Simulated Input Voltage Step

The analog controller is a good starting point, however, the physical system is a mixture of analog and digital signals. The actual compensator is implemented in a DSP, which involves converting the feedback signal with an A/D, processing it, and then converting it back to analog with a D/A to send to the phase-shift control chip. This process is illustrated in the block diagram in Figure 3.24.

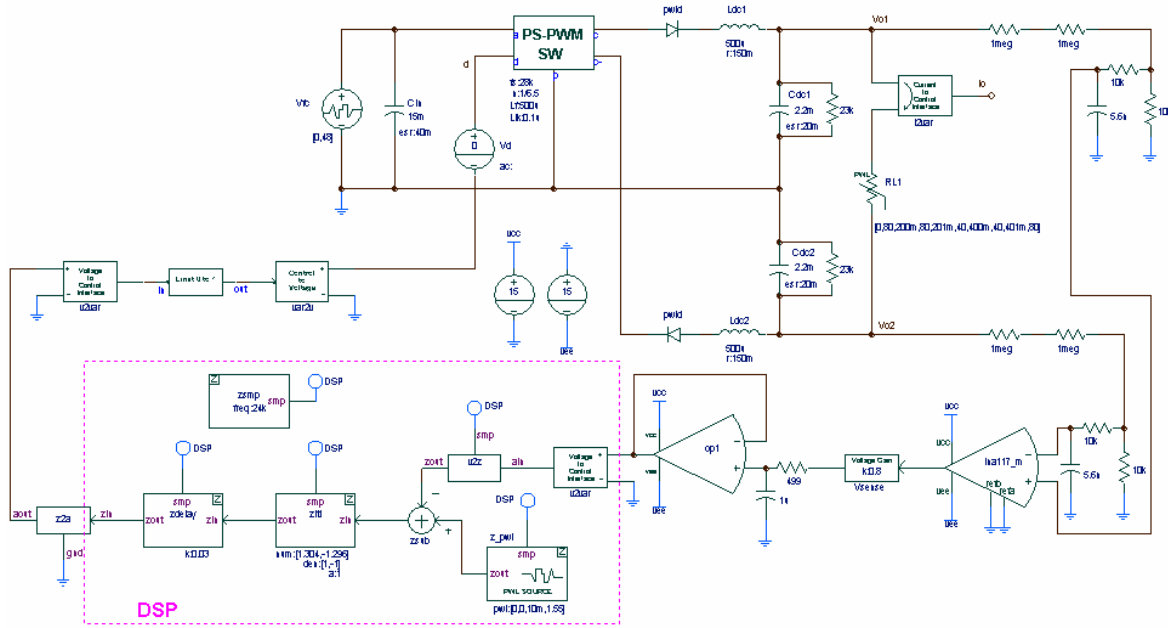


**Figure 3.24 Digital Control Block Diagram**

To accurately model a digital system using the  $z$  domain, a sampler and hold must be used. There are several types of data holds, but the most common is the zero-order hold (ZOH), which uses only the first term of the Taylor expansion of the analog signal. The first-order hold (FOH), also quite common, uses the first two terms of the Taylor expansion and may be more accurate depending on the frequency range of the signal and sampler [38]. For this application the DSP is sampling at 24 kHz and the feedback signal is DC, so a FOH should be beneficial. The continuous compensator in (3.9) must be converted to the digital domain, which is done by taking the inverse Laplace transform, converting to discrete time and then taking a  $z$ -transform (easily done in Matlab). The resultant discrete time compensator with FOH incorporated is shown in (3.10), where  $T_s$  is the sampling time of the DSP.

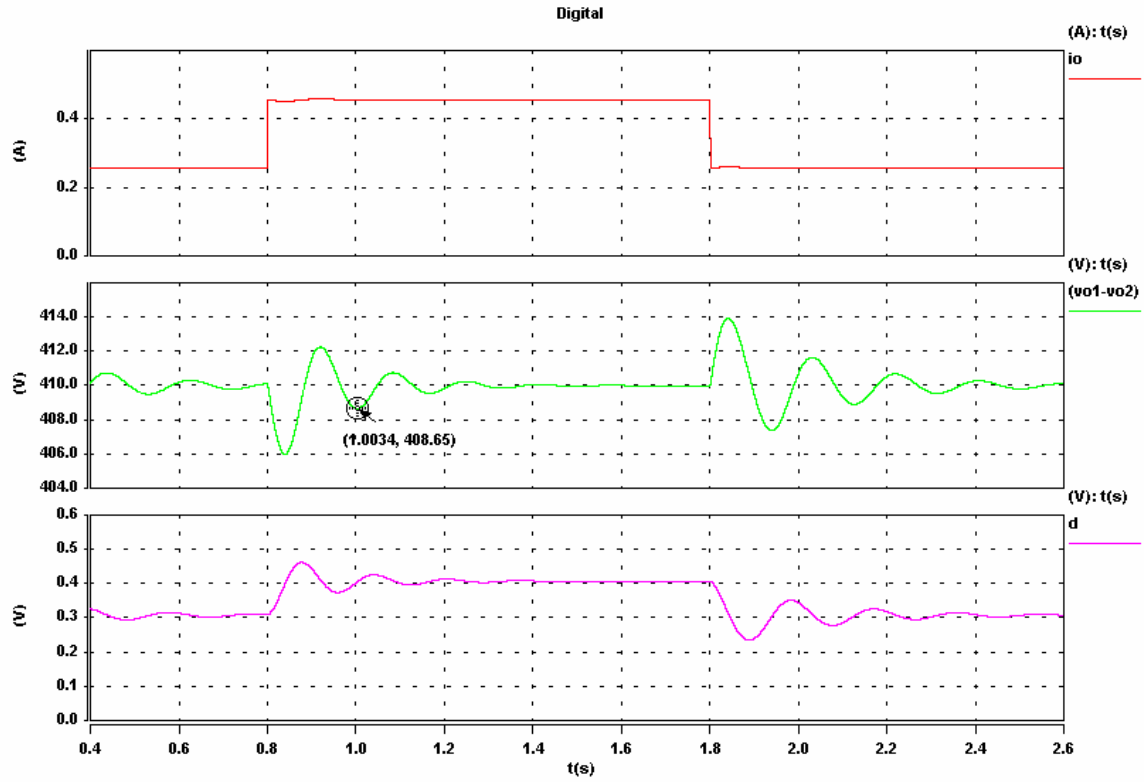
$$C_z(z) = K_p + K_i \frac{T_s}{2} \frac{z+1}{z-1} \quad (3.10)$$

The model is implemented in Saber as shown in Figure 3.25, where the v2z block is the sampler and the zlti block is the compensator. A piecewise linear source is used as the voltage reference to provide a soft-start command. Once again, a small delay was incorporated in the loop to account for computations during the DSP interrupt cycle.

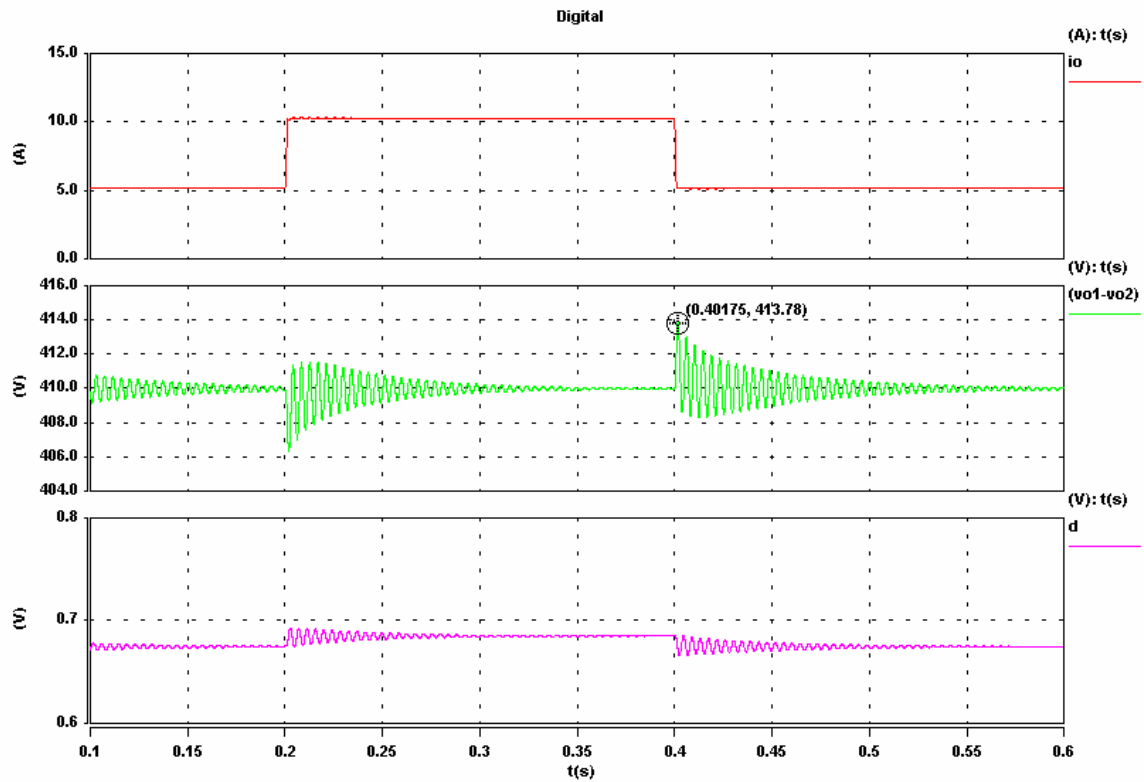


**Figure 3.25 Saber Closed Loop Digital Circuit for Front-End**

A closed loop bode plot is not easily obtainable from the digital simulation model. Therefore, only transient analysis was performed to verify the system stability with the digital compensator. Figure 3.26 shows the same DCM load step in discrete time as simulated in the continuous time case, and the performance is almost identical. On the other hand, the digital controller in the CCM condition results in some small oscillations during load transients. The overshoot is the same as the continuous case, but the settling time is increased due to the oscillations. Since the compensator was designed with a relatively small phase margin in CCM, the digital delay seemed to decrease the stability enough to cause it to oscillate during transient conditions. Decreasing the gain of the compensator slightly (to  $K_p=0.9$ ) can significantly reduce the oscillations without increasing the overshoot much.



**Figure 3.26 Simulated Load Transient in DCM Operation with Digital Control**



**Figure 3.27 Simulated Load Transient in CCM Operation with Digital Control**

### 3.2.3 Experimental Verification

In attempt to verify the model and lend insight into the system, transfer functions of the DC-DC converter were measured using an impedance analyzer. Details on the various measurement set-ups can be found in Appendix E.

Before the compensator was implemented, several open loop transfer functions were measured. In each case, the simulated version is also shown for comparison. In some cases, scaling of the simulated signal was required to account for the gain of the differential probe used. The control to output response with the converter operating in DCM (100 W) is given in Figure 3.28. There is a little bit of noise on the measured signal because the signal to noise ratio at light loads is smaller, and the impedance analyzer's averaging function may not have been set high enough. Despite the deviation at high frequencies, the simulation matches the experimental bode plot well. It should be noted that the validity of the average model simulation is questionable beyond about half of the switching frequency, or about 14 kHz. This high frequency region is also affected by parasitics that were probably ignored in the simulation model.

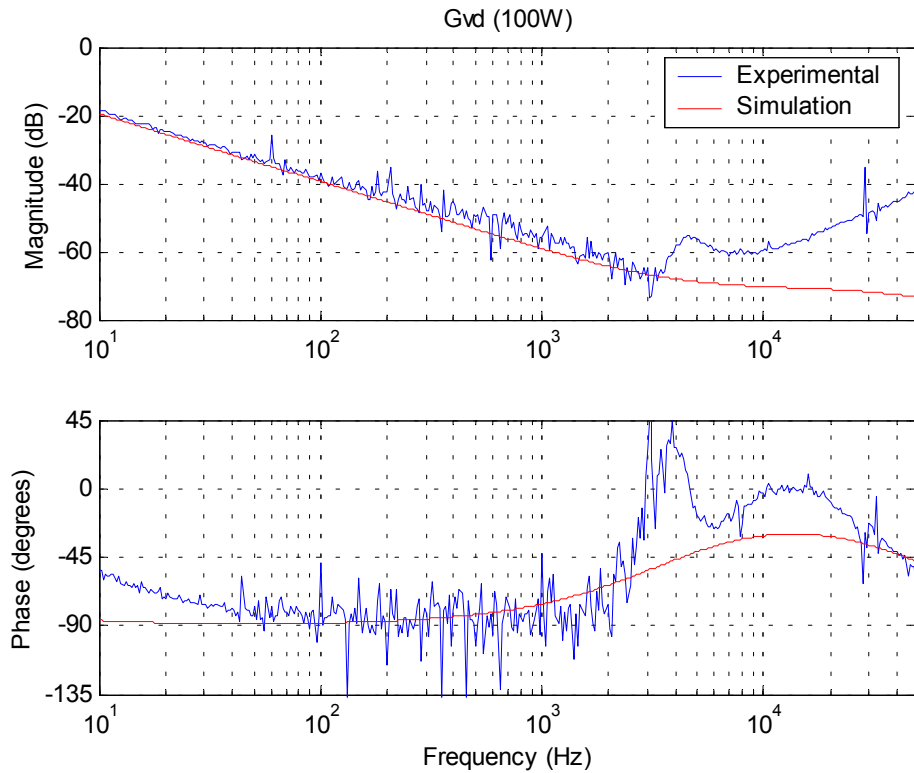
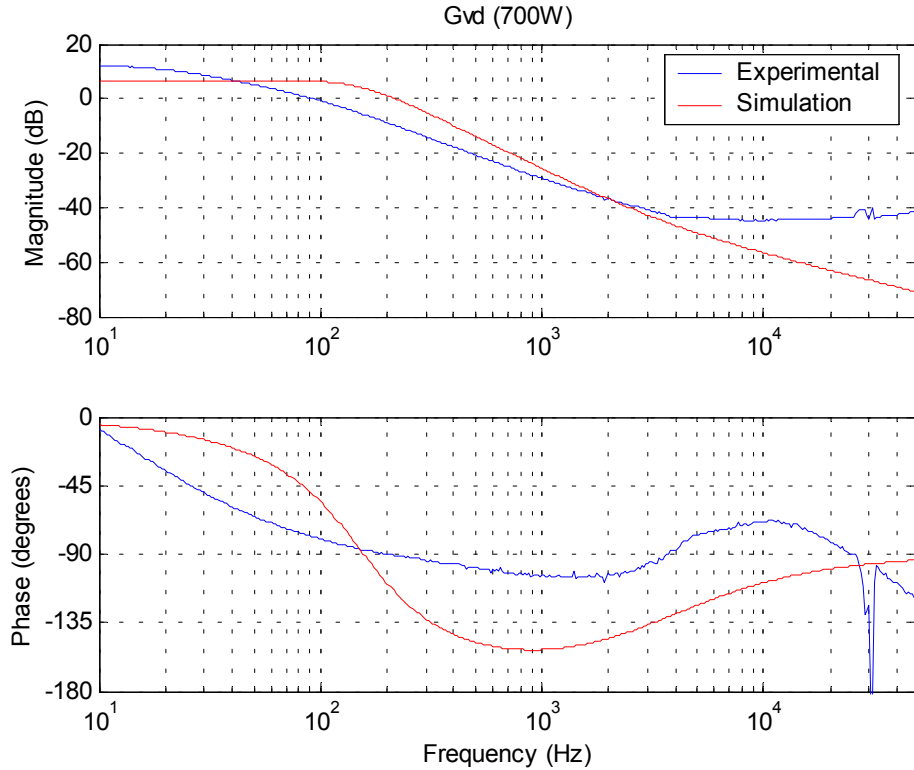


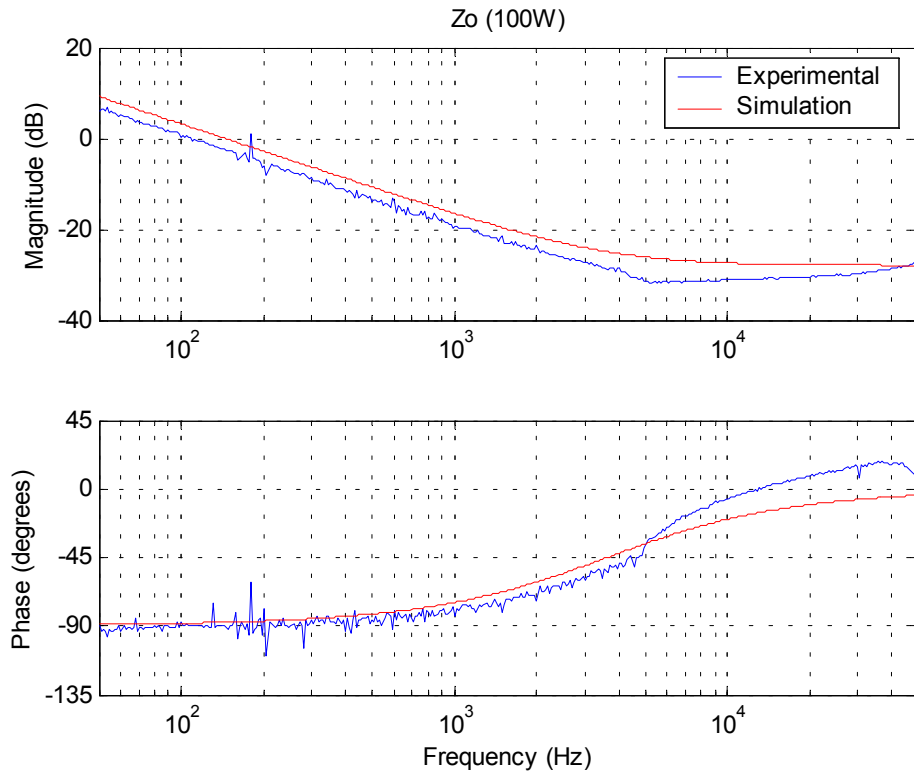
Figure 3.28 Experimental Control to Output Transfer Function in DCM

The control to output response in CCM operation of 700 W is shown in Figure 3.29. The measured transfer function begins to roll-off before the simulated version and the slope is less steep. This indicates that the converter sees a lower filter resonance than anticipated.

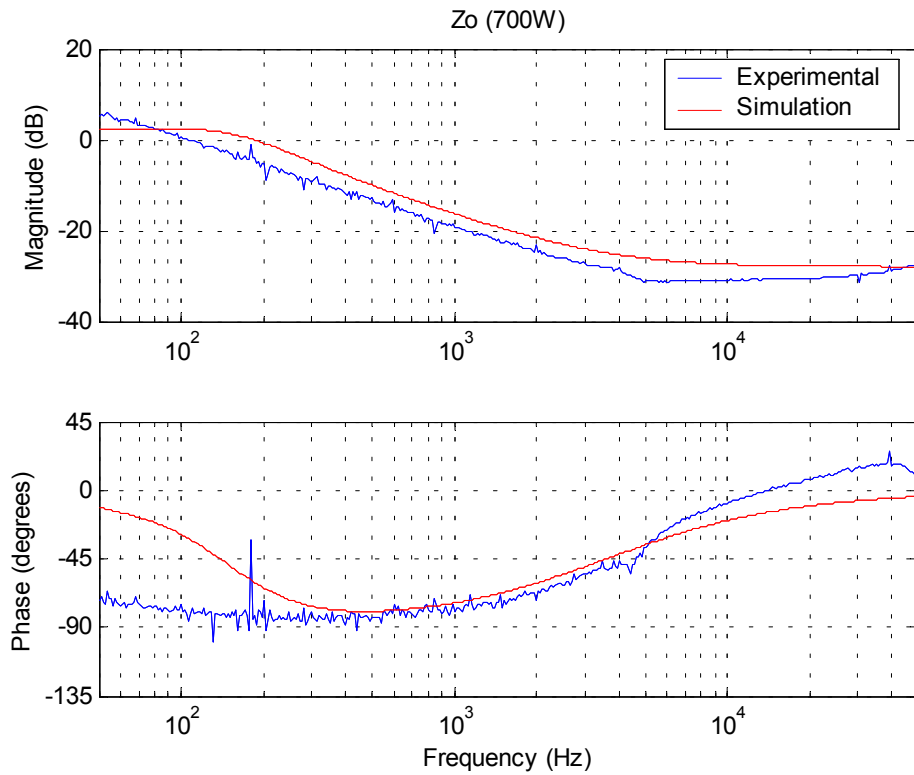


**Figure 3.29 Experimental Control to Output Transfer Function in DCM**

The open loop audio susceptibility and output impedance of the converter were also measured. The results for open loop output impedance are shown in Figure 3.30 and Figure 3.31. The output impedance measurement was somewhat complicated as explained in E.3, and the isolation transformer used did not work well at low frequencies. Consequently, the starting frequency was limited to 50 Hz. For the DCM case, the experimental and simulation results match almost perfectly. In CCM, the results are close except that the cut-off point occurs at a lower frequency in the experimental case; unfortunately the exact point was not able to be determined. This difference is the same phenomena that occurred in Figure 3.29, and suggests that the LC filter is much larger than it actually is.

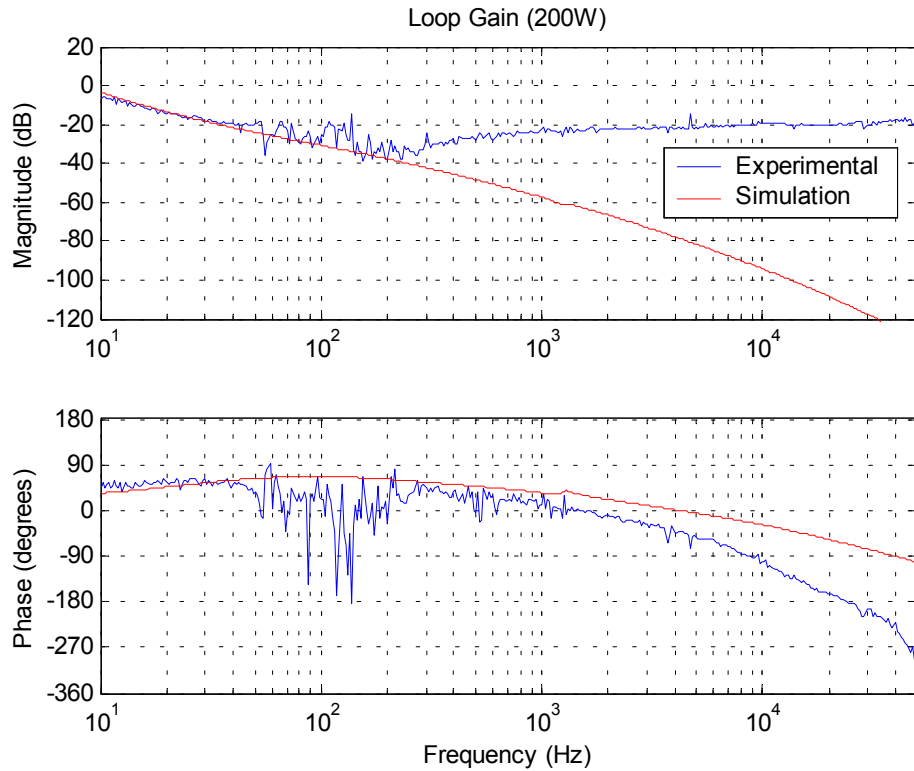


**Figure 3.30 Experimental Open Loop Output Impedance in DCM**



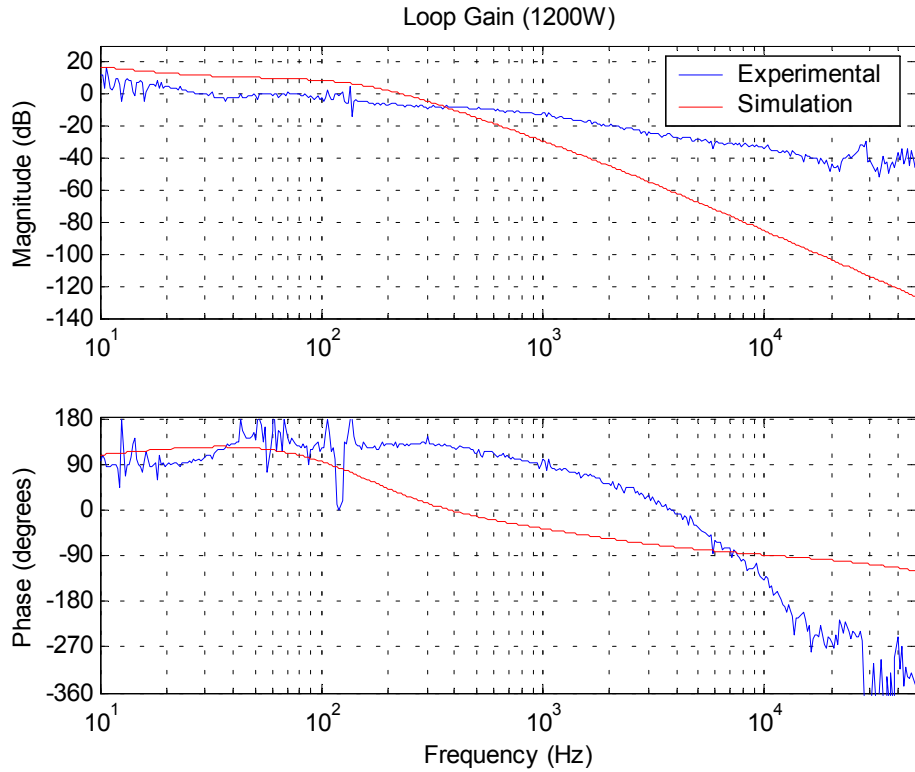
**Figure 3.31 Experimental Open Loop Output Impedance in CCM**

Judging from the open loop measurements, it was determined that the model was reasonably accurate and the designed PI compensator was implemented in the DSP to close the loop. After verifying that the converter was stable at several operating points, closed loop transfer functions were measured. Initially, a gain of  $K_p=0.9$  was implemented in the DSP, but the crossover was lower than expected so the gain was increased back to the original design value of  $K_p=1.3$ . The experimental loop gain results for DCM and CCM are shown in Figure 3.32 and Figure 3.33.



