

DESIGN AND ANALYSIS OF A HIGH
FREQUENCY RESONANT SEPIC
BIDIRECTIONAL CONVERTER AND THE
ASSOCIATED BATTERY MODELING FOR
BATTERY EQUALIZATION APPLICATIONS

by

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ABSTRACT

In this thesis, a high frequency resonant SEPIC bidirectional converter is proposed that has applications in battery equalization and charging/discharging. Motivations and applications for battery equalization are explored. Previous work on battery equalization is presented, and the benefits and challenges of high frequency power electronics are explored.

The proposed high frequency converter is small in size due to reduced size of the magnetic components. The design, simulation, and experimental procedure is outlined. Experimental results show that the efficiency is greater than 81%. Ways to improve the efficiency are also explored and mentioned.

This thesis also presents a new electrical circuit topology that can be used to accurately model batteries. The proposed model has only two circuit components, making it very simple when compared to some of the popular battery models used today. Also, the method to formulate the nonlinear component values that vary based on the current state of the battery is very straightforward because no optimization software or long battery tests are required. In this thesis, the new battery model is proposed and its performance is compared with that of the four most common electrical circuit battery models that are used today. Model formulation and the required battery testing is also detailed and explained. It is shown that the performance of the proposed model exceeds that of the common models.

The work of this thesis can be seen as a contribution toward improving the current state of battery management systems (BMS). Batteries are fragile and complex systems that need careful management in order for battery technology to be usable and sustainable. It is

believed that this thesis laid some of the groundwork necessary to improve current BMS. The modeling will help predict and monitor battery performance. The implementation of a RF DC-DC converter will decrease the size, improve the possibilities of power electronic integration with battery cells, and decrease the cost.

Key words: batteries, electrical model, lithium-ion, zero voltage switching, resonant DC-DC converter, battery equalizer, SEPIC converter, high frequency

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Chapter 1. Background

1.1 Public Policy Trends

The desire for improved energy efficiency in areas such as the electric power grid and automobile technology in recent years has led to an increase in power electronics research. States such as California have mandated that 33% of its energy come from renewable energy by 2020 [1]. Recently, the Obama administration passed new standards that increase fuel economy to the equivalent of 54.5 MPG for cars and light-duty trucks [2]. This was done to protect the environment and to encourage innovation and investment in advanced technologies. This new program also includes incentives for automobile manufacturers to make advancements in electric vehicles (EV) and plug-in hybrid electric vehicles (HEV). In order to implement renewable energy into the grid or develop EVs and HEVs, it is essential to have an efficient battery equalizer [1].

1.2 Lithium-Ion Technology

Lithium-ion based energy storage is becoming the most popular form of energy storage system. Lithium-ion technology is currently the chemistry of choice due to its high power and energy density, high working cell voltage, low self-discharge rate, and high charge-discharge efficiency [3]. Table 1.1 compares some of the key characteristics between Lead-acid, the previous chemistry of choice, and Lithium-ion technologies. In addition to what is listed in Table 1.1, when compared to Lead