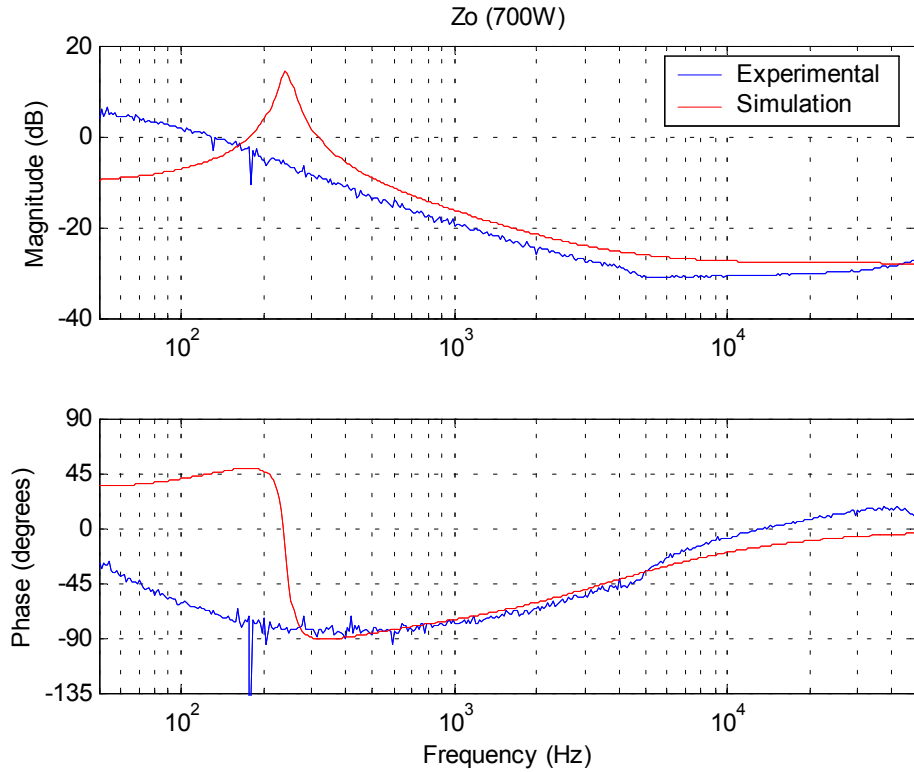
**Figure 3.32 Experimental Closed Loop Gain in DCM**

The DCM experimental result agrees with the simulation exactly at low frequencies, but the gain begins to deviate around 300 Hz, although the phase follows closely. This result implies that there is a right half plane (RHP) zero, however this does not seem reasonable and could be a measurement problem. The loop gain result for CCM condition is given in Figure 3.33. The crossover frequency is slightly lower than anticipated, but the puzzling phenomena again occurs around 300 Hz, where it seems as though there is a pole missing.



**Figure 3.33 Experimental Closed Loop Gain in CCM**

The closed loop audio susceptibility and output impedances were also measured, however only the CCM output impedance is shown here. The DCM plots are the same as the open loop cases; this result is to be expected because closing the loop in DCM only affects the very low frequency portion of the curve, which can't be measured accurately. The closed loop output impedance, in Figure 3.34, again demonstrates that the resonant frequency is lower than the simulation predicts. When comparing the experimental plot in Figure 3.34 to the open loop version, the low frequency portion begins to show some deviation, but the major difference seems to be below 50 Hz and therefore could not be measured.



**Figure 3.34 Experimental Closed Loop Output Impedance in CCM**

Overall the experimental bode plots are reasonable, but there seems to be a common theme present in the majority of them. The output filter resonance point seems to be almost a decade below where it should be.

The transient response of the DC-DC converter was also measured in DCM and CCM conditions. Figure 3.35 shows the output voltage during a small load step from 120 W to 200 W. A CCM load step from 600 W to 1.2 kW is shown in Figure 3.36. The output voltage was AC coupled into the scope so the transient could easily be measured. As expected, the DCM response is considerably slower than the CCM response. For both cases, the voltage droop is less than 5 V, which is minimal as predicted. The ringing that is evident in the digital simulation is not present in the actual response, even with a  $K_p$  value of 1.3. This effect agrees with the lower crossover frequency shown in the experimental transfer function measurement of Figure 3.33.

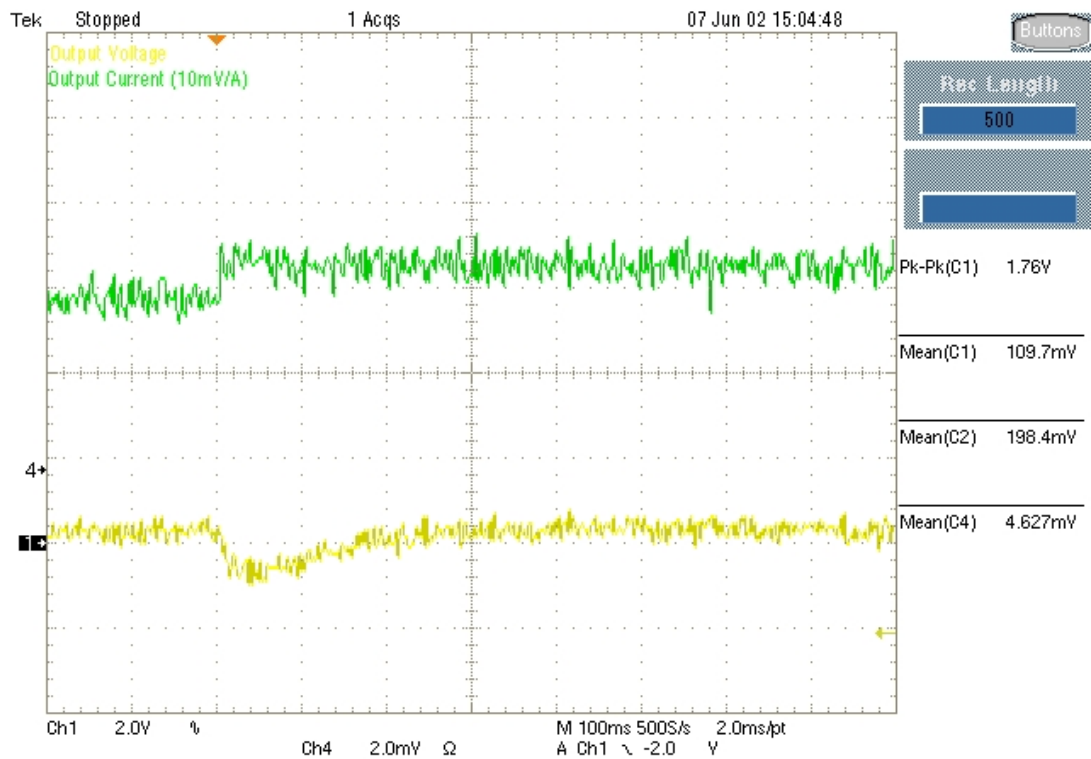


Figure 3.35 Experimental Load Transient in DCM Operation

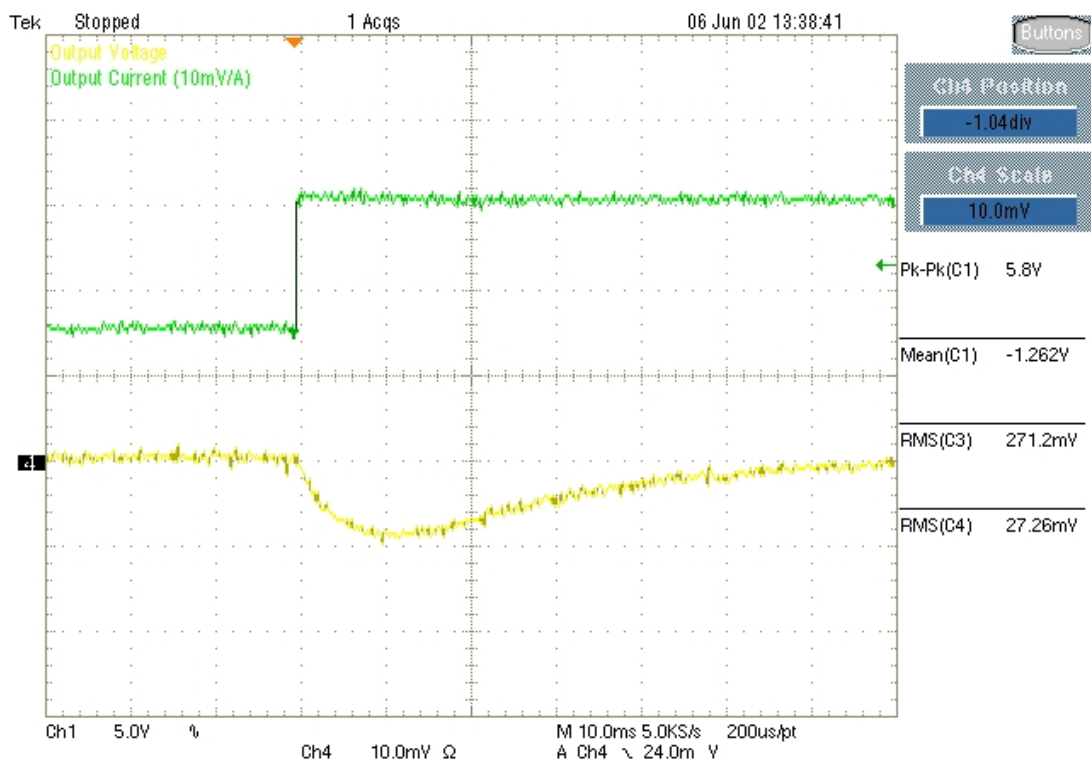
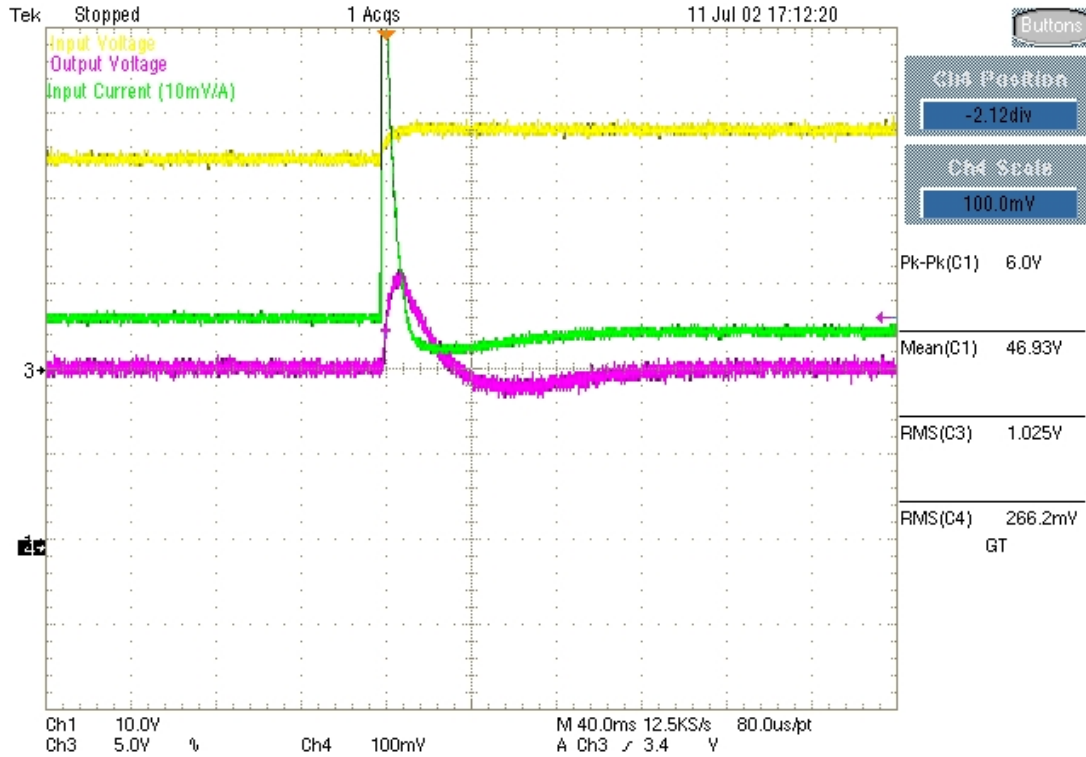


Figure 3.36 Experimental Load Transient in CCM Operation

The output voltage response to an input voltage step is displayed in Figure 3.1. The input voltage was increased approximately four volts, and the resultant output voltage overshoot was six volts with a settling time of about 40 ms. The performance is actually better than the simulation predicts.



**Figure 3.37 Experimental Input Voltage Step Response**

Overall the converter works well; however, it regulates to a slightly higher voltage in DCM than in CCM. This steady-state error is due to the lower DC gain in DCM. This problem did not present itself in the simulations. The output voltage variation can be resolved with the inverter control strategy. Table 3.1 summarizes the results of the control loop and the transient performance.

**Table 3.1 Simulation vs. Experimental Results**

| Characteristic | Simulation |        | Experiment |       |
|----------------|------------|--------|------------|-------|
|                | DCM        | CCM    | DCM        | CCM   |
| Phase Margin   | 21.5°      | 27.5°  | 22°        | 100°  |
| Gain Margin    | 64.9 dB    | 5.2 dB | 24 dB      | 28 dB |
| Overshoot      | 1 %        | 1 %    | 0.5 %      | 1.2 % |
| Settling Time  | 300 ms     | 80 ms  | 160 ms     | 60 ms |

### 3.3 Inverter

#### 3.3.1 Modeling

In order to evaluate the control options for the half-bridge inverter, it was also accurately modeled. The switching model of the half-bridge inverter, shown in Figure 3.38, is much simpler than the switching model of the DC-DC converter. The Saber model consists of ideal switches and diodes with parasitic capacitances for the IGBTs. Comparing a high frequency triangle wave to a 60 Hz sine wave generates the complementary logic gate signals for the switches. The RDC circuit is used to introduce dead time between the gate signals.

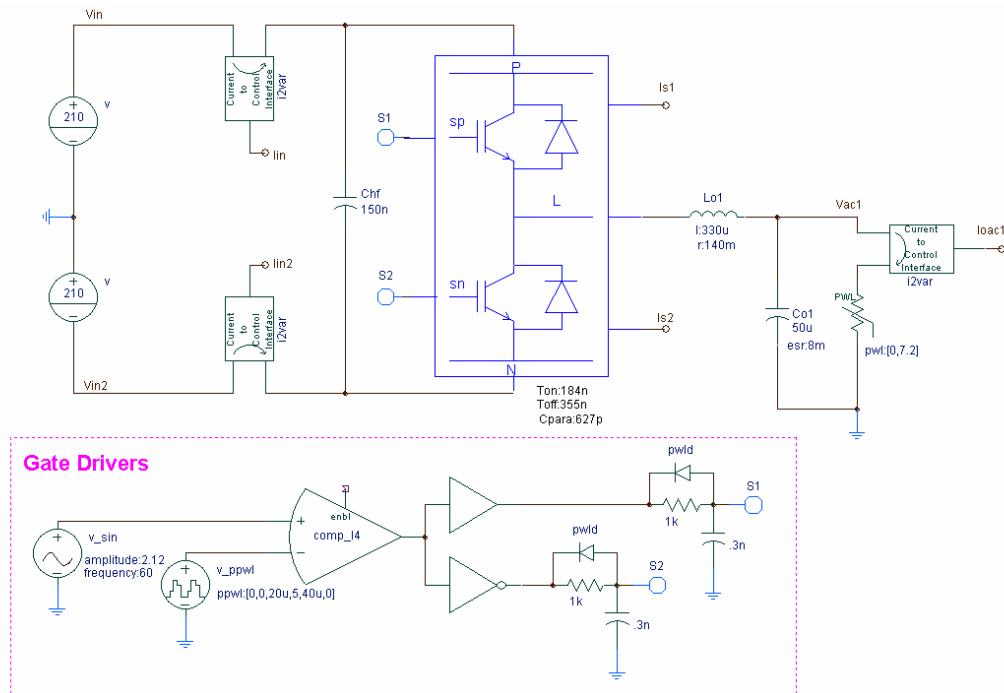


Figure 3.38 Saber Switching Model of Inverter

Simulations were run to verify the simple operation of the circuit at various load conditions. Figure 3.39 shows the input current and the output voltage and current for a load of 2 kW. The device voltages and currents for the same load are shown in Figure 3.40.

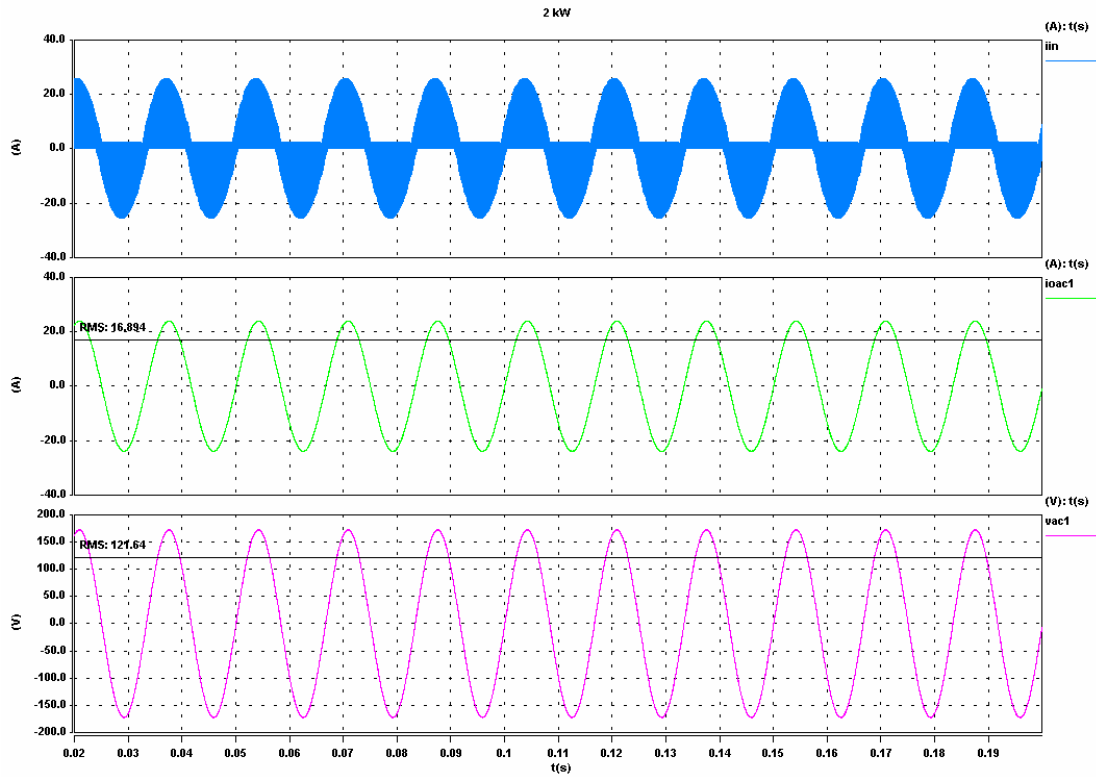


Figure 3.39 Simulated Inverter Output Waveforms at 2 kW

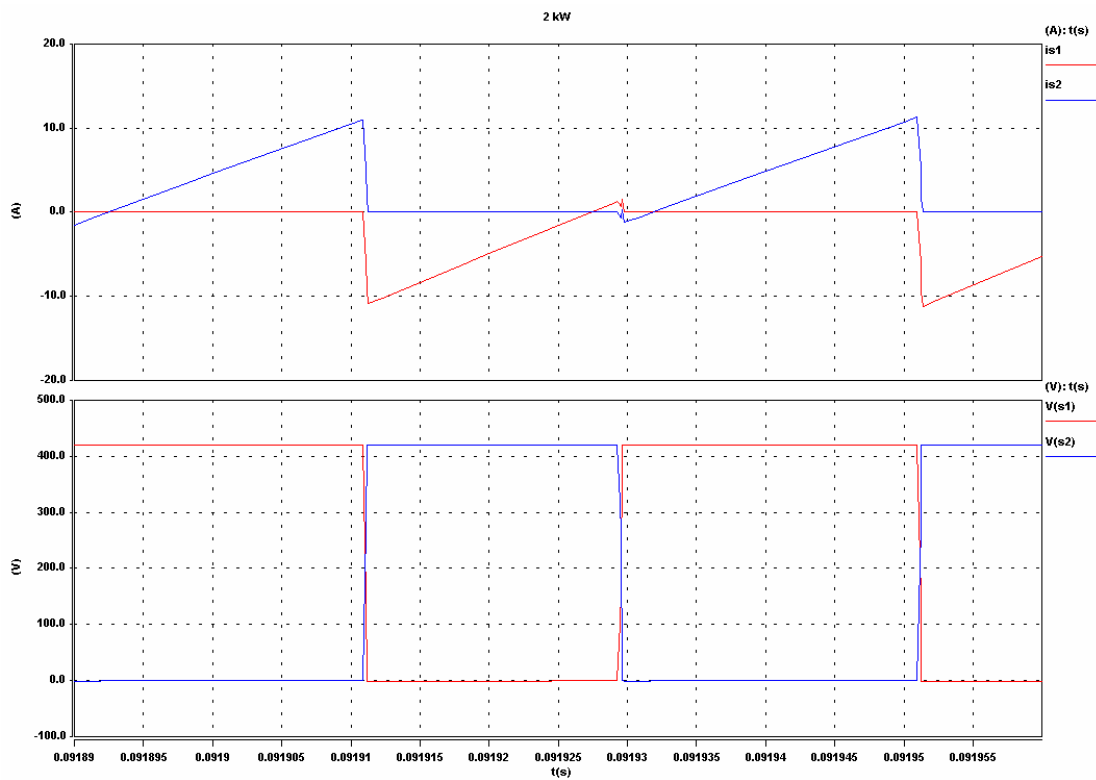


Figure 3.40 Simulated Inverter Device Waveforms at 2 kW

As with the DC-DC converter, it is desirable to create an average model of the inverter for design of the control loop and to decrease simulation time. Assuming that the switching frequency is much larger than the line frequency, the switching function can be averaged as shown in (3.11), where  $m$  is the modulation index and  $\omega$  is the line frequency.

$$d(t) = \frac{1}{2}(m \sin(\omega t) + 1) \quad (3.11)$$

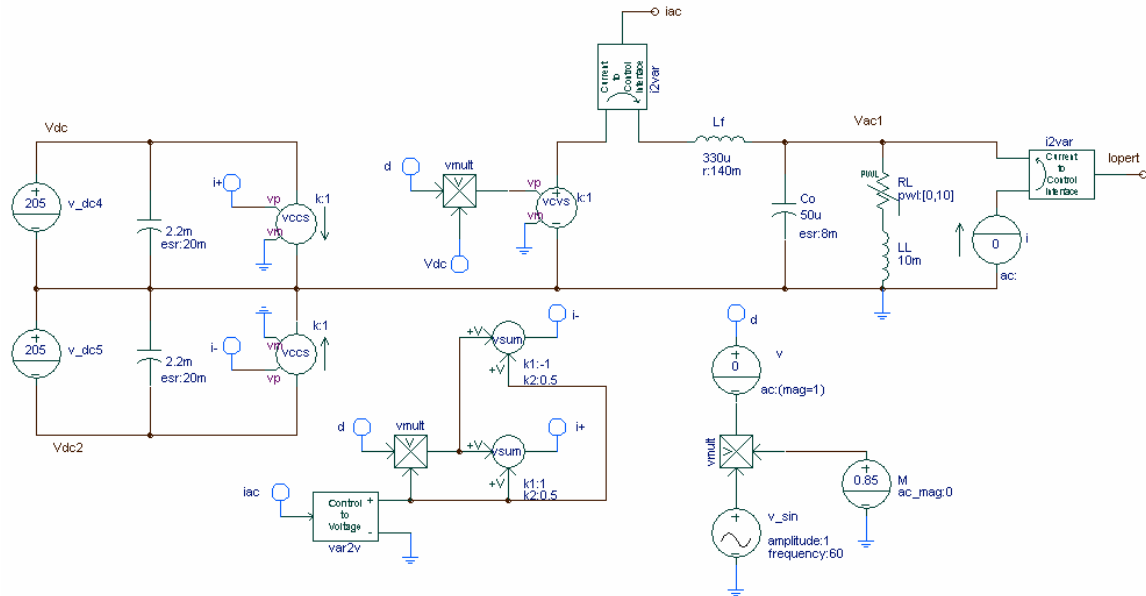
Now an averaged set of differential equations can describe the system as shown in (3.12), where  $L_f$  is the filter inductance,  $R_f$  is the inductor filter resistance, and  $C_o$  is the output capacitance [40].

$$\frac{d}{dt} \begin{bmatrix} i_f(t) \\ i_o(t) \\ v_o(t) \end{bmatrix} = \begin{bmatrix} -R_f/L_f & 0 & -1/L_f \\ 0 & R_L/L_L & 1/L_L \\ 1/C_o & -1/C_o & 0 \end{bmatrix} \begin{bmatrix} i_f(t) \\ i_o(t) \\ v_o(t) \end{bmatrix} + \begin{bmatrix} \frac{mV_{dc} \sin(\omega t)}{L_f} \\ 0 \\ 0 \end{bmatrix} \quad (3.12)$$

Creating a small signal model for a single-phase inverter is not that intuitive since the output is constantly varying with time. This makes it difficult to linearize the equations around a specific operating point. However, small signal models have been developed by the perturbation and approximation method [40]. The control (modulation index) to output voltage transfer function is shown in (3.13). This transfer function does not include the output capacitor ESR.

$$\frac{\hat{v}_o}{\hat{m}} = \frac{\frac{V_{dc}}{L_f C_o} \left( s + \frac{R_L}{L_L} \right)}{s^3 + \left( \frac{R_L}{L_L} + \frac{R_f}{L_f} \right) s^2 + \left( \frac{R_L R_f}{L_L L_f} + \frac{L_L + L_f}{L_L L_f C_o} \right) s + \frac{R_L + R_f}{L_f L_L C_o}} \quad (3.13)$$

Similar to what was developed in [40] and [41], an equivalent circuit was created and implemented in Saber as shown in Figure 3.41. The input current, although not really useful in this model, was added for completeness and will be important when building the average model of the entire system.



**Figure 3.41 Saber Open Loop Average Circuit of the Inverter**

Using the Saber model, the control to output transfer function and the output impedance were simulated for a single dc operating point. The results are shown in Figure 3.42 and Figure 3.43. By varying the phase of the sinusoidal reference, frequency analysis can be done for different dc operating points. Several points along the sine wave cycle were chosen and the results are exactly the same as given in the figures on the follow page. Although not ideal, this method does provide some insight into the dynamics of the system. The output impedance of the converter is important for this application, because the load is completely unknown. Figure 3.43 shows that the output impedance around 60 Hz is sufficiently attenuated, and that the full load case results in the largest peaking due to minimal damping by the load.

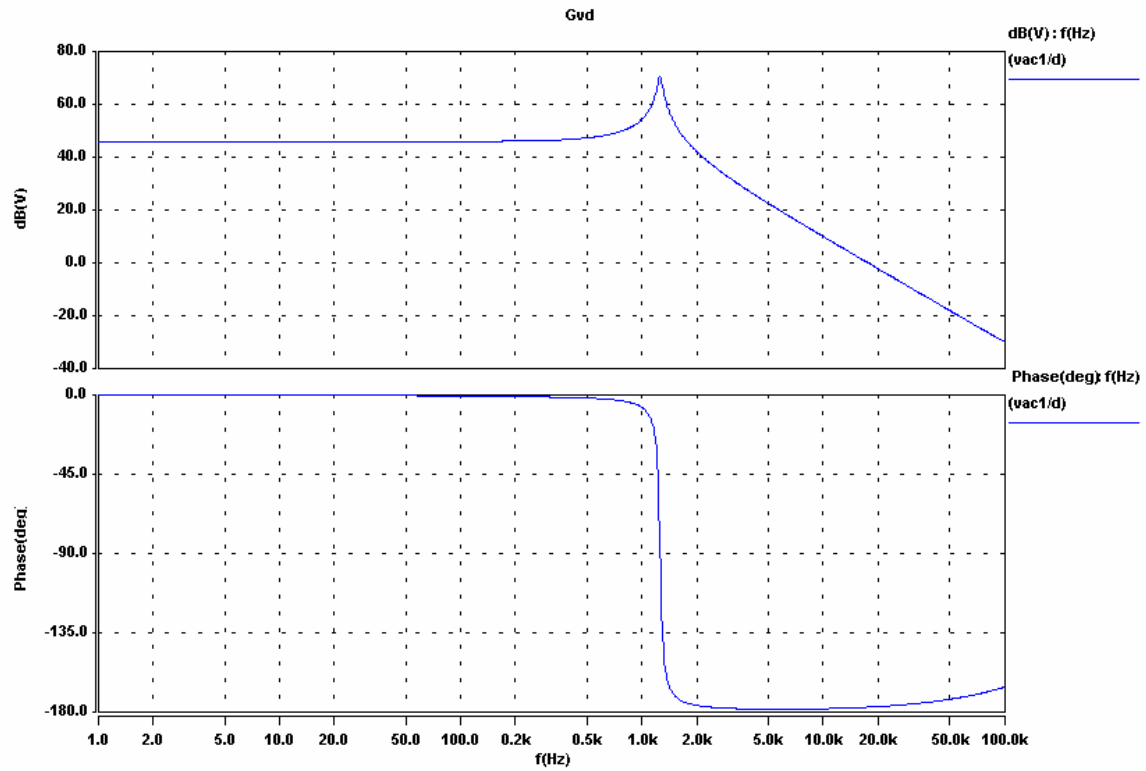


Figure 3.42 Simulated Control to Output Transfer Function of the Inverter

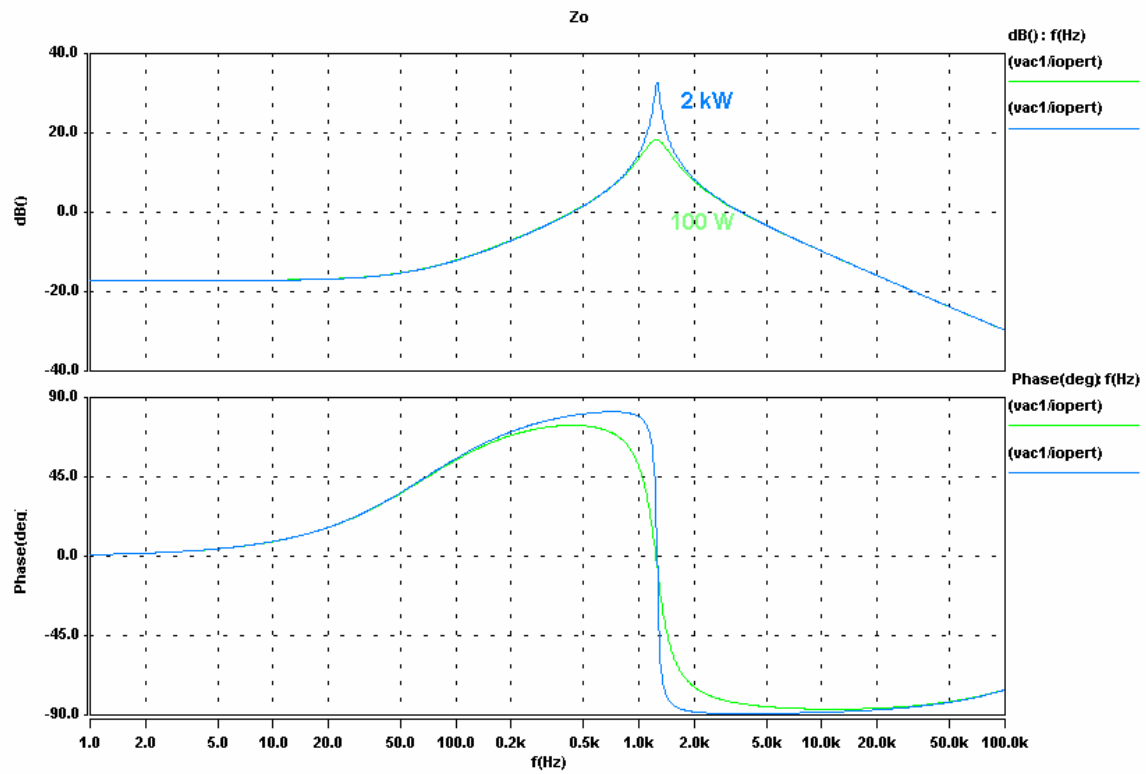


Figure 3.43 Simulated Open Loop Output Impedance of the Inverter

### 3.3.2 Control Design

Unlike the DC-DC converter, the inverter can operate open loop, which is clearly the simplest and lowest cost control method. Assuming that the DC input voltage is fixed and there are no parasitic losses in the circuit, then ideally the inverter could be operated with a fixed modulation index and therefore a fixed duty cycle. For purely resistive loads the output current does not affect the output voltage of an inverter. Of course there are parasitic losses in the devices and the passive components, but is it still possible to maintain the AC output voltage within the ITI/CBEMA specifications (Appendix A) over the entire range of load conditions?

Theoretically, the voltage drop of the inverter from no load to full load can be calculated. Using a first order loss model of an IGBT [42], the conduction losses for the switch and the diode can be calculated as a function of current as shown in (3.14) and (3.15).

$$V_{sw}(I_o) = V_{CEsat} \left( \frac{1}{\pi} + \frac{m}{4} \cdot pf \right) + I_o R_{CE} \left( \frac{\sqrt{3}}{8\sqrt{\pi}} + \frac{m}{3\pi} \cdot pf \right) \quad (3.14)$$

$$V_D(I_o) = V_f \left( \frac{1}{\pi} - \frac{m}{4} \cdot pf \right) + I_o R_{ak} \left( \frac{\sqrt{3}}{8\sqrt{\pi}} - \frac{m}{3\pi} \cdot pf \right) \quad (3.15)$$

These equations assume sinusoidal PWM waveforms, where  $m$  is the modulation index and  $pf$  is the power factor of the load. The other significant loss is due to the output filter inductor. The inductor voltage drop as a function of output current is shown in (3.16), where  $\omega$  is the line frequency (377 rad/sec).

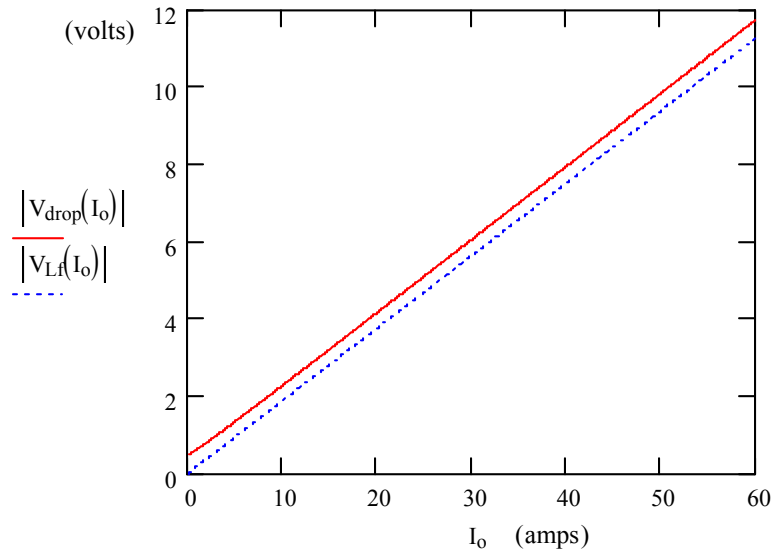
$$V_{Lf}(I_o) = (R_f + j\omega L_f) I_o \quad (3.16)$$

Adding (3.14), (3.15), and (3.16) together, the total voltage drop as a function of load can be determined and is given in (3.17).

$$V_{drop}(I_o) = V_{sw}(I_o) + V_D(I_o) + V_{Lf}(I_o) \quad (3.17)$$

Plugging in values from the IGBT datasheet, with a modulation index of 0.85 and a power factor of 1, the peak voltage drop can be plotted as a function of peak output current as demonstrated in Figure 3.44. As seen from the graph, the inductor (shown as

dotted line) accounts for the majority of the total voltage drop. Therefore varying the power factor and the modulation index has little effect on the total voltage drop.



**Figure 3.44 Inverter Voltage Drop versus Load Current**

At full load (5 kW), the peak output current is about 59 amps, which corresponds to a peak voltage drop of about 12 volts. Assuming that, at no load, the input voltage and/or modulation index are adjusted to give the maximum allowable output voltage (127 V<sub>rms</sub>) then at full load the output voltage would drop to about 118 V<sub>rms</sub>. This difference, although significant, is acceptable for household applications and even meets both the ITI/CBEMA and FEC specifications.

The next test for the open loop control strategy is to simulate some load transients and observe the voltage response. Using the average model from Figure 3.41, several transients were simulated and the worse case condition is shown in Figure 3.45. The load step up from no load to 5 kW is no problem with a minor glitch and the steady-state peak voltage dropping only 9 volts. The load transient from 5 kW to no-load results in a voltage spike of up to 187 volts. This spike is over the 6% steady-state limit, but still meets the CBEMA requirements because it only occurs for about 800 μs.

Another concern with running the inverter open loop is the performance with a nonlinear load. A 300 W purely nonlinear load, equivalent to at least two personal computers, was simulated and the results are displayed in Figure 3.46. The THD was calculated to be 4.03%, which is under the 5% specification. However, if the nonlinear

load were increased to 1 kW the THD would be about 6.5%. A purely nonlinear load is unlikely in a household, and adding some load resistance will help reduce the THD value.

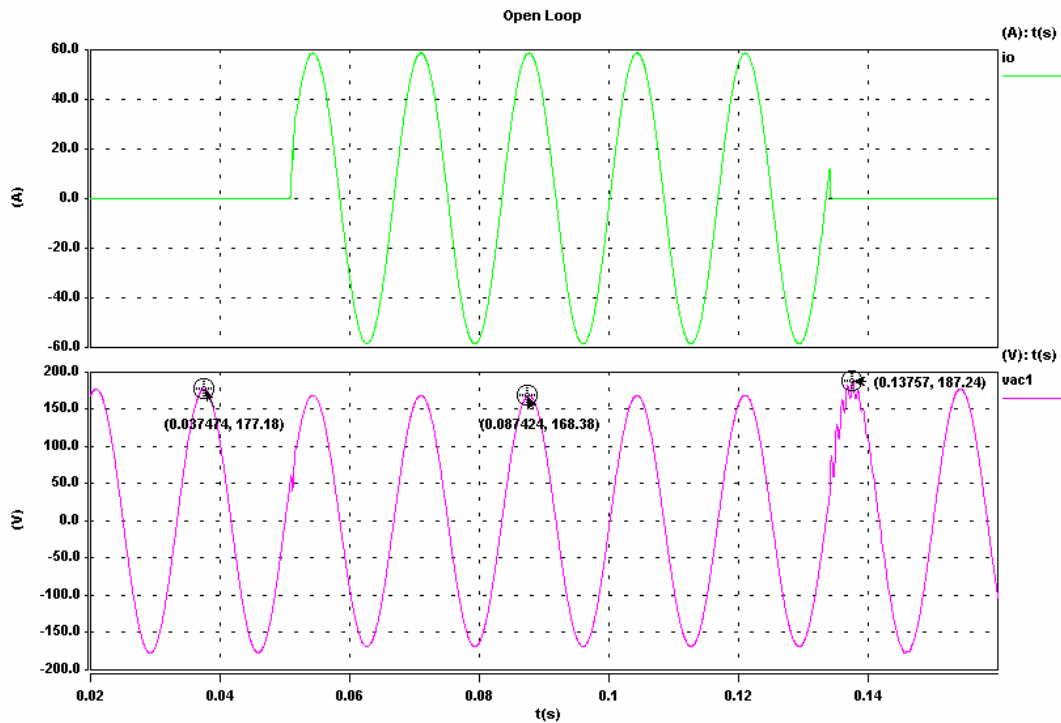


Figure 3.45 Simulated Open Loop Transient with Inverter

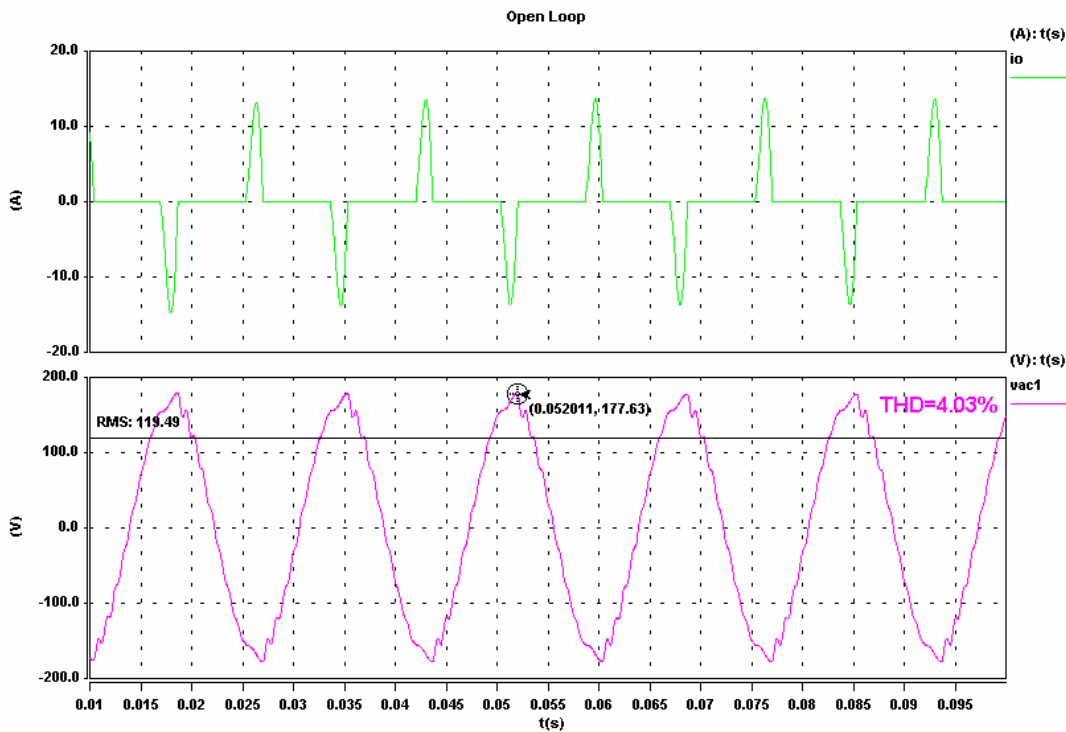
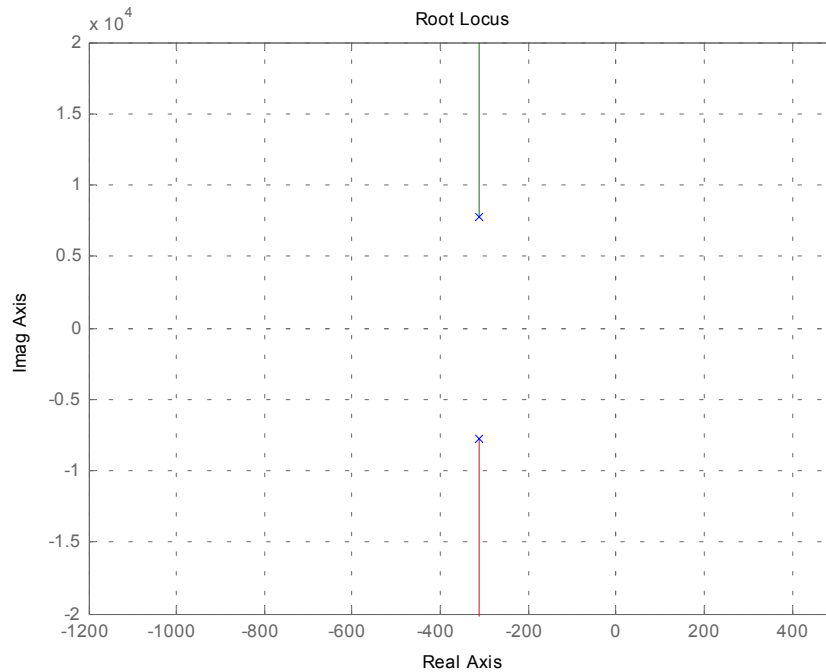


Figure 3.46 Simulated Open Loop Nonlinear Load with Inverter

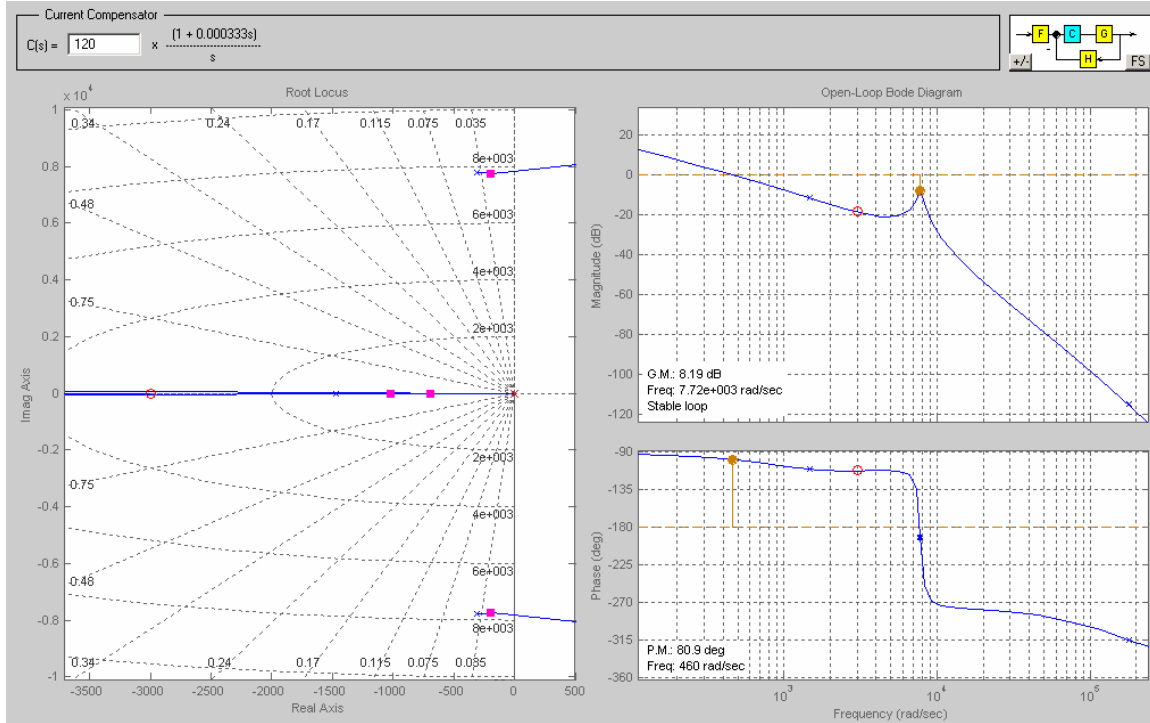
Using voltage feedback to control the modulation index seems to be the next logical step and can still result in a relatively simple and inexpensive solution. In the past many inverters relied on an rms feedback loop to regulate the voltage magnitude [20], but this system uses a continuous controller that adjusts the modulation index every switching cycle. The response does not have to be extremely fast, but cycle-by-cycle voltage regulation will eliminate the inductor voltage drop problem and should improve transient response.

Although the closed loop simulation model for the inverter can be developed, it does not allow for easy control design by analyzing the bode plot of the closed loop gain. Therefore a root locus method was employed using the transfer functions developed earlier in this chapter. Figure 3.47 shows the open loop complex poles of the plant and their trend as the gain is increased for the worse case load condition, which occurs at very light load. There is also a very high frequency pole and zero located on the real axis that not shown due to the resistance of the inductor and the small load inductance. Although this dominant pole behavior may not be ideal for UPS systems [40], voltage feedback may still be applicable for this application.



**Figure 3.47 Open Loop Root Locus of Inverter**

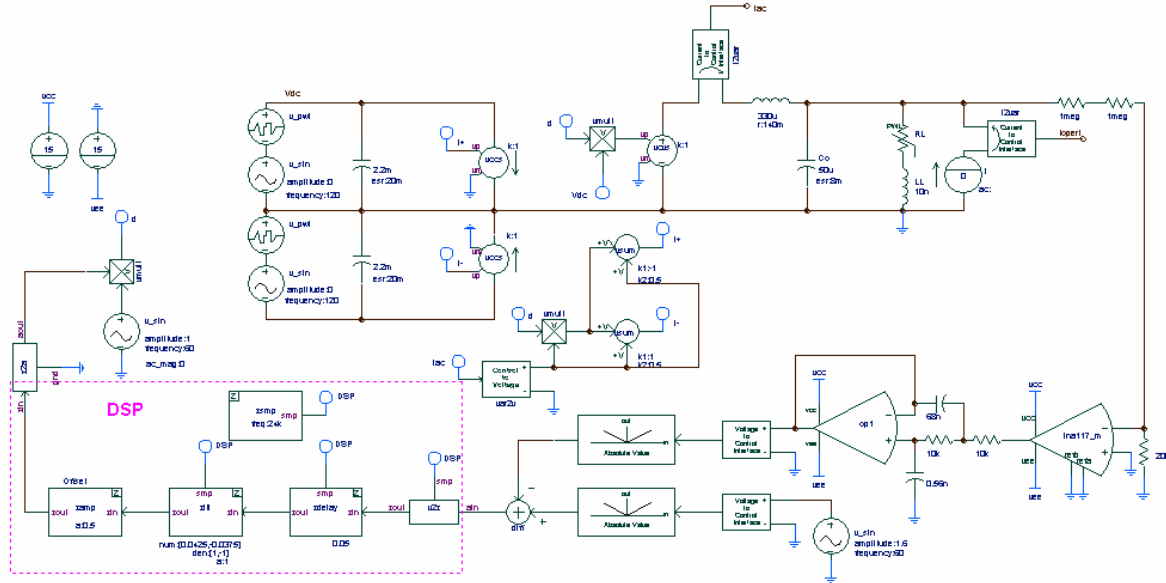
A PI compensator like the one shown in (3.9) was designed using the Single Input Single Output (SISO) Design Tool in Matlab. The transfer functions of the plant, sensor and compensator were developed and imported into the design tool, and the resultant graphical interface is shown in Figure 3.48. The gain can be easily adjusted to give the desired pole placement or desired crossover and phase margin.



**Figure 3.48 Closed Loop Root Locus of Inverter with PI Compensator**

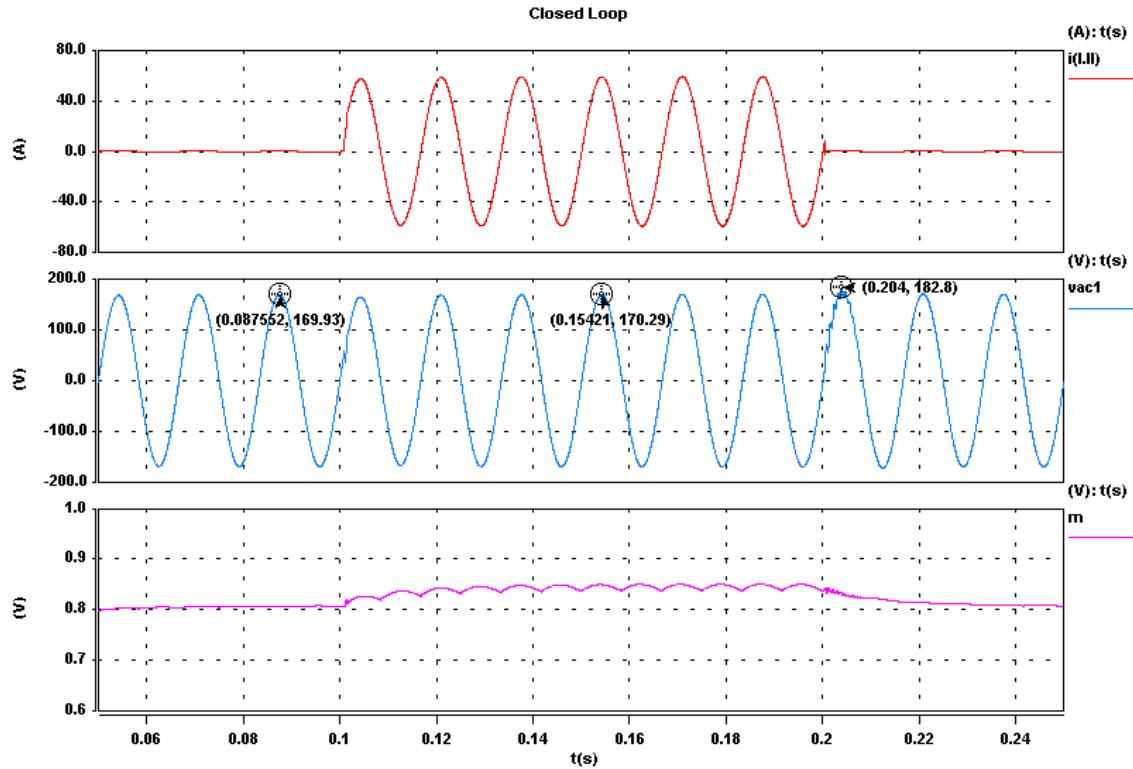
Adding a PI compensator causes the trajectory of the complex pole pair to enter the right half plane, which if the gain is increased, will result in an unstable system. Therefore the  $K_p$  value of the compensator must remain very small. Since the inverter output has a 60 Hz fundamental component, it is of crucial importance that the gain at 60 Hz be as large as possible to eliminate steady-state errors. Using a  $K_p$  value of 0.04 and a  $K_i$  value of 120 results in a stable system with a gain at 60 Hz of about 2 dB. From Figure 3.48, the phase margin is shown to be  $80.9^\circ$  and the gain margin is 8.18 dB. This compensator was implemented in Saber to analyze the transient response and steady-state regulation. The closed loop inverter circuit with digital control is shown in Figure 3.49. The absolute value of the feedback and the reference signal are taken to assure negative feedback throughout the entire sine wave cycle. In the physical system, the absolute

value is performed by the DSP, but Saber does not have a z domain absolute value model. As with the DC-DC converter, the modulator was ignored in the modeling process of the inverter, but it should not have a significant impact on the control design. Other than that, the sensing circuit and control implementation are modeled as close as possible to the physical implementation. The sinusoidal voltage source that is multiplied with the calculated modulation index emulates the sine wave look-up table in the DSP.



**Figure 3.49 Saber Closed Loop Average Circuit of Inverter**

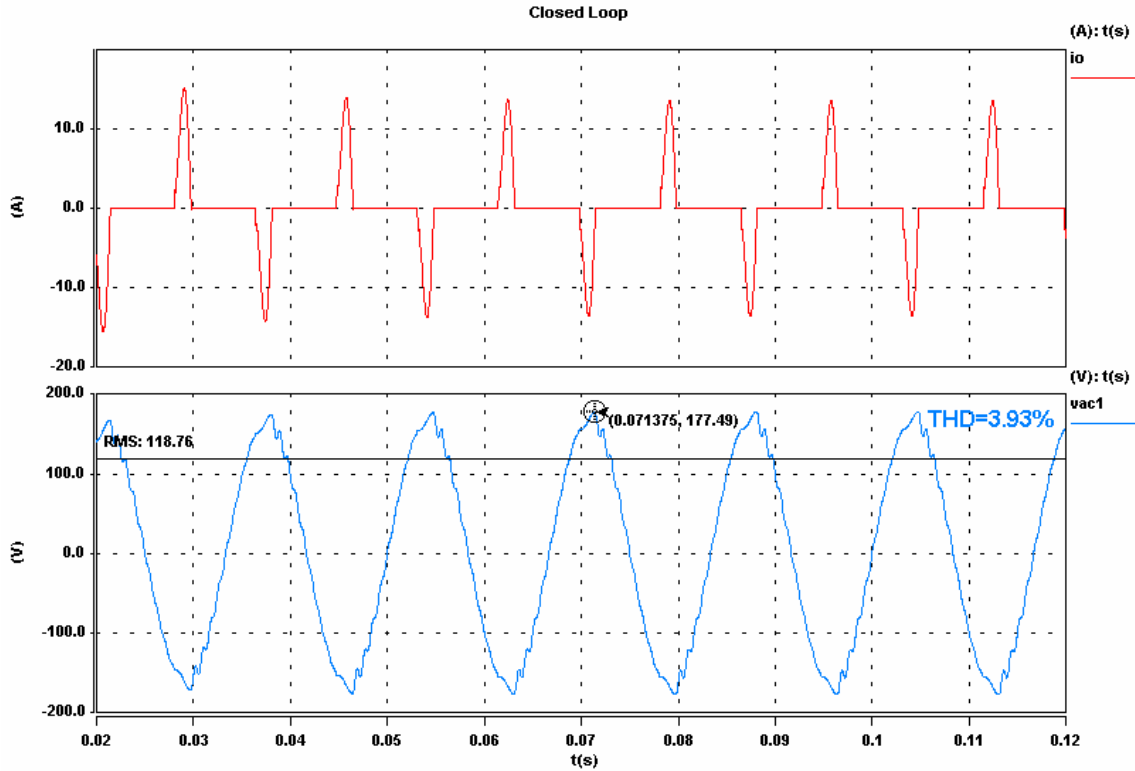
Using the model above, the same load transient from no load to 5 kW was simulated and the results are shown in Figure 3.50. Comparing it to the open loop case, there is still some overshoot, although not as large, during the transient. The steady-state regulation is much better with less than one-volt difference between no load and full load.



**Figure 3.50 Simulated Closed Loop Transient with Inverter**

Again, a purely nonlinear load was simulated and the results are shown in Figure 3.51. The THD is 3.93% versus 4.03% for the open loop case, so there is virtually no improvement for nonlinear loads. Even with the voltage loop closed, the THD will exceed 5% at larger loads and may even exceed the open loop case.

A more complex voltage compensator with a complex zero was developed using the root locus method. The trajectories of the dominant poles were pushed further into the left half plane as desired and the system seemed stable; however, the simulation revealed unstable behavior even in steady-state conditions.



**Figure 3.51 Simulated Open Loop Nonlinear Load with Inverter**

The only way to greatly improve the THD is to add current feedback to the inverter control. Using current feedback can provide the opportunity to work in the stationary frame, which would make the controller design much easier [21 and 22]. However, as discussed in 3.2.2, inexpensive current sensing is not trivial for a system of this power level. The current controllers also tend to require more computation time and could require a faster, more expensive DSP. Furthermore, although the harmonic performance using only voltage feedback is not desirable, it may be acceptable for this application.

The simulations suggest that there is very little difference in performance between an open loop strategy (assuming a  $\pm 6$  output voltage swing is acceptable) and a voltage feedback strategy. The experimental results in the section analyzing the system will prove otherwise.

### 3.4 System

Although both the DC-DC converter and the inverter work well alone, it does not mean that they will work well when put together. There can be many unforeseen complications, such as control loop interaction and impedance mismatching, when cascading two converters. Before putting the two converters together, it is important to look at the output impedance of the DC-DC converter and the input impedance of the inverter. In an ideal case, the output impedance will be low and the input impedance will be high. Figure 3.52 shows both transfer functions on the same graph for a 700 W load with the inverter running open loop and closed loop. Although the magnitudes of the two waveforms are close around 200 Hz, there is no overlap. The concern is with the phase of the closed loop input impedance at low frequencies. A  $-180^\circ$  phase shift may cause the system to go unstable.

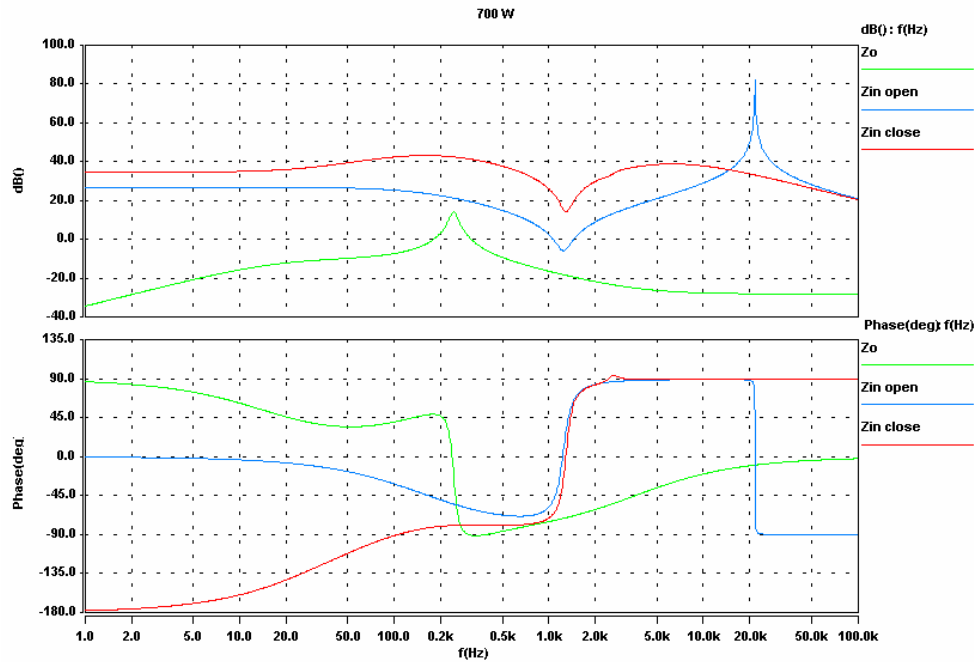


Figure 3.52 Simulated Impedances of DC-DC and Inverter

#### 3.4.1 Modeling

Cascading the two switching models from Figure 3.5 and Figure 3.38 can be done, but the simulation time is extreme long and convergence problems can easily arise. An easier method, which will still reveal some information regarding the switching interaction, is to use the average model of the DC-DC converter and switching model of

the inverter. This system model involves running the front-end closed loop and the inverter open loop and is shown in Figure 3.53.

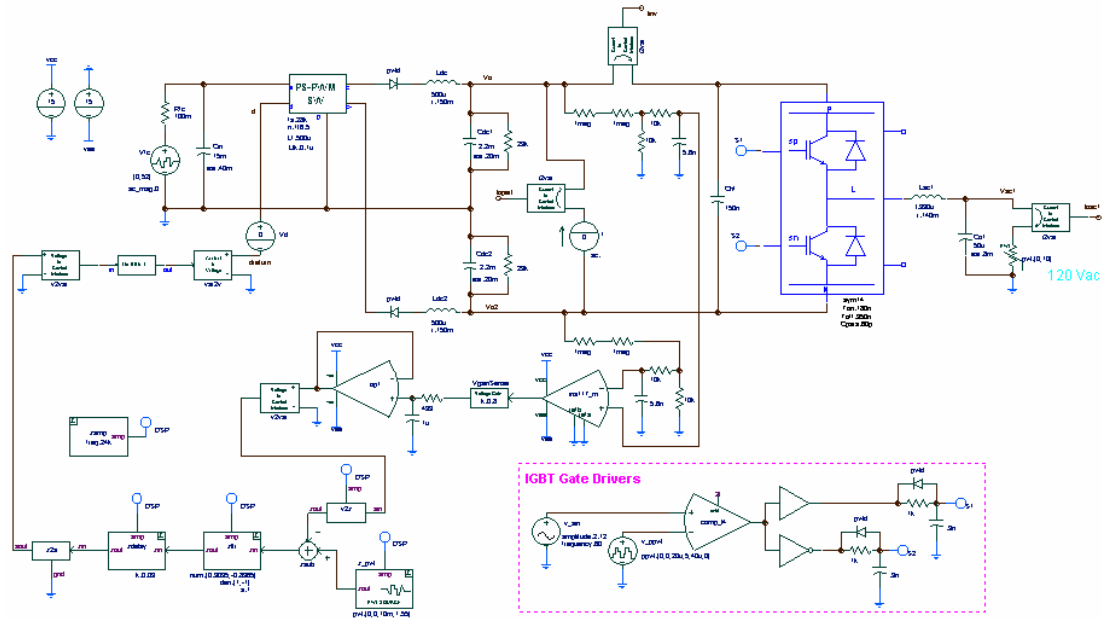


Figure 3.53 Saber Model of System

The single-phase inverter will introduce a 120 Hz ripple on the high voltage DC bus, which can be verified by simulation as shown in Figure 3.54. The DC bus ripple arises from the sinusoidal current being drawn by the inverter as shown in the figure.

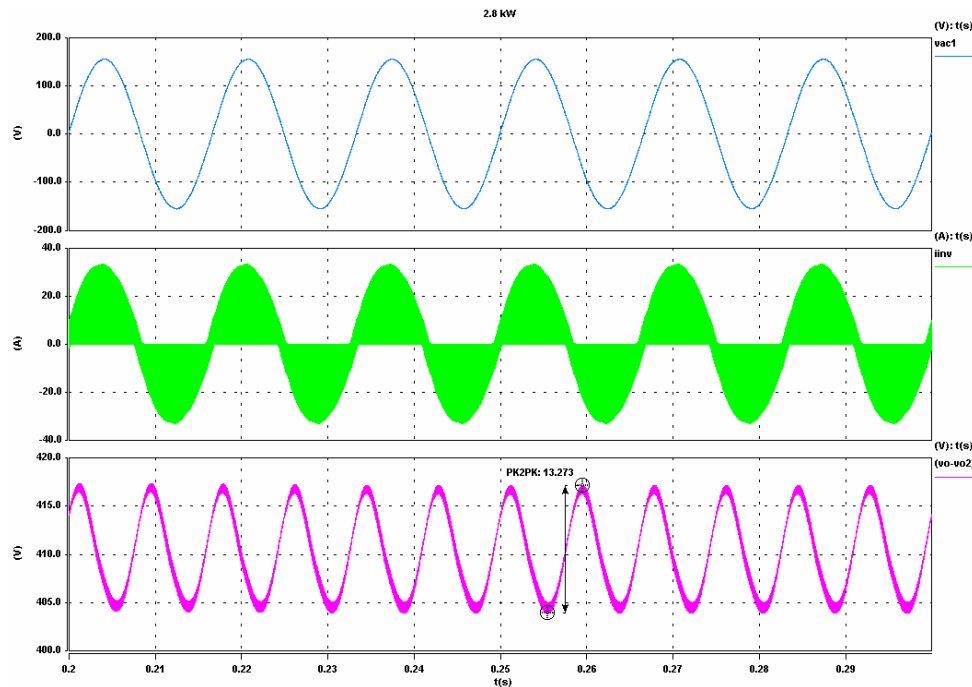
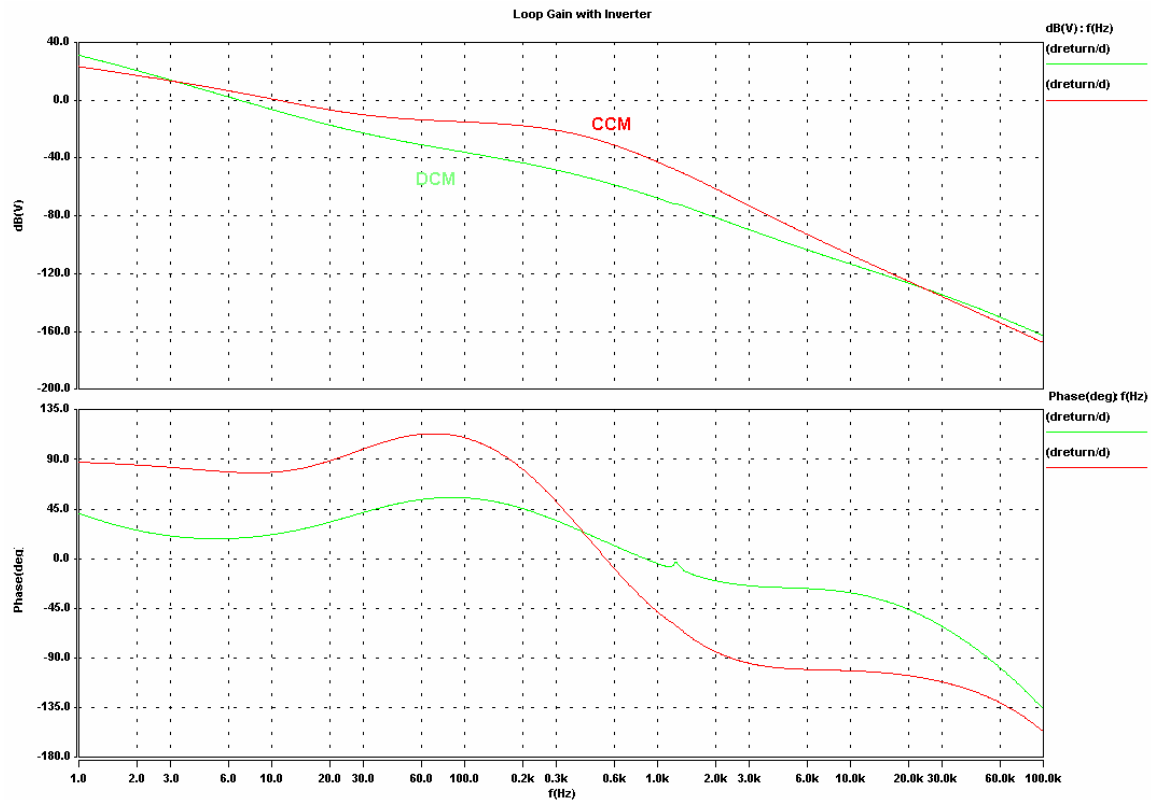


Figure 3.54 Simulated DC Bus Ripple

The transfer functions of the DC-DC converter can also be affected by adding an inverter as the load. The simulated loop gain with the inverter is shown in Figure 3.55. If compared to the DC-DC converter alone (Figure 3.18), the shapes are the same, but the gain of the CCM case is reduced. This suggests that the gain of the DC-DC converter could be increased when operating with the inverter as a load.



**Figure 3.55 Simulated Closed Loop Gain of Front-End with Inverter**

The output impedance is also affected as shown in Figure 3.56. There is a slight glitch just above 1 kHz, which may be due to the AC output filter resonance. However, the biggest difference when compared to Figure 3.20 is the low frequency portion. There is no resonance and the DCM and CCM cases are very similar. The inverter is an active load which makes output impedance slightly different.

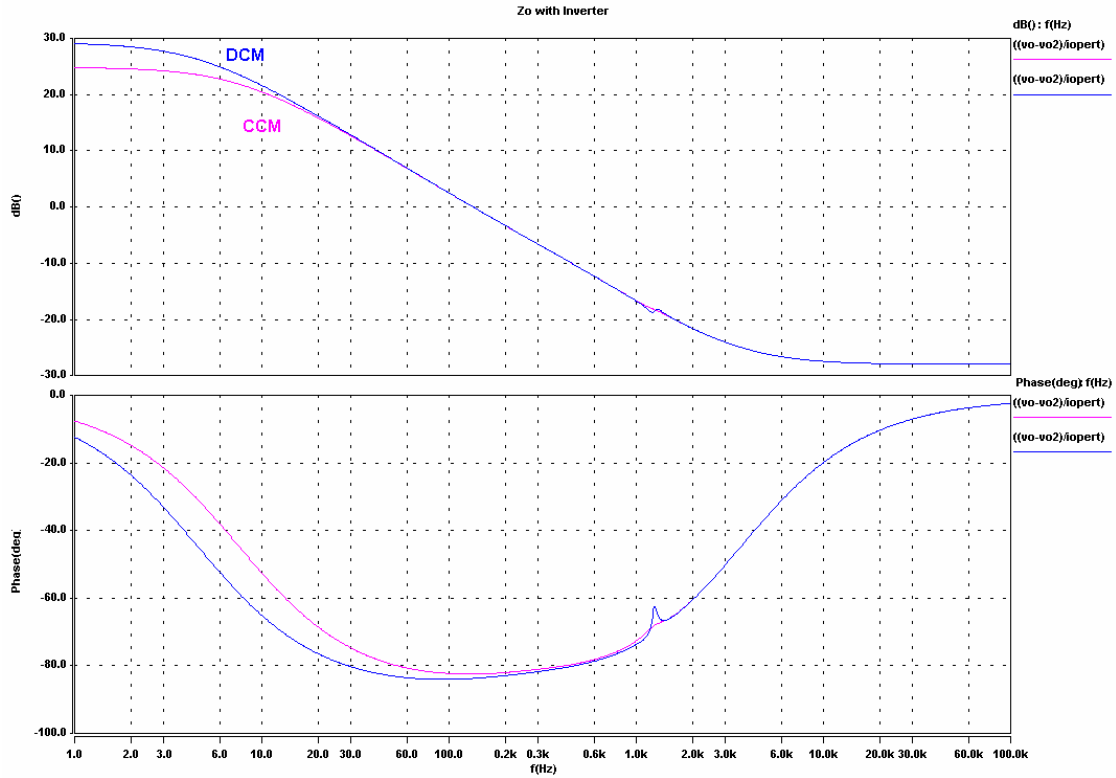


Figure 3.56 Simulated Output Impedance of Front-End with Inverter

For control and simulation purposes, it is useful to develop an average model of the entire system. Due to the input current sources of the inverter model in Figure 3.49, the two closed loop average models can be cascaded together as shown in Figure 3.57.

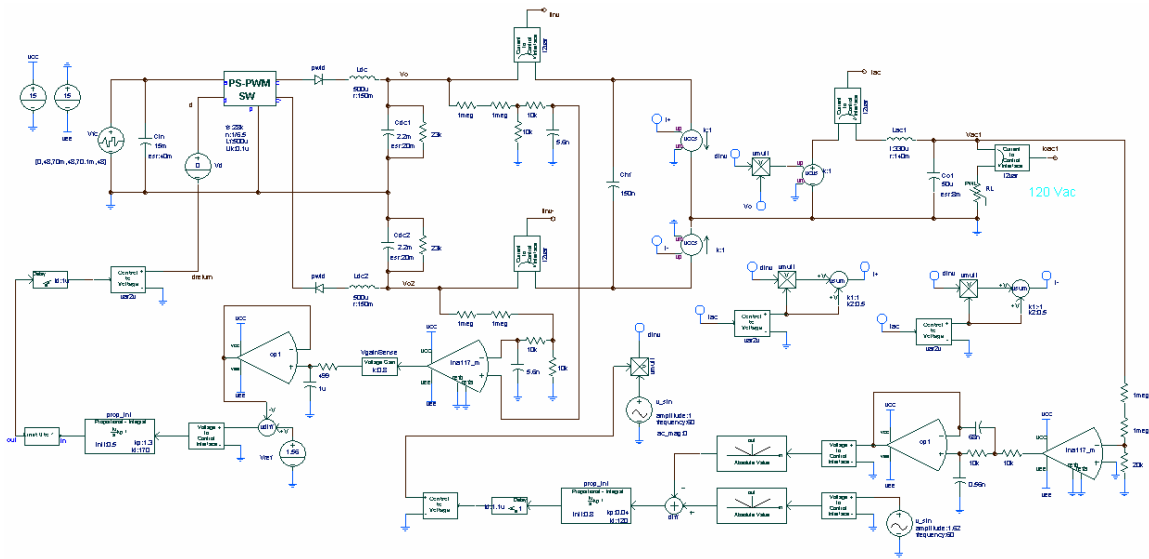
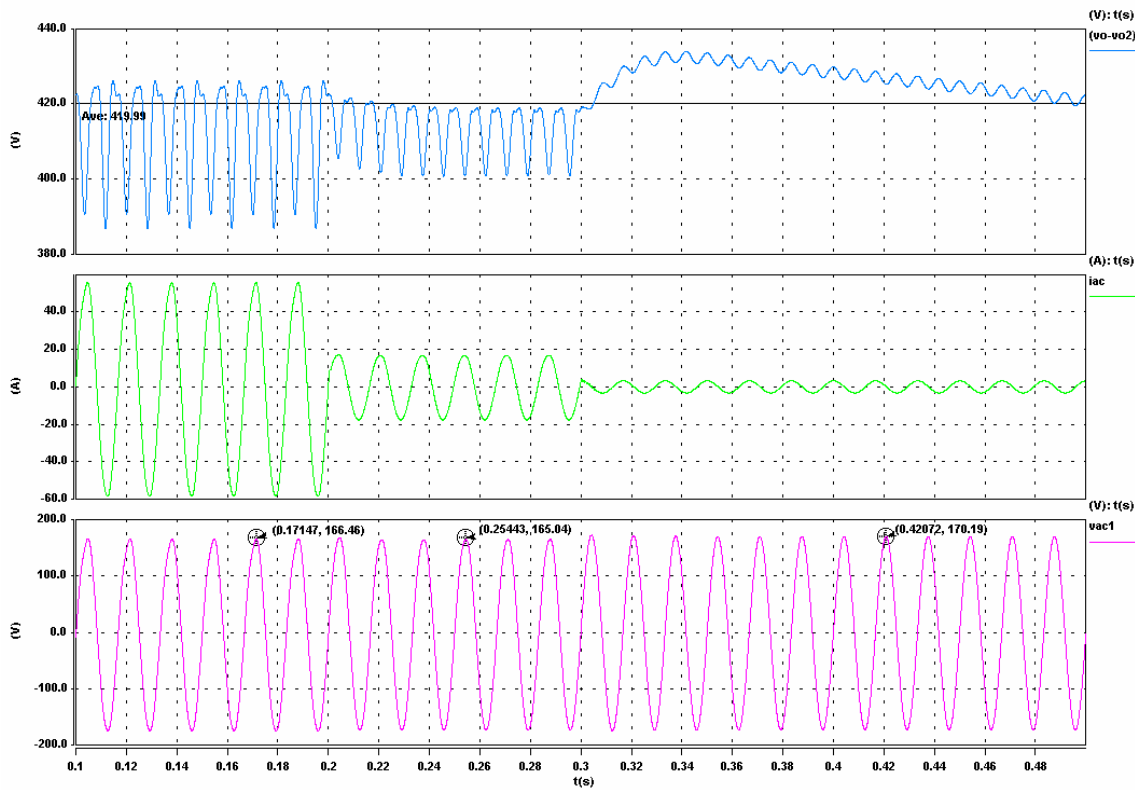


Figure 3.57 Saber Model of System

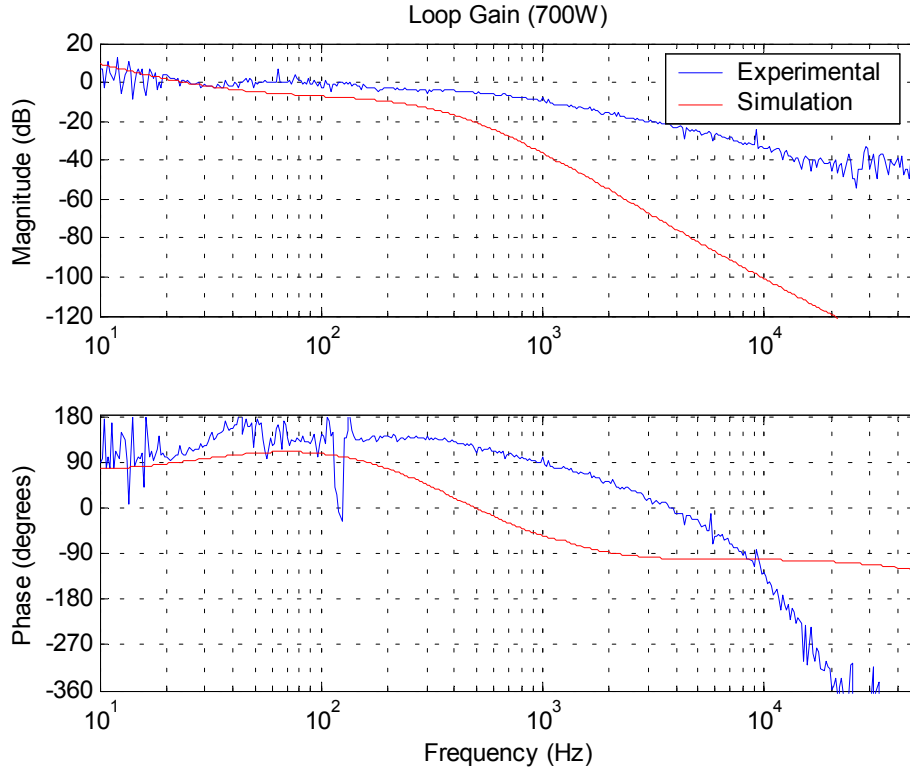
Now load transients can easily be simulated for open loop and closed loop conditions and the response of the system can be analyzed. A series of load steps with both loops closed is shown in Figure 3.58. As expected, the DC bus ripple varies with load and at full load it is almost 30 V peak to peak. The AC output voltage is not perfectly regulated due to the 120 Hz ripple on the DC bus. The maximum of the output voltage occurs at the maximum of the ripple, and the inverter control loop is not fast enough to completely adjust to the 120 Hz component of the DC bus. Despite this shortcoming, the output voltage still regulates well within specification across the entire load range. Also shown in Figure 3.58 is the slow response when the DC-DC converter goes into DCM operation. However, the inverter control loop is able to adjust and maintain a constant output voltage. The open loop inverter simulation (not shown) displays a significant decrease in performance due to the DC bus ripple and DCM response.



**Figure 3.58 Simulated Load Transients of System**

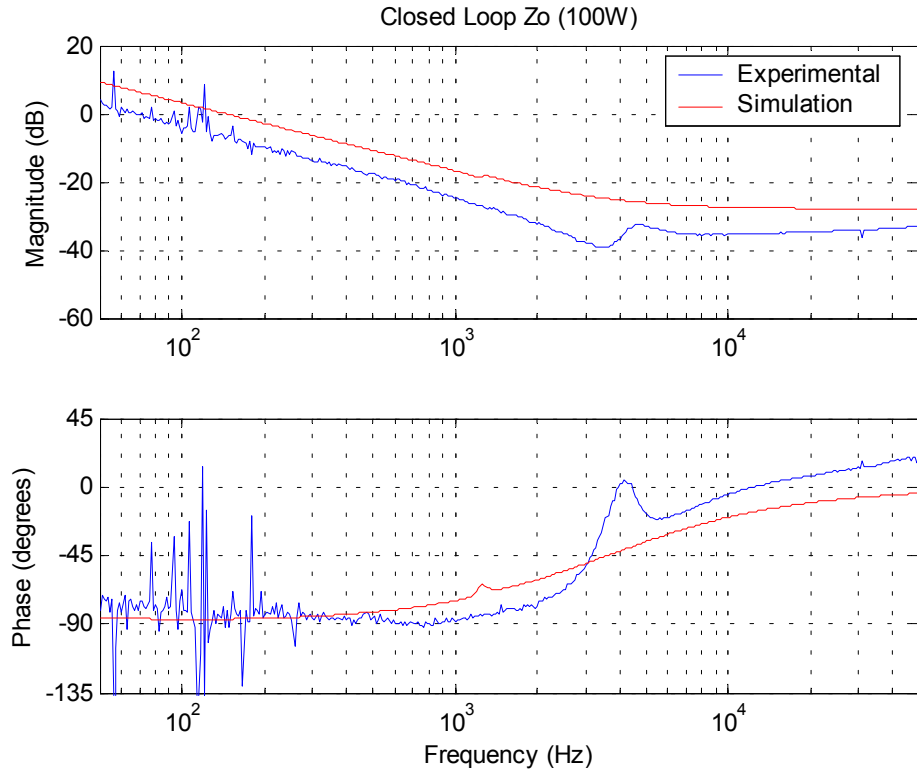
### 3.4.2 Experimental Verification

A few DC-DC converter transfer functions were measured with the inverter running open loop to verify the differences that the simulations show. The DCM case is virtually the same so it is omitted, and the CCM case is shown in Figure 3.59. The experiment agrees with the simulation except for the attenuation problem past the cut-off frequency, which was discussed earlier.

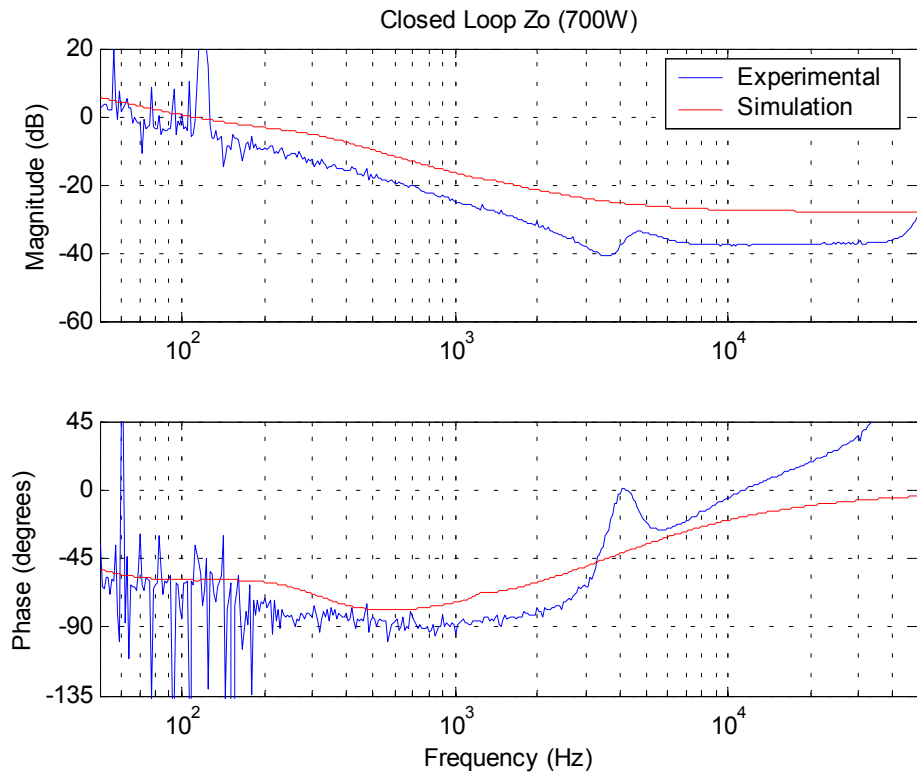


**Figure 3.59 Experimental Closed Loop Gain in CCM with Inverter**

The closed loop output impedance with the inverter was measured for DCM and CCM conditions, and the results are displayed in Figure 3.60 and Figure 3.61. The results contain a little more noise than the DC-DC alone because of the ripple induced on the DC bus from the inverter. Overall, the results match the simulations extremely well in both cases. The only difference is the glitch that shows up in the simulation around 1.2 kHz is now in the 4 kHz region. These experimental results confirm that the inverter does effect, albeit slightly, the operation of the DC-DC converter.

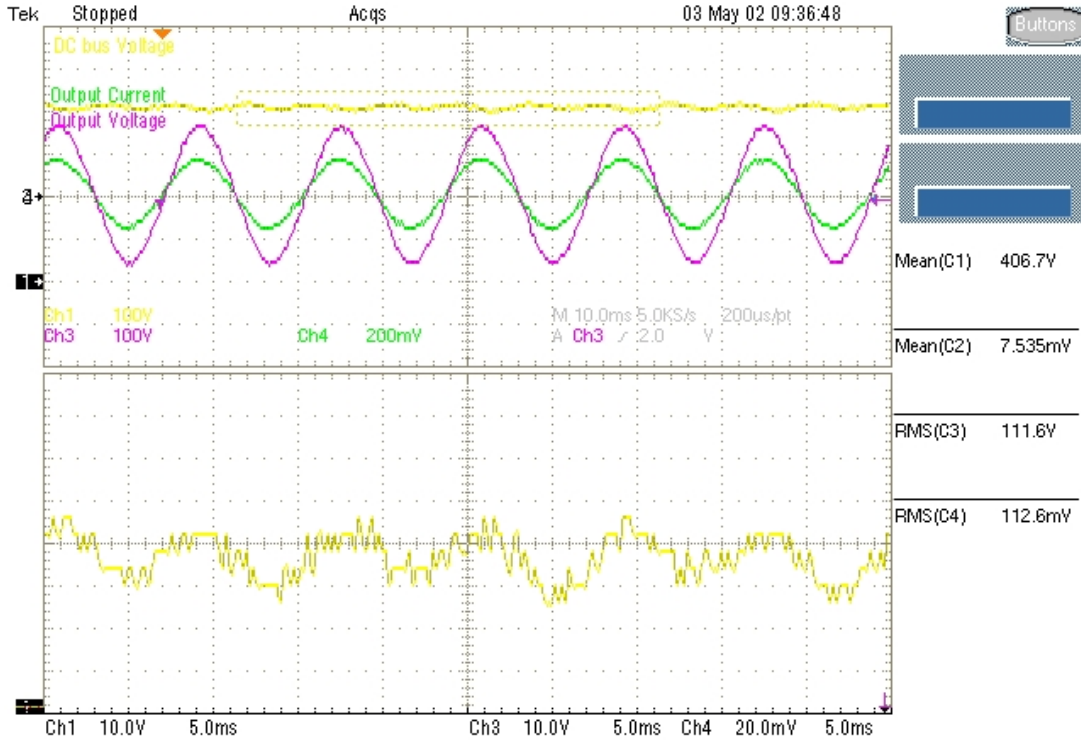


**Figure 3.60 Experimental Closed Loop Output Impedance in DCM with Inverter**



**Figure 3.61 Experimental Closed Loop Output Impedance in CCM with Inverter**

The system was operated with the inverter open loop and closed loop and data was taken to compare the performances of each control strategy. First, however, the DC bus ripple behavior at a load of 1.3 kW was captured and is shown in Figure 3.62. The exploded view of the DC bus, shown in the bottom half of the figure, exhibits about a 12 V peak to peak ripple of 120 Hz.



**Figure 3.62 DC Bus Voltage Ripple**

Open loop output waveforms during a load transient from 1.2 kW to 1.7 kW are shown in Figure 3.63. The large spike in current is due to the inrush current of the light bulbs that were used as loads. The close-up shows that the voltage drops about 20 V for about half of a cycle and then slowly ramps back up. This is not desirable behavior, but is still within the limits of the ITI/CBEMA specification. As discussed in the previous section, the THD can be a problem with nonlinear loads. Figure 3.64 shows the measured harmonic spectrum using three desktop computers as a load. Using the first 21 harmonics shown, the THD can be calculated, using (3.18), to be 4.61%.

$$THD = \frac{\sqrt{\sum h_i^2}}{h_1} * 100\% \quad (3.18)$$

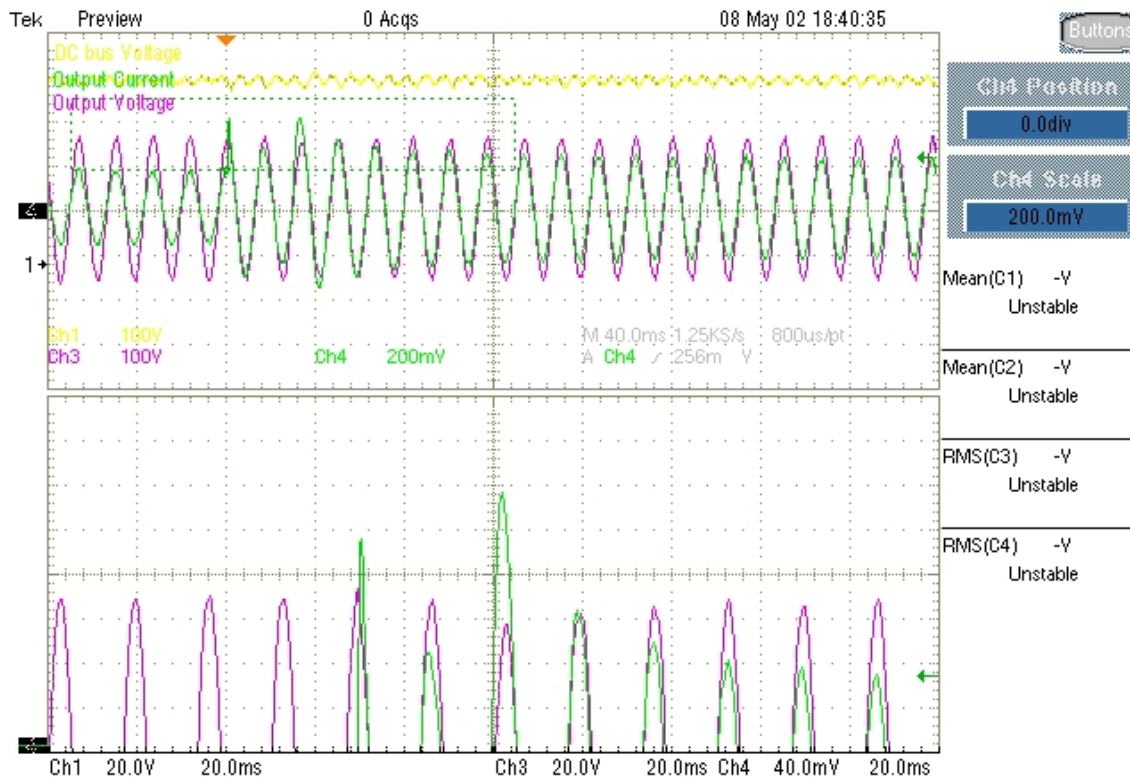


Figure 3.63 Load Transient with Inverter Open Loop

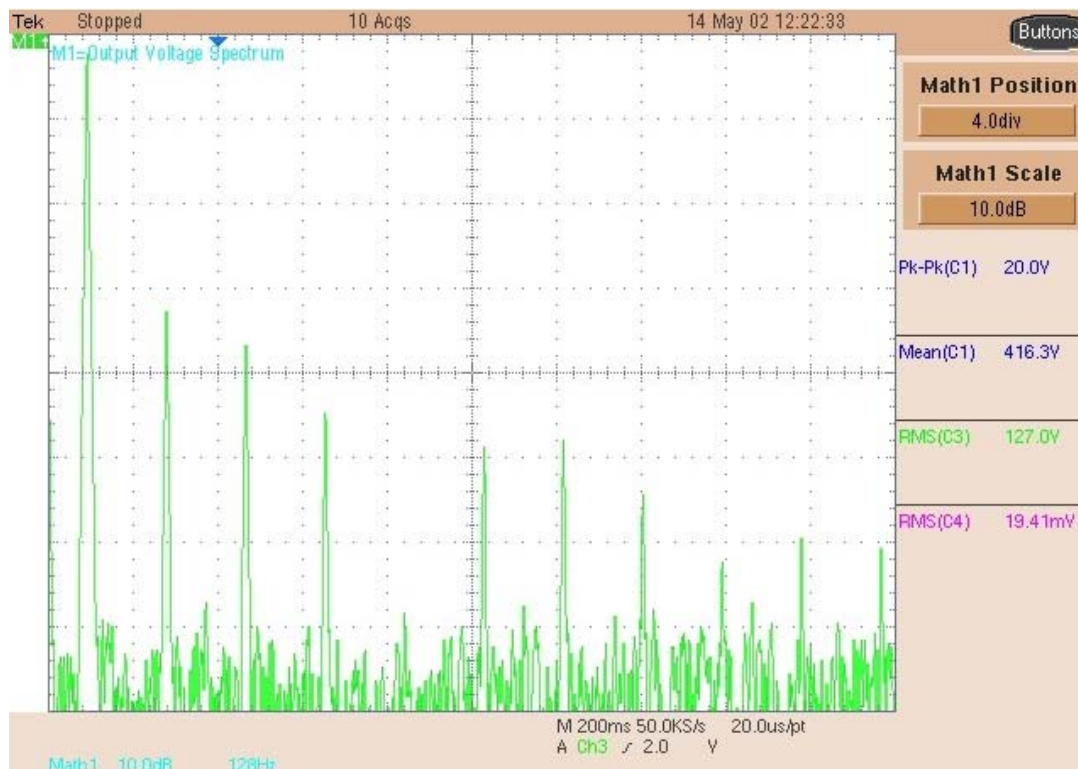
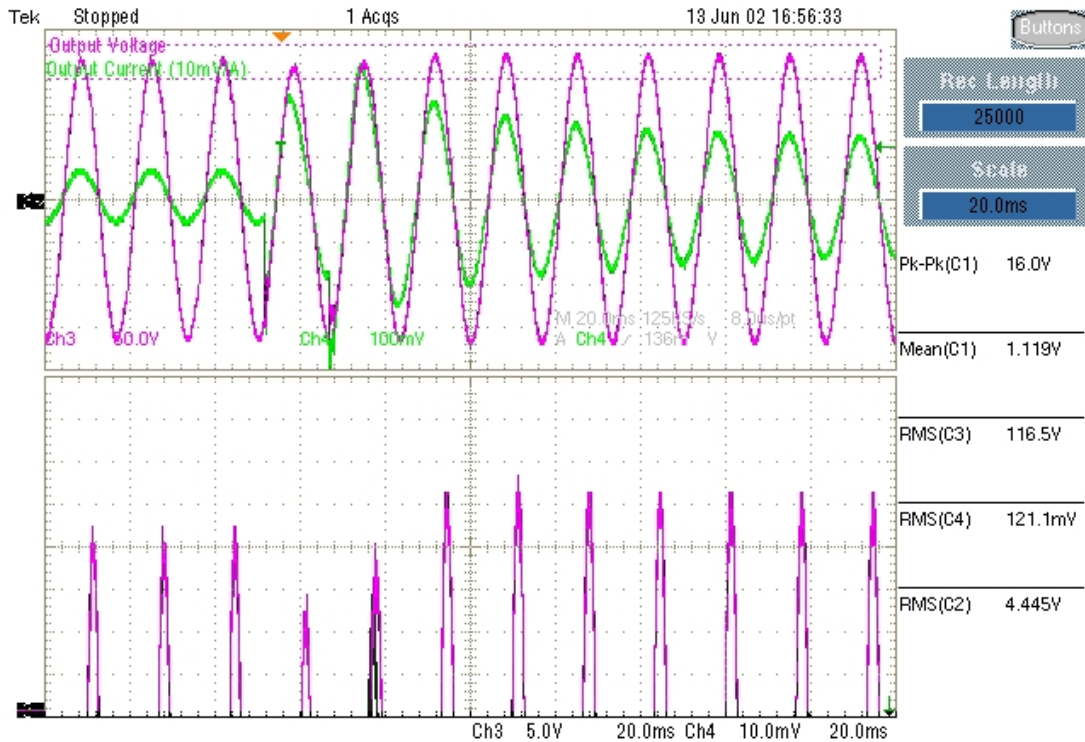


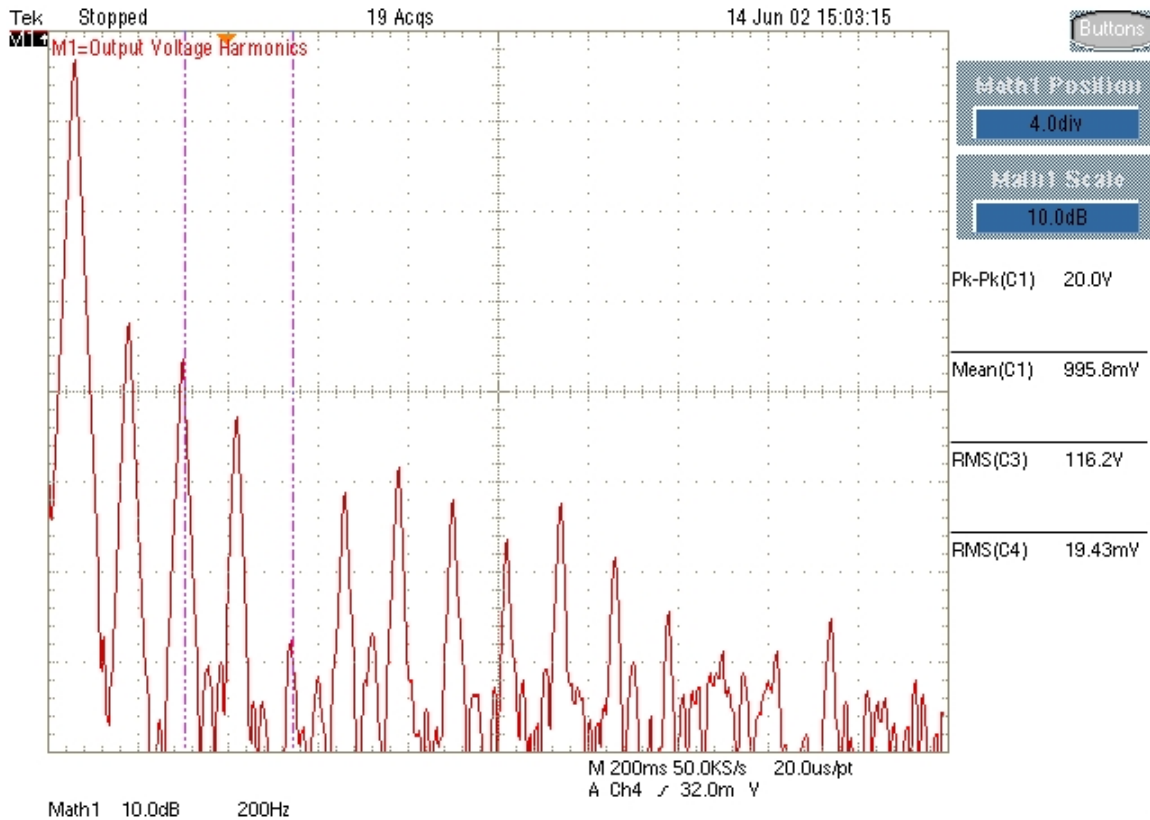
Figure 3.64 Harmonic Spectrum with Inverter Open Loop

The same tests were run with the inverter running closed loop and the results are shown in Figure 3.65 and Figure 3.66. The transient response, although not perfect, is much improved. The magnified output voltage shows a drop of only seven volts, which is well within the limits.



**Figure 3.65 Load Transient with Inverter Closed Loop**

The closed loop harmonic spectrum with the same computer load is shown in Figure 3.66, and the THD is calculated to be 4.64%. As the simulation predicted, closing the voltage loop does not improve the THD when operating with a purely nonlinear load.



**Figure 3.66 Harmonic Spectrum with Inverter Closed Loop**

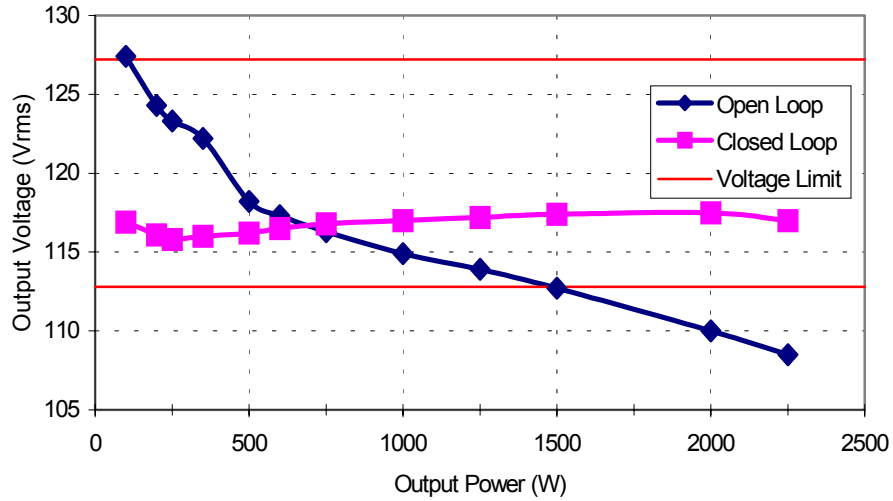
A small resistance was added in parallel with the computer power supplies to compare the difference. The harmonic spectrums (not shown) exhibit a reduction in THD for both the open loop and closed loop cases. A summary of the results is shown in Table 3.2.

**Table 3.2 THD for Various Conditions**

| Load                                | Open Loop | Closed Loop |
|-------------------------------------|-----------|-------------|
| 300 W Nonlinear                     | 4.61 %    | 4.64 %      |
| 300 W Nonlinear and 150 W Resistive | 3.97 %    | 4.28 %      |

Figure 3.67 compares the output voltage regulation for the inverter running open loop and closed loop. The open loop version does not meet the 6% voltage limit for all load conditions, but the closed loop output is nicely regulated. Although the calculations and simulations show that it may be possible to meet the specification running open loop, the experimental results show otherwise. This is because there are parasitics in the system that are not modeled in the simulation and in reality the DC-DC converter voltage

is not the same in DCM and CCM conditions. The DCM/CCM boundary occurs in the vicinity of 400 W, which is exactly where the open loop voltage dropped eight volts.



**Figure 3.67 Experimental Output Voltage vs. Output Power**

The experimental results clearly show that running the inverter with a fixed modulation index is not acceptable. Therefore the simple closed loop design was selected and the system was tuned for near optimal performance. A summary of the final compensator parameters is shown in Table 3.3. Device waveforms and experimental results for a wide variety of load conditions are given in the next chapter.

**Table 3.3 Final Compensator Values**

| PI Parameters | DC-DC | Inverter |
|---------------|-------|----------|
| $K_p$         | 1.3   | 0.05     |
| $K_i$         | 170   | 120      |

## Chapter 4 – Experimental Results

### 4.1 Testing with a Power Supply

The fuel cell and ultracapacitors were not readily available during most of the testing phase. Consequently, the majority of the experimental data was taken using a power supply as the input source and without supplemental energy storage.

#### 4.1.1 Test Set-up

The system power source was a 300 A, 50 V DC power supply, which was more than capable to deliver the current for high power loads and for transients. A 1.5 kW light bulb bank was used for the majority of the resistive load testing. Light bulbs are easily switched on and off and are a good representation of a common household load. The DSP was operated via a serial connection, so changes to the code could be easily made. A picture of the bench test set-up is shown in Figure 4.1.

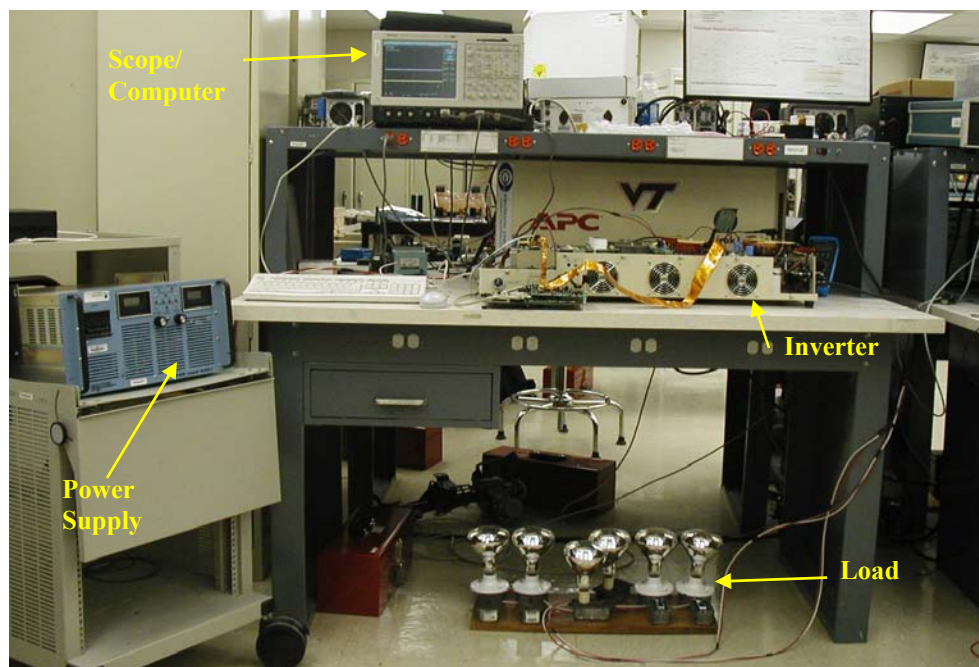


Figure 4.1 Power Supply Test Set-up

#### 4.1.2 Device Waveforms

Waveforms of some of the major components were captured to assure correct operation. Figure 4.2 shows the voltage across two of the MOSFETs (Q2 and Q4) and the corresponding voltage across the transformer. The phase-shift operation is clearly seen here, because when the two voltages overlap the transformer voltage is zero. The

leading leg MOSFET voltage and corresponding gate signal is shown in Figure 4.3. The bottom half of the figure clearly shows that the device voltage goes to zero before the gate is turned on, thus achieving ZVS.

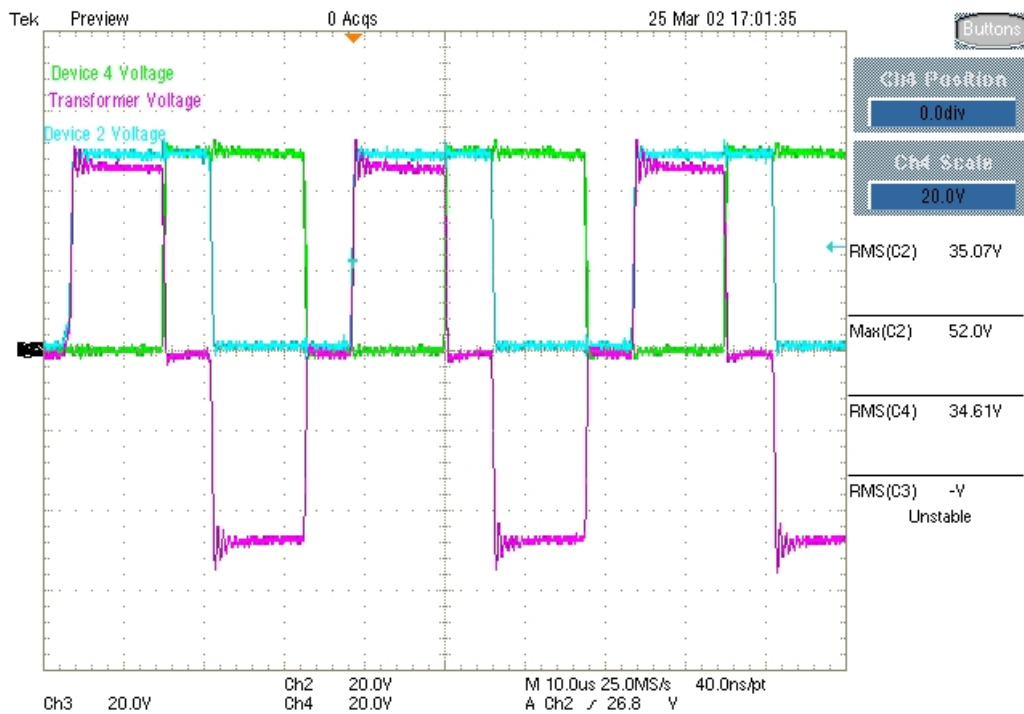


Figure 4.2 Transformer Voltage at 650 W

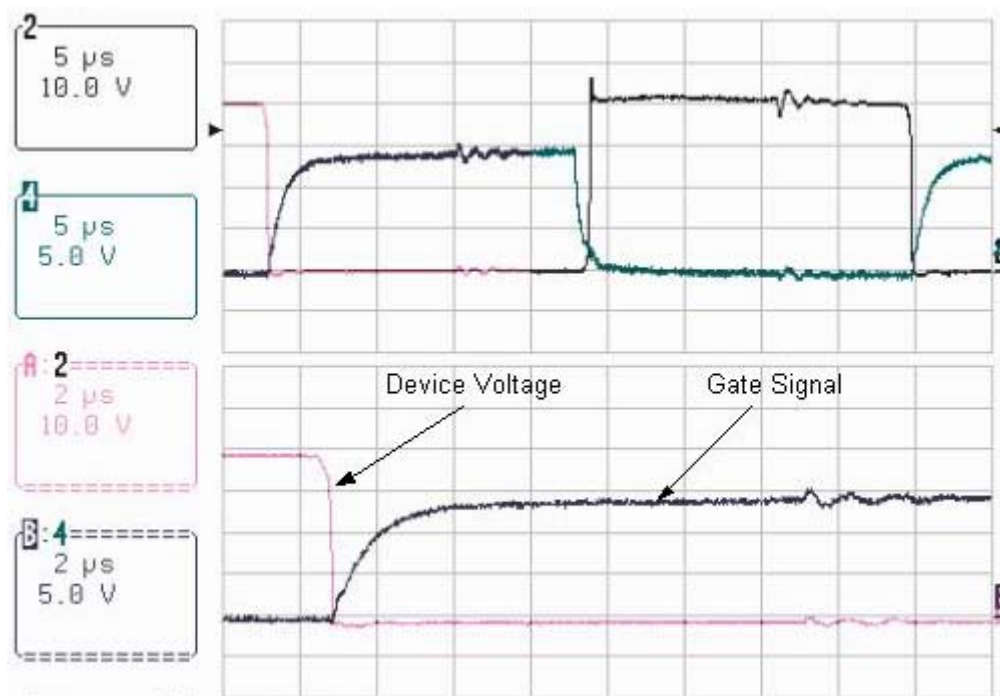
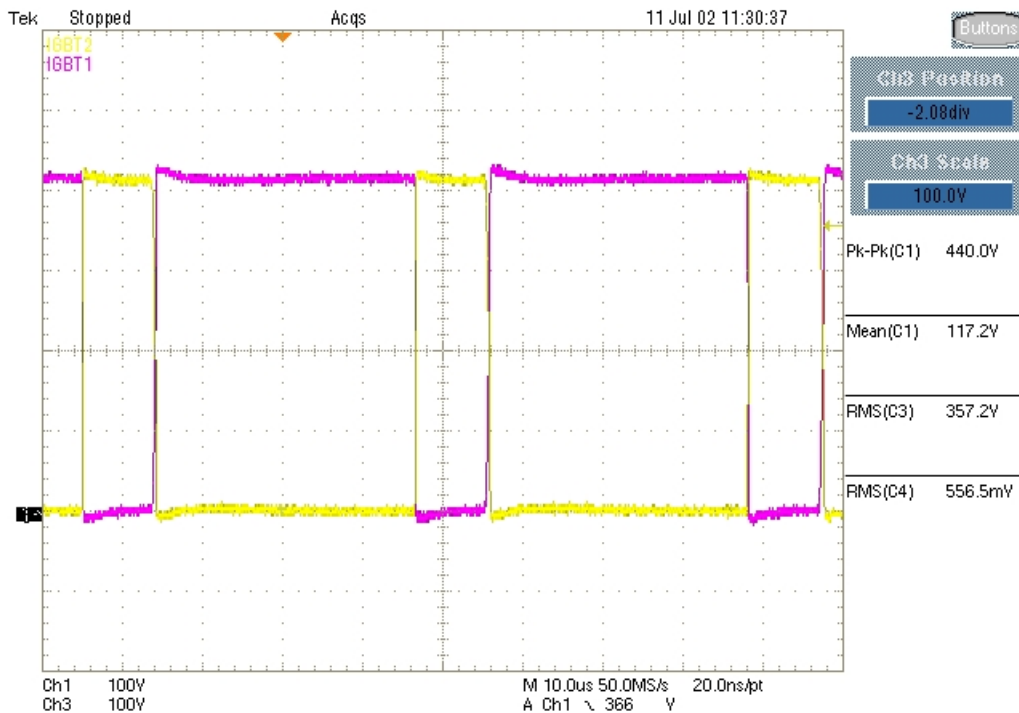


Figure 4.3 ZVS of Leading Leg MOSFET at 500 W

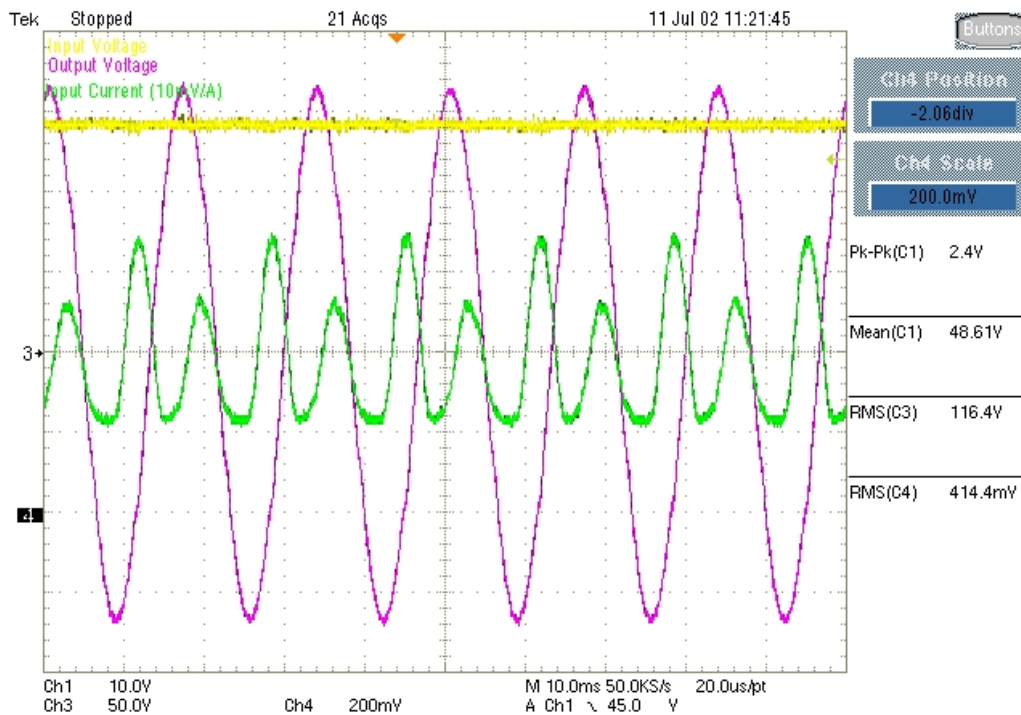
The two complementary IGBT voltages are shown in Figure 4.4. The IGBTs used are considered relatively fast for their class, but there is significant voltage crossover shown in the figure. The voltage crossover does not necessarily represent device turn-on, but it does present a potential predicament at high power levels. The dead time, already rather long at 1.5  $\mu$ s, could be increased, but then the DC bus voltage or modulation index would need to increase to keep the output voltage at an acceptable level. This concern leads to a future optimization problem involving device selection, gate drive design, dead time value, and DC bus voltage.



**Figure 4.4 IGBT Voltages at 1 kW**

### 4.1.3 Input Waveforms

The input voltage and current were measured using a power supply to compare with the fuel cell results shown in the next section. Figure 4.5 demonstrates that the inverter 120 Hz ripple current is reflected all the way to the input of the DC-DC converter. The ripple is actually quite severe with a peak-to-peak value of about 20 A (the scale in Figure 4.5 is actually 20 mV/A). The input voltage is extremely stiff, which can be attributed to the large power supply used.



**Figure 4.5 Input Waveforms at 750 W**

#### 4.1.4 Output Waveforms

The system was tested over a wide range of load conditions, many of which are similar to what a typical house might see. Load transients of 1250 W are shown in Figure 4.6 and Figure 4.7. The load step exhibits very good performance with a voltage sag of only 7 volts. The load dump condition (Figure 4.7), as expected from the simulations, is not quite as good. The overshoot is about 12 volts initially and then it decays back to the steady-state voltage in about 5 cycles. Although not great, this will easily meet the ITI/CBEMA standards.

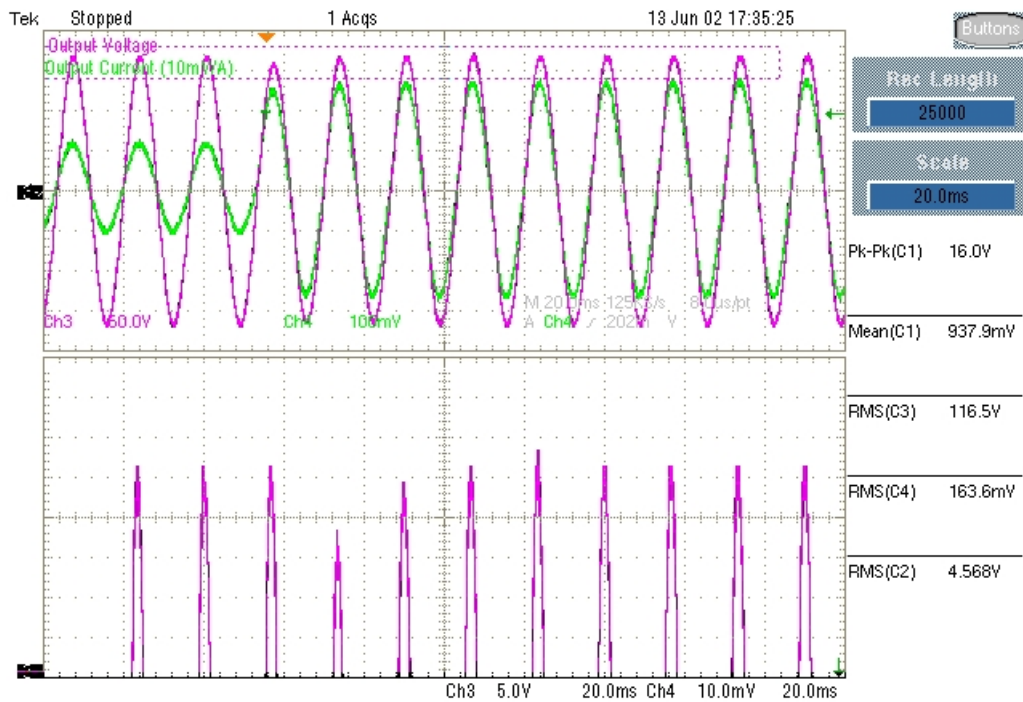


Figure 4.6 Load Transient – 1 kW to 2.25 kW

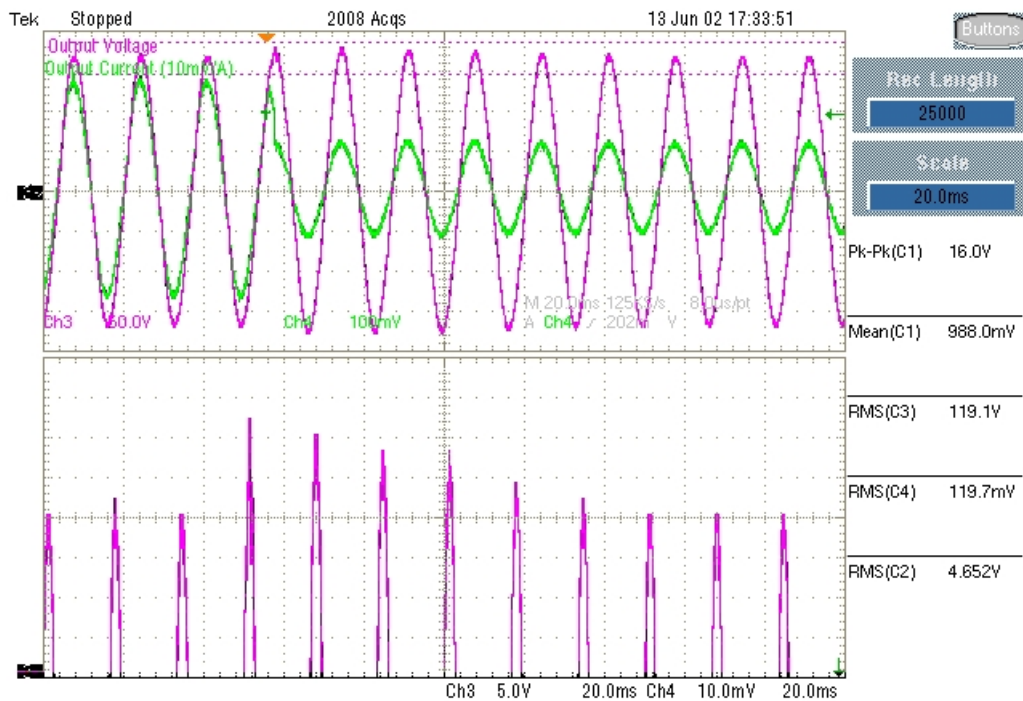
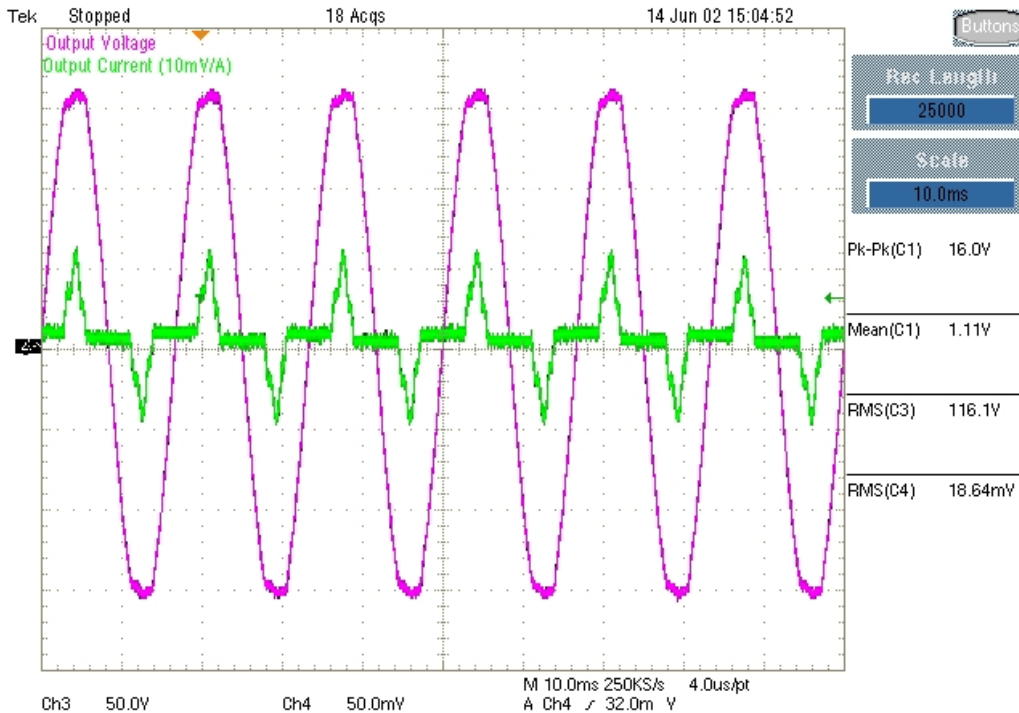


Figure 4.7 Load Transient – 2.25 kW to 1 kW

Nonlinear loads, such as computers and stereo equipment can be a common occurrence for households. The output voltage and current of the inverter while powering three desktop computers is shown in Figure 4.8. As expected, the current is far from sinusoidal and the voltage is somewhat distorted, with a THD of 4.6%.



**Figure 4.8 Output Waveforms with Nonlinear Load**

Households can also have significant inductive loads, such as air conditioning and refrigerators. The output voltage and current with a RL ( $R=10\Omega$ ,  $L=20\text{ mH}$ ) load operating at 800 VA is shown in Figure 4.9. The output voltage is slightly distorted at the peaks, but the THD is less than 3%.

As mentioned in Chapter 1, motor starts can cause huge inrush currents and require the inverter to be overrated. Figure 4.10 shows the start-up of a jigsaw on top of a 1 kW resistive load. The current shoots up to nearly 5 kW for one cycle and then decays back to the steady-state value of 1.8 kW. The large inrush current causes some temporary distortion on the negative cycle of the output voltage. This is due to the DSP's pulse width elimination algorithm that occurs at high modulation index values. Short pulse elimination is necessary due to the slow transition time of the IGBTs. By increasing the DC bus voltage, the modulation index can be reduced and the problem can be eliminated.

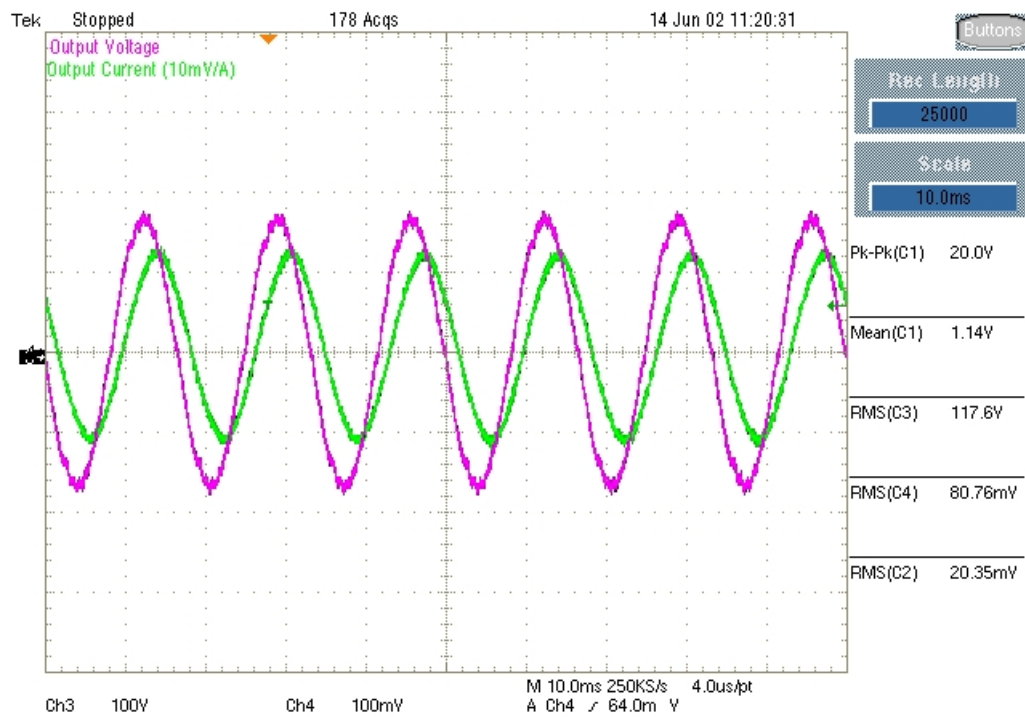


Figure 4.9 Output Waveforms with 0.8 Leading Power Factor

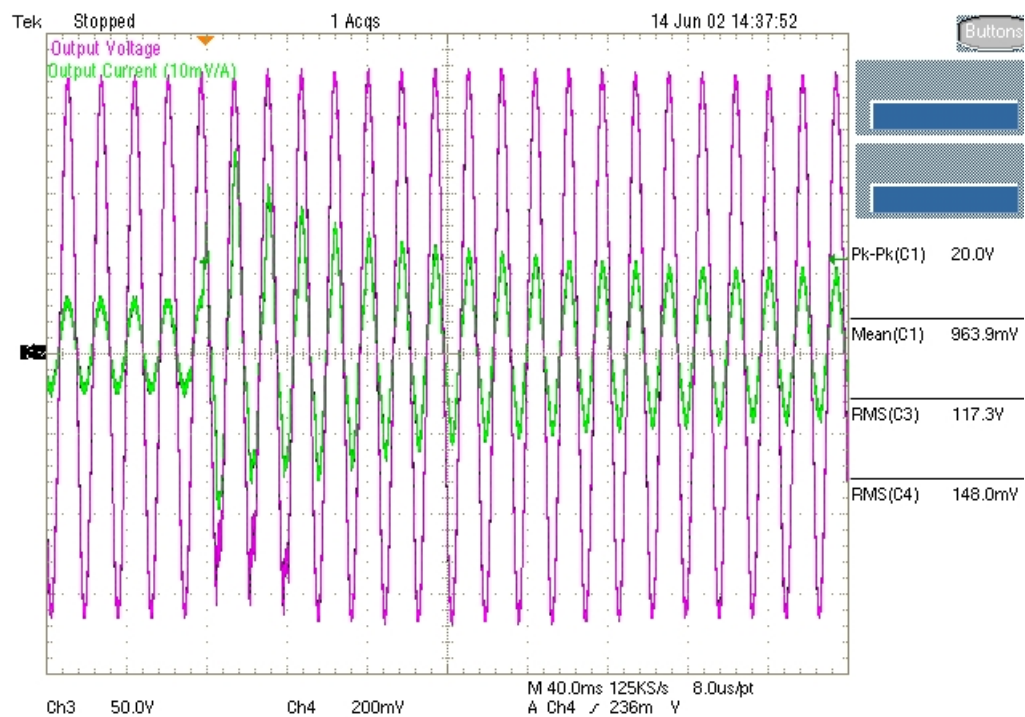


Figure 4.10 Output Waveforms During Motor Start

Figure 4.11 shows the output voltage and current of both phases, with phase one operating at 1.5 kW and phase two operating at 1.7 kW. This is a good representation of an average output for a household.

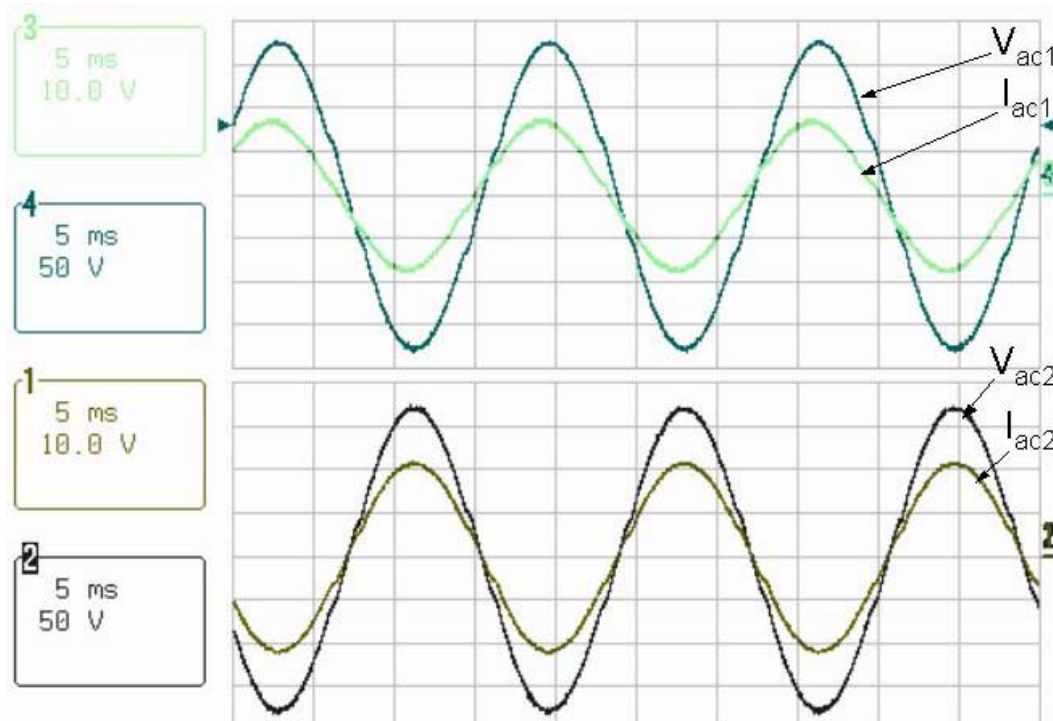


Figure 4.11 Output Waveforms of Both Phases at 3.2 kW

#### 4.1.5 Efficiency

Efficiency measurements for one phase of the converter were taken up to about 75% of rated power. The results for the DC-DC converter alone and with the inverter are shown in Figure 4.12.

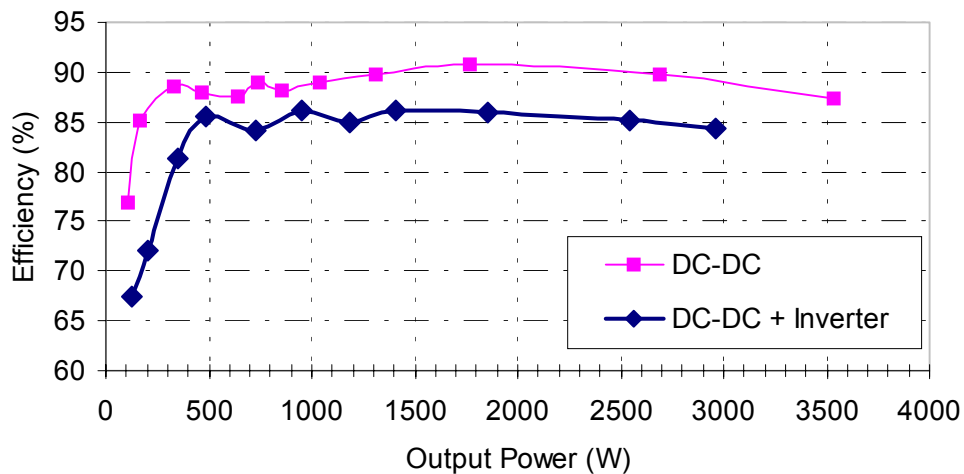


Figure 4.12 Efficiency vs. Output Power

The efficiency of the system is around 85% for loads above 500 W, which does not exactly meet the specification but is quite good for a prototype two-stage converter. It is presumed that for higher power levels, the efficiency should improve. However, Figure 4.12 actually shows a minor decline around 2.6 kW. This is mostly related to the copper losses of the DC-DC PCB. The PCB copper was accidentally underrated and therefore gets extremely hot when the power is increased. This is also one of the reasons why the system could not be run continuously at full power.

## 4.2 Testing with a Fuel Cell

Fortunately, thanks to EPRI Power Electronics Application Center, some testing was done with a fuel cell and ultracapacitors, and the results are summarized below.

### 4.2.1 Test Set-up

The inverter was tested on a Proton Exchange Membrane (PEM) hydrogen fuel cell from Enable™ Fuel Cell Corporation. The fuel cell, rated at 3 kW, consists of three 17 V stacks, giving a nominal output voltage of 51 V. Symmetrical ultracapacitors were placed across the fuel cell output. The test setup, not including the ultracapacitors, is shown in Figure 4.13. The ultracapacitors were located about 5 meters behind the fuel cell, which may have resulted in some unnecessary ringing as discussed later.

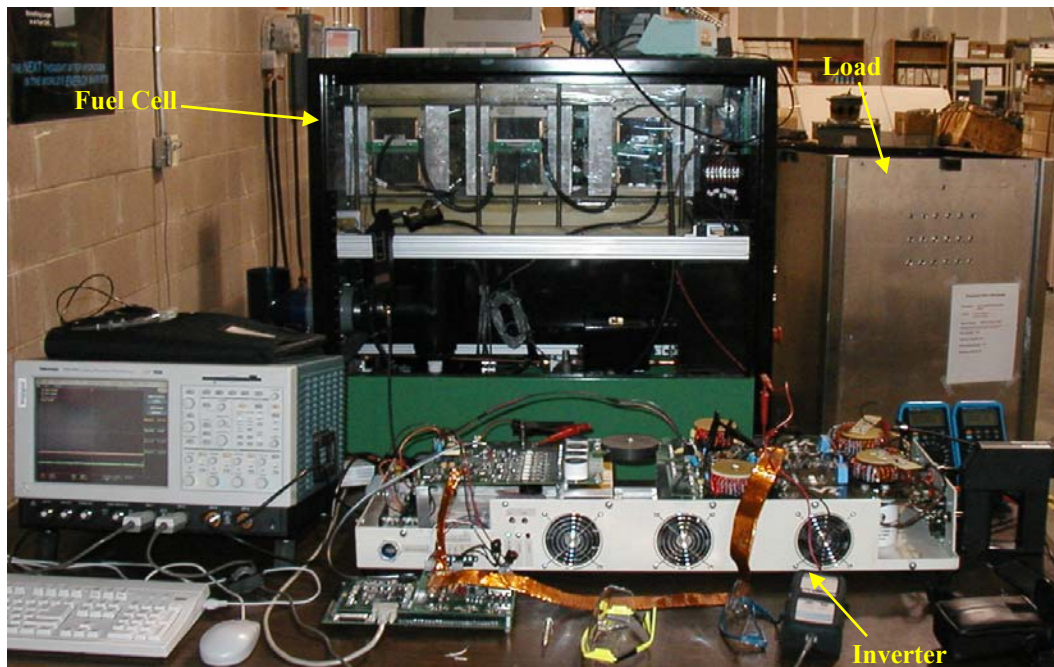


Figure 4.13 Fuel Cell Test Set-up

### 4.2.2 Input Waveforms

Several tests were performed in attempt to measure the steady-state and transient response of the fuel cell system. The load consisted of a light bulb bank and a small motor. The data for the transient responses was captured and graphed using Excel to help distinguish the results. Trend lines for the transient results are drawn in the darker colors.

Steady-state input and output waveforms for a resistive load are displayed in Figure 4.14. The average input voltage is about 50 volts with a 5-volt peak-to-peak ripple at approximately 120 Hz. The 10% voltage ripple may be due to the input source being soft or due to the ultracapacitor location, as this phenomena was not evident during the power supply testing (Figure 4.5). The input current drawn by the converter is shown as the blue waveform in Figure 4.14 and also exhibits a significant 120 Hz component, which is similar to what was seen with the power supply testing. The AC components of the input waveforms are potential problems for some fuel cells, but did not seem to cause any short term problems with this system. However, as shown in the subsequent plots, the ripple can increase dramatically during transient conditions.

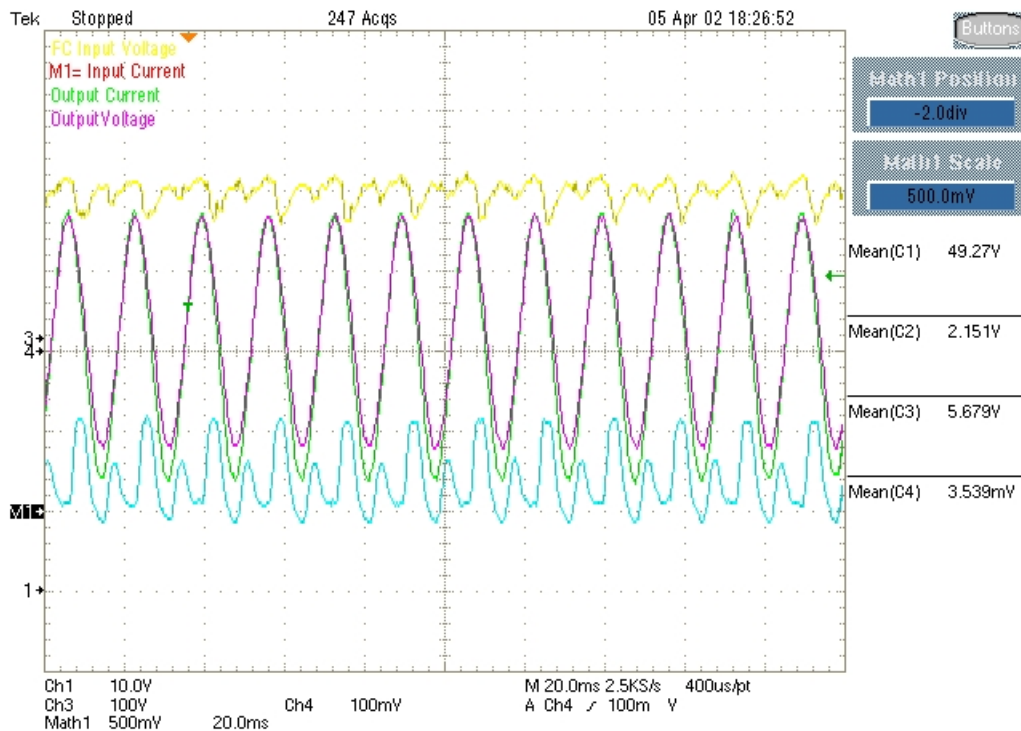
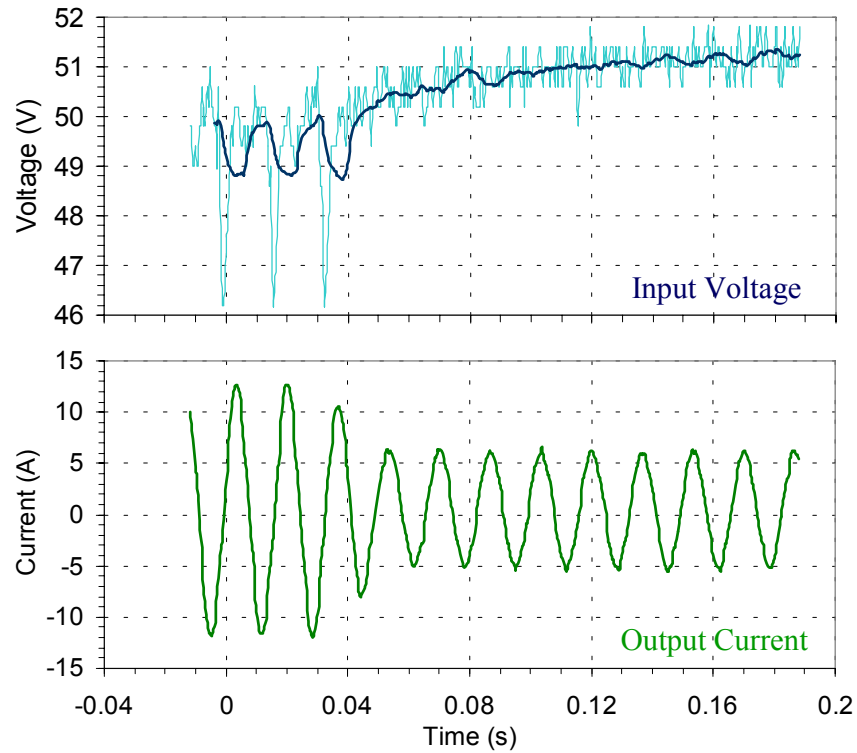


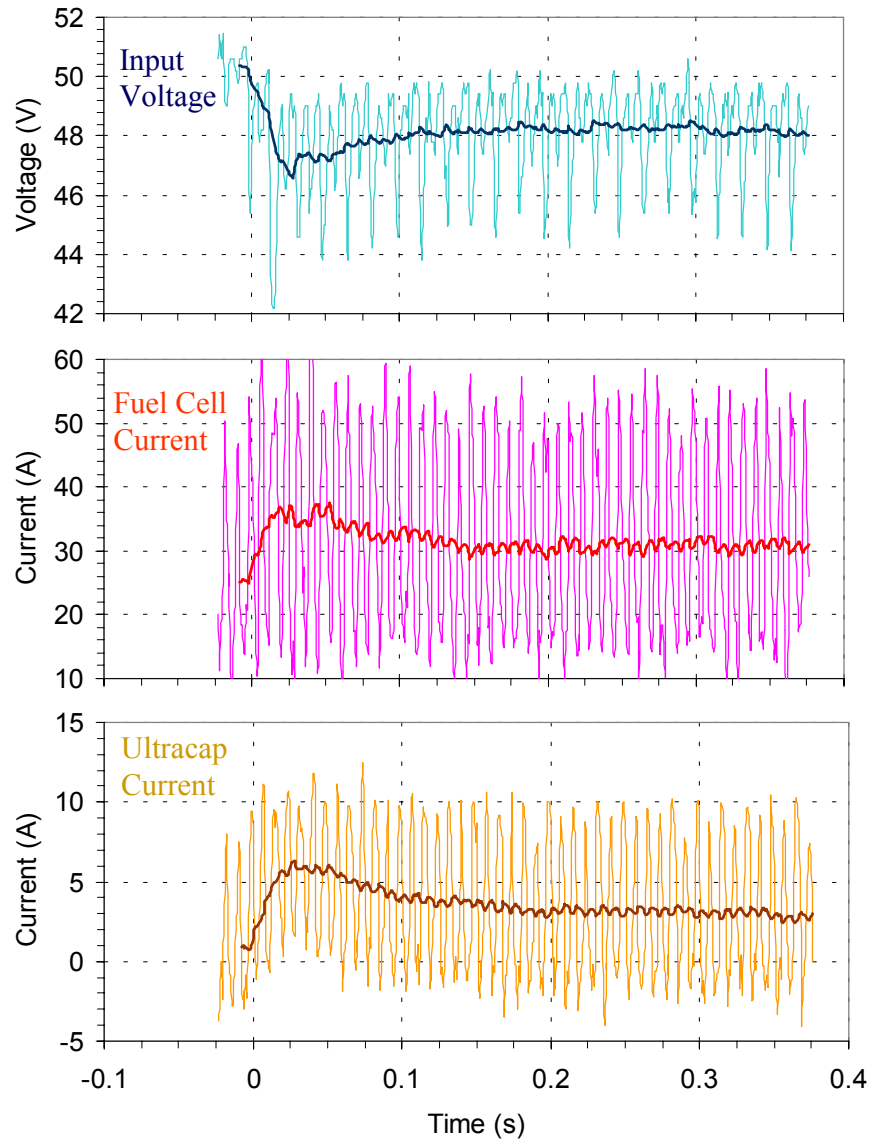
Figure 4.14 Fuel Cell Inverter Waveforms at 1.5 kW

Figure 4.15 shows the fuel cell voltage and the inverter output current during a 600 W load decrease. The input voltage increases about 2 V in 150 ms, and it seems to be still increasing slightly. It should be noted, that the steady-state fuel cell voltage ripple is relatively small, but the transient ripple is quite severe.



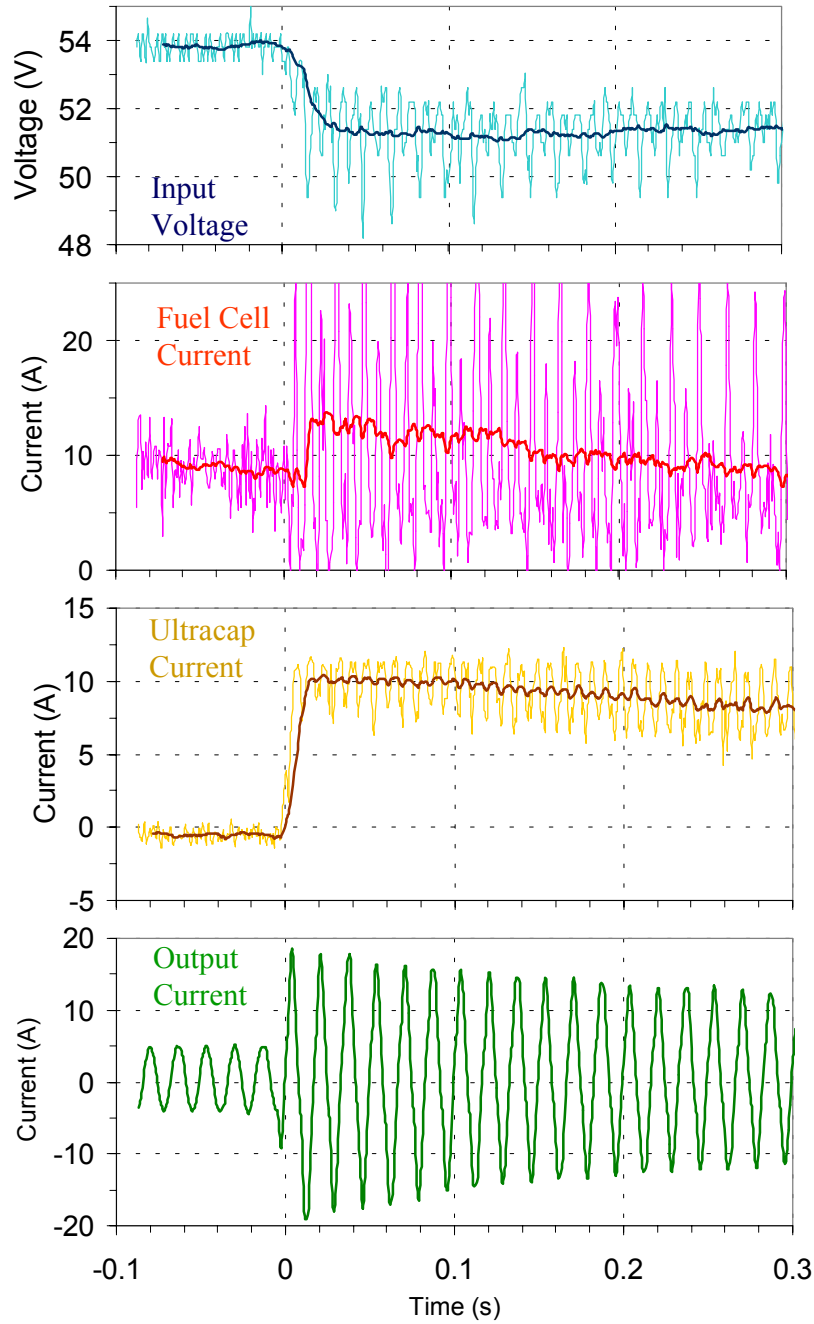
**Figure 4.15 Fuel Cell Load Transient – 1.1 kW to 500 W**

Shown in Figure 4.16 is the fuel cell voltage and current along with the ultracapacitor current during a load step of 500 W. The average input voltage starts around 51 V and then initially drops to about 47 V and finally rises back up to about 48 V for steady-state at 1.5 kW. The response time of the fuel cell current is relatively fast, but it is evident that the ultracapacitor helps out during the transient and then slowly ramps back down to around zero. The measured fuel cell ripple current under dynamic conditions is nearly 100% at a frequency of 120 Hz. The ultracapacitor current also exhibits a significant 120 Hz ripple.



**Figure 4.16 Fuel Cell Load Transient – 1 kW to 1.5 kW**

Figure 4.17 shows measured input voltage and currents and the output current during a motor transient. Two 200 W light bulbs were on initially, and at time zero the motor was turned on. The fuel cell voltage drops from 54 V to about 51 V, and the fuel cell current responds well by increasing almost immediately. However, it is also noticed that the ultracapacitor helps absorb the transient by providing about 10 amps instantaneously.



**Figure 4.17 Fuel Cell Load Transient – 400 W to Motor Start**

Again, the dynamic fuel cell voltage and current ripples are noticeably larger than the steady-state condition. It is apparent that the fuel cell response time to load changes is reasonably fast, but can contain large oscillations. The voltage time constant of the fuel cell can be estimated from to be about 2 ms. The input voltage waveforms in Figure 4.15, Figure 4.16, and Figure 4.17 all exhibit a frequency component around 120 Hz, which is twice the fundamental output frequency. This low frequency ripple may be problematic for some fuel cell systems, and depending on the magnitude and frequency, it can even cause cell degradation. Therefore, this ripple should be avoided if possible. The voltage ripple is caused by the non-ideality (sloping V-I curve) of the fuel cell source and the AC current drawn by the inverter. The long cable separating the ultracapacitor from the rest of the system and the large ESR of the ultracapacitor added unwanted impedance, not allowing maximum absorption of the current ripple. Adding an input filter inductor may help alleviate the current ripple problem, but a 10 kW filter inductor can be costly and bulky.

## Chapter 5 – Conclusion

### 5.1 Summary

The work presented in this thesis explored some of the topology and control options for a 10 kW fuel cell inverter system, and verified a low cost control strategy that still achieves adequate performance. Issues involving fuel cell characteristics and energy storage selection were discussed. Several power stage topology options were explored and compared, and the design of a prototype system was given.

Models of the fuel cell, the DC-DC converter, and the inverter were developed to aid in the control design and to analyze interactions between each subsystem. A simple voltage controller was designed and implemented to control the output of the phase-shifted full-bridge converter. Experimental transfer function measurements of the DC-DC converter were taken for various load conditions to confirm the modeling assumptions. Although the measured transfer functions varied slightly from the simulations, the converter was stable and performed well. The voltage regulation and transient response suffered during very light load conditions, because the converter was operating in DCM. However, according to the household load profile in Figure 1.1, operation of the system below 400 W will be a rare occurrence.

Two control strategies for the half-bridge inverter were analyzed in detail, and experimental results were compared for each case. It was shown that although theoretically possible, the open loop strategy does not meet the output voltage specifications. Therefore a continuous voltage feedback signal was employed to regulate the output voltage and improve transient response. Although steady-state errors can be a concern when using a sinusoidal reference, the AC output voltage was shown to regulate nicely over a wide range of loads.

The system was implemented using only voltage feedback, thereby eliminating the need for costly current sensors. Furthermore, a low cost DSP was used to do the cycle-by-cycle control, protect the system, and communicate with the user. Due to the minimal computation time required for the control algorithms, it would be feasible to incorporate the fuel cell controller into the same DSP and as a result reduce system cost. If the fuel cell needed a power request signal from the inverter, it would also be possible to estimate

the output power of the system from the duty cycle of the DC-DC converter or from the modulation index of the inverter, so output current sensing would not be required.

The final prototype was tested on a power supply with a wide variety of household loads, including light bulbs, computers, and power tools. A chart comparing the performance of the system to the specification from Chapter 1 is given in Table 5.1. The system was also tested on a 3.0 kW Enable™ fuel cell with ultracapacitor energy storage. During fuel cell testing, it was discovered that 1) the fuel cell time constant is reasonably fast, but the voltage can oscillate severely after load transients, 2) the DC bus bulk capacitors, which are normally sized for steady-state ripple, should be increased to help reduce the fuel cell current and voltage ripples during dynamic conditions, and 3) using ultracapacitors for supplementary energy storage is effective, but the ESR needs to be as small as possible to help alleviate the fuel cell voltage ripple.

**Table 5.1 Output Performance**

|                            | Measured         | Specification   |
|----------------------------|------------------|-----------------|
| Voltage Regulation         | 117 V $\pm$ 2 %  | 120 V $\pm$ 6 % |
| Frequency Regulation       | 60 $\pm$ 0.05 Hz | 60 $\pm$ 0.1 Hz |
| THD with Nonlinear Load    | 4.6 %            | 5 %             |
| THD with pf=0.8            | 2.9 %            | Not Given       |
| Overshoot (1 kW transient) | 4.2 %            | Not Given       |
| Efficiency                 | 85 %             | 90 %            |

## 5.2 Future Considerations

The research presented to this point clearly demonstrates the ability to implement a simple and cost effective control strategy on a medium power fuel cell inverter system. However the system is far from perfect. The chosen topology is not necessarily the best one, and the power stage design, albeit sufficient, was not optimized completely. There is ample room to improve the layout, tune the filter design, and optimize the switching frequencies. All of these changes will have an effect on the control design and possibly allow for improved performance.

The existing control parameters were chosen mainly from the simulation results, and they result in stable and relatively accurate performance. However, the measured transfer functions differed slightly from the simulated ones. Therefore by better understanding those differences, there may be some room for improvement of the transient response of

the system. It is also important to operate the system continuously at full power and even slightly above to see if there are any issues related to stability or performance.

The complete system performance, including fuel cell and ultracapacitors, definitely needs to be analyzed further. The fuel cell's non-ideal output impedance seems to cause some ripple effects not seen with a power supply. The 120 Hz voltage and current ripples present in the experimental results may not be allowed for many fuel cell systems. It seems reasonable that, if placed close to the inverter and if the ESR was small, the ultracapacitors could significantly attenuate the voltage ripple and help improve the current ripple. Assuming the models are accurate, simulations can verify this ripple reduction during steady-state conditions. However, the ultracapacitor performance during transient conditions is not as good as expected, which is mainly due to mismatches in time constants and impedances of the two energy sources. Therefore, it would be advantageous to take experimental data without the ultracapacitors and do a detailed comparison. If the current ripple is not reduced enough, there are a few possible solutions that could be explored. First an input filter could be added that might help smooth out the current. The second option would be to increase the DC bus capacitance, allowing less ripple current to be reflected to the input. Lastly it may be advantageous to investigate control options or improvements that may help decrease the ripple effects. Finally, it would be beneficial to test the inverter on different fuel cells, especially ones that exhibit a slower transient response.

## Appendix A – ITI (CBEMA) Curve

### A.1 Scope and Application

The Computer Business Manufacturers Association (CBEMA, now the ITI Council) established a voltage profile to aid manufacturers in the design of protection circuits during the early 1980's. Since then, the profile has become a standard reference within the industry for steady-state and transitory conditions.

The “standard” basically describes an AC input voltage envelope that typically can be tolerated by most information technology equipment. It is intended to serve as a guideline, not a specification, for products and for AC distribution systems. The curve and application note are mainly aimed at single-phase 120 V, 60 Hz nominal systems.

### A.2 The Curve

The curve shown in Figure A.1 is an approximation of the actual ITI curve. The actual curve can be found in [43]. It shows the acceptable region of voltage deviation from the nominal value versus the duration of the deviation. The steady-state conditions are considered to be  $\pm 10\%$ .

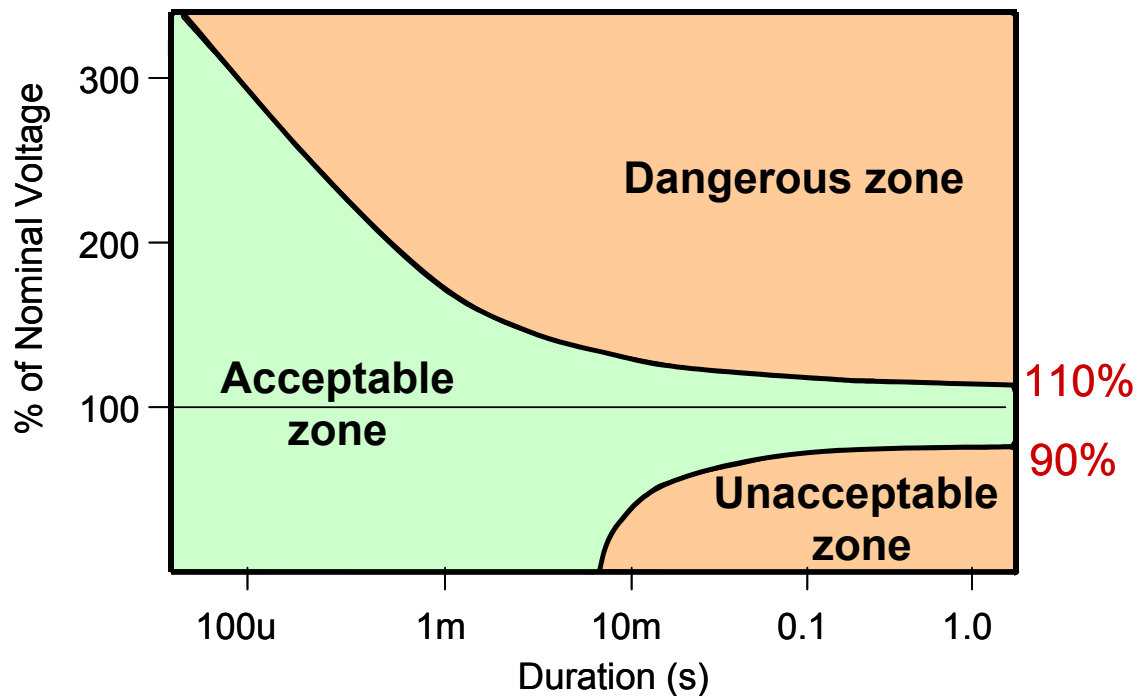


Figure A.1 Approximate ITI Curve

## Appendix B – Parts List

A parts list of the major components used in the prototype inverter is shown in Table B.1. Estimated pricing information based on large quantities is included in the last column.

**Table B.1 Parts List and Estimated Price Information**

| Estimated Inverter Cost Based on Available Price Quotes |                             |                     |                  |                         |     |                    |
|---------------------------------------------------------|-----------------------------|---------------------|------------------|-------------------------|-----|--------------------|
|                                                         | Component                   | Manufacturer Part # | Manufacturer     | Price in 10000 Qty (\$) | Qty | Extended Cost (\$) |
| <b>POWER SEMICONDUCTORS</b>                             |                             |                     |                  |                         |     |                    |
|                                                         | 85 V, 180 A (130 A) MOSFET  | IXFX180N085         | IXYS             | 3.9                     | 10  | \$ 39.00           |
|                                                         | 600 V, 100A (60 A) IGBT     | IRG4PSC71UD         | IR               | 7.8                     | 4   | \$ 31.20           |
| <b>DIODES</b>                                           |                             |                     |                  |                         |     |                    |
|                                                         | 1200 V, 32 A Dual Ultrafast | HFA32PA120C         | IR               | 3.5                     | 8   | \$ 28.00           |
| <b>GATE DRIVES</b>                                      |                             |                     |                  |                         |     |                    |
|                                                         | 2A w/ DSAT                  | HCPL-316J           | Agilent          | 3.32                    | 12  | \$ 39.84           |
| <b>INDUCTORS</b>                                        |                             |                     |                  |                         |     |                    |
|                                                         | 500uH MPP Core              | 55104-A2            | Magnetics        | 4.5                     | 8   | \$ 36.00           |
|                                                         | 330 uH MPP Core             | 55866-A2            | Magnetics        | 13                      | 4   | \$ 52.00           |
| <b>TRANSFORMER</b>                                      |                             |                     |                  |                         |     |                    |
|                                                         | Ferrite Pot Core            | Custom              | Ceramic Magnetic | 16                      | 2   | \$ 32.00           |
| <b>CAPACITORS</b>                                       |                             |                     |                  |                         |     |                    |
|                                                         | 400 V, 50u Polypropolene    | UL30AX0400          | Elcon            | 20                      | 2   | \$ 40.00           |
|                                                         | 80 V, 15000u Electrolytic   | ECET1KA153FA        | Panasonic        | 5.45                    | 1   | \$ 5.45            |
|                                                         | 250 V, 2200u Electrolytic   | ECET2EA222EA        | Panasonic        | 6                       | 4   | \$ 24.00           |
|                                                         | 600 V, 0.15u Polypropolene  | 376KP/MMKP          | Phillips         | 0.43                    | 2   | \$ 0.86            |
|                                                         | 100 V, 50u Polypropolene    | 3MP-16219K          | Elcon            | 4                       | 2   | \$ 8.00            |
| <b>CONTROL</b>                                          |                             |                     |                  |                         |     |                    |
|                                                         | DSP                         | ADMC401             | Analog Devices   | 18                      | 1   | \$ 18.00           |
|                                                         | Phase Shift Controller      | UC3895              | Unitrode         | 4.8                     | 1   | \$ 4.80            |
| <b>AUX POWER SUPPLY</b>                                 |                             |                     |                  |                         |     |                    |
|                                                         | 12 V Top-Switch Circuit     | NA                  |                  | 4                       | 1   | \$ 4.00            |
|                                                         | 15 V Top-Switch Circuit     | NA                  |                  | 4                       | 1   | \$ 4.00            |
|                                                         | 5 V Top-Switch Circuit      | NA                  |                  | 4                       | 1   | \$ 4.00            |
|                                                         | Isolated Gate Drive Supply  | NA                  |                  | 1                       | 5   | \$ 5.00            |
| <b>SENSING</b>                                          |                             |                     |                  |                         |     |                    |
|                                                         | Linear Optocoupler          | HCNR201-300         | Agilent          | 2.4                     | 4   | \$ 9.60            |
|                                                         | Instrumentation Amplifier   | INA126U             | TI               | 1.1                     | 4   | \$ 4.40            |
|                                                         | Op Amp                      | LM833M              | National Semi    | 0.32                    | 15  | \$ 4.80            |
| <b>PCB</b>                                              |                             |                     |                  |                         |     |                    |
|                                                         | 6 Layer Inverter Board      | Custom              |                  | 4.5                     | 1   | \$ 4.50            |
|                                                         | 4 Layer Front End Board     | Custom              |                  | 2                       | 1   | \$ 2.00            |
|                                                         | Control Board               | Custom              |                  | 2                       | 1   | \$ 2.00            |
|                                                         | 14 Layer Transformer Board  | Custom              |                  | 15                      | 2   | \$ 30.00           |
| <b>PACKAGING</b>                                        |                             |                     |                  |                         |     |                    |
|                                                         | 13" Bonded Fin Heat Sink    | 420003 u2 1500      | Aavid            | 4                       | 3   | \$ 12.00           |
|                                                         | Case                        | Custom              |                  | 15                      | 1   | \$ 15.00           |
| <b>FANS</b>                                             |                             |                     |                  |                         |     |                    |
|                                                         | 120 Vac, 9W                 | 3906                | EBM PAPST        | 6                       | 3   | \$ 18.00           |
| <b>MISC</b>                                             |                             |                     |                  |                         |     |                    |
|                                                         | 2.5 V, 4 F Ultracap         | PC5                 | Power Cache      | 2                       | 4   | \$ 8.00            |
| <b>MANUFACTURING</b>                                    |                             |                     |                  |                         |     |                    |
|                                                         | Labor, Resources, ect.      |                     |                  |                         |     | \$ 8.00            |
| <b>Total</b>                                            |                             |                     |                  |                         |     | <b>\$ 494.45</b>   |

## Appendix C – Small Signal Model

### C.1 Phase-Shifted Full-Bridge

The small signal model of the phase-shifted full bridge circuit as developed in [33] is shown in Figure C.1.

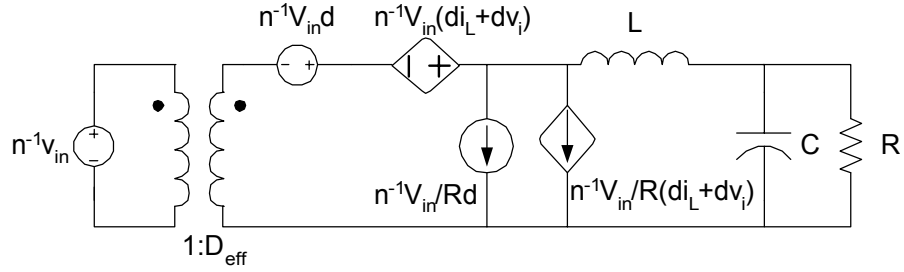


Figure C.1 Small Signal Model of the Phase-Shifted Full-Bridge

## Appendix D – Saber Models and MAST Code

### D.1 Fuel Cell

The Saber MAST code for the nonlinear fuel cell model is shown below, where the variable *in* represents the fuel cell current and *out* represents the fuel cell terminal voltage. The Saber circuit used to test the nonlinear model is shown in Figure D.1.

```

=====
# Template to calculate Fuel Cell V-I Curve
# Ver 1.0 by C.Liu Nov. 14, 2000
# Modified by T. Nergaard May 14, 2002 & June 17, 2002
=====
element template lookuptable in out
input nu in
Output nu out
{
  val nu i,o,tmp
  values {
    if (in <= 2) tmp = 1.6502*in*in-8.3461*in+69.4715
    else if (in <= 60) tmp = 59-0.25*in
    else tmp = -0.0804*in*in+8.8335*in-196.48
    o = tmp}
  equations {
    out = o}
}

```

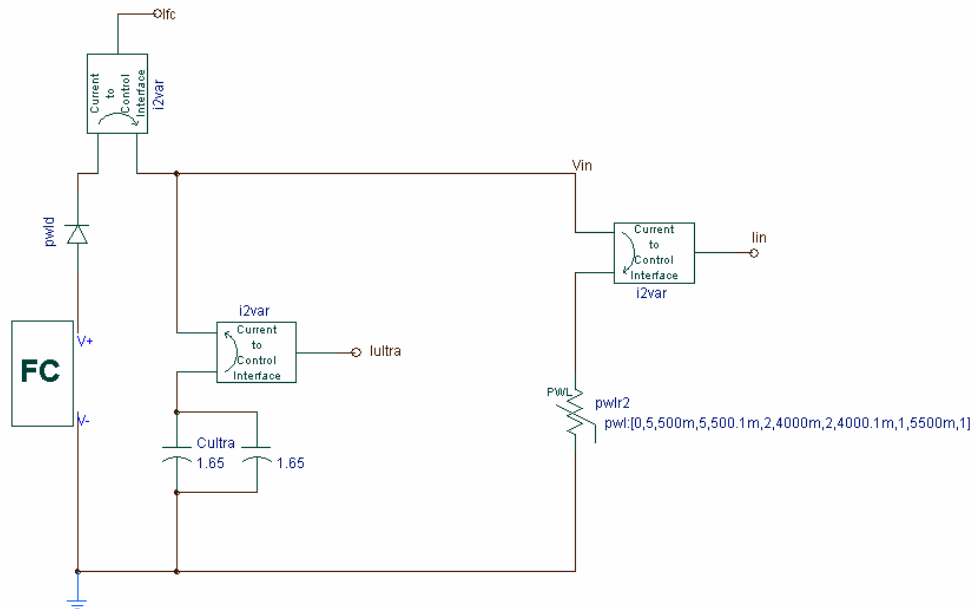
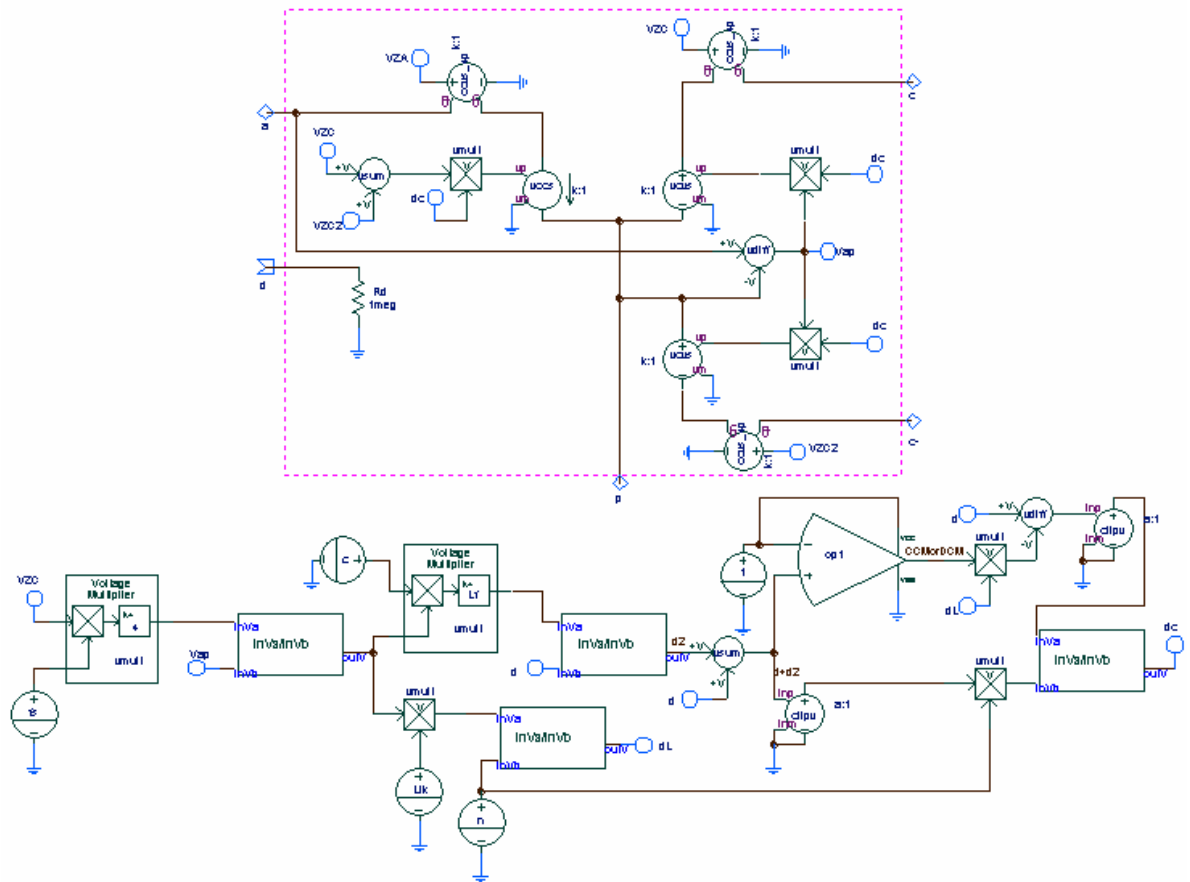


Figure D.1 Saber Circuit of Nonlinear Fuel Cell

### D.2 Phase-Shifted PWM Switch

The Saber implementation of the average switch model for the phase-shifted full-bridge converter is shown in Figure D.2. The dotted box shows the basic four terminal switch and the components below it are used to implement the equations based on what mode (DCM or CCM) the converter should be operating in.

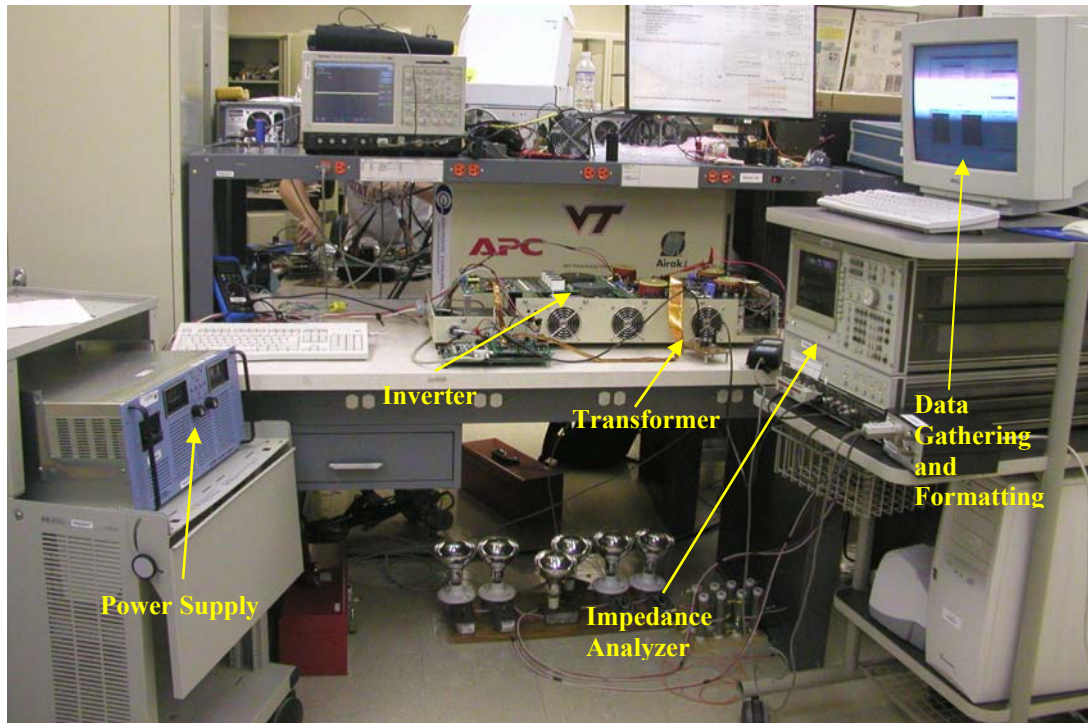


### Figure D.2 Saber Model of Phase-Shifted Switch

## Appendix E – Transfer Function Measurements

### ***E.1 Test Set-up***

The tests varied slightly for each transfer function, but a picture of the basic test set-up is shown in Figure E.1. The impedance analyzer used was a HP4149, and a transformer was used to inject the perturbation into the circuit.



**Figure E.1 Transfer Function Measurement Bench Set-up**

### ***E.2 Control to Output***

The open loop control to output transfer functions were measured by AC coupling the perturbation into the DC control signal. A small transformer was used to isolate the impedance analyzer from the rest of the circuit. A differential probe was used to measure the test voltage on the output of the converter. A diagram of the configuration is shown in Figure E.2.

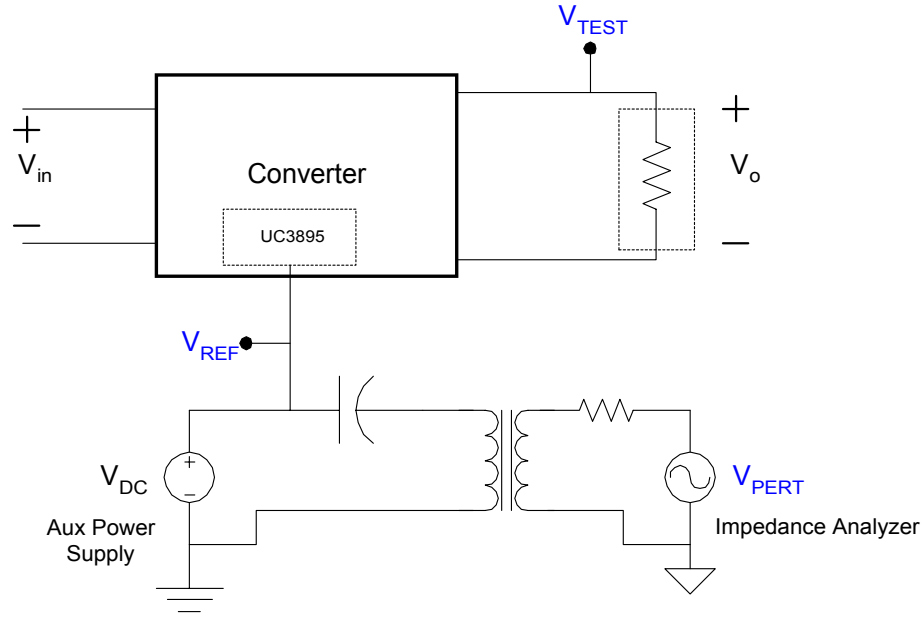


Figure E.2 Control to Output Set-up

### E.3 Output Impedance

The output impedance measurement, shown in Figure E.3, was much more complicated due to the high voltage output of the converter. The impedance amplifier had trouble directly driving the output of the converter. Therefore an audio amplifier was inserted before the converter to boost the amplitude of the signal. This provided enough power to perturb the current, but caused the isolation transformer to saturate at low frequencies. Consequently, the measurements could only be taken as low as 50 Hz. The DC blocking capacitor was 500 V, and the current perturbation was measured with a current probe.

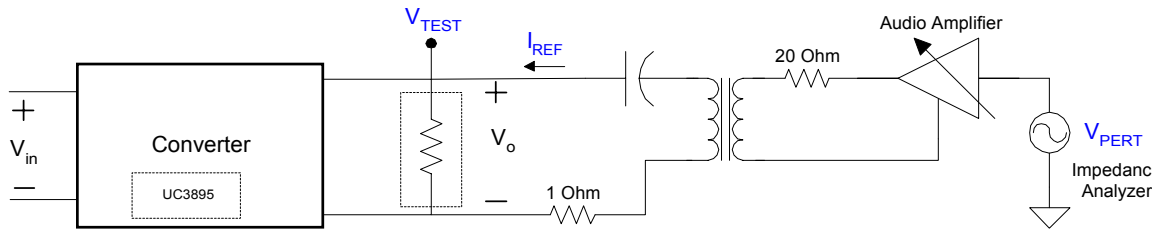


Figure E.3 Output Impedance Set-up

## E.4 Loop Gain

The closed loop gain measurement involved using the same transformer from the control to output measurement and basically injecting a signal inline with the feedback. An illustration of the measurement is shown in Figure E.4.

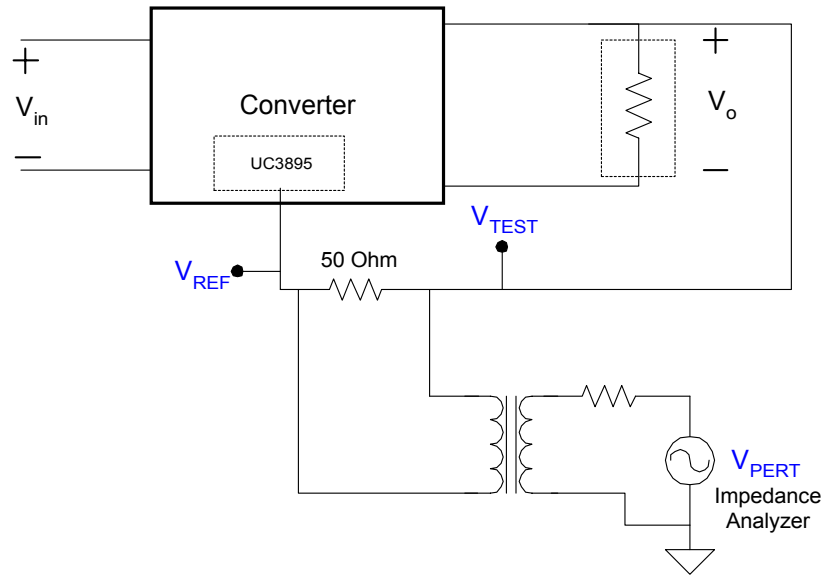


Figure E.4 Loop Gain Set-up

## E.5 Audio Susceptibility

To perturb the input voltage, a MOSFET was placed in series with the input power supply. The MOSFET operated in the linear region and was biased by the input voltage. The AC signal from the impedance analyzer was coupled into the gate via a capacitor, as shown in Figure E.5. There was a considerable voltage drop across the MOSFET, so the circuit could not be operated above about 200 W.

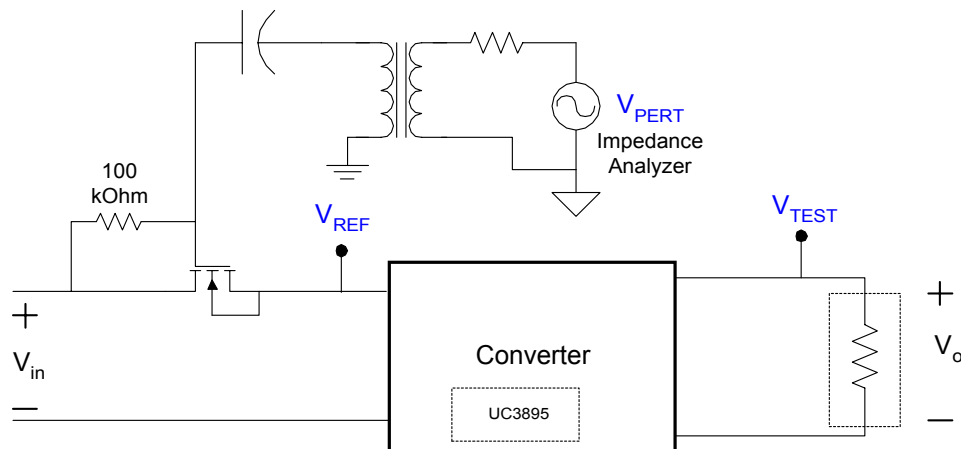


Figure E.5 Audio Susceptibility Set-up

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## **Vita**

Troy A. Nergaard was born in Marinette, Wisconsin in October of 1976. He graduated valedictorian from Eau Claire North High School in June 1995. Troy went on to pursue an engineering degree from the University of Wisconsin-Madison, where he was awarded the Bachelor of Science in Electrical and Computer Engineering in May 2000. In the fall of 2000 he entered the graduate program at Virginia Polytechnic Institute and State University, where he was awarded a Bradley Fellowship. During his time at Virginia Tech, Troy worked as a research assistant in The Center for Power Electronics Systems with research focused in the areas of fuel cell power conversion, high power DC-DC converters, and DSP control. Troy will begin full-time employment with the Energy Systems Development Center of AeroVironment, Inc. upon completion of his Master of Science degree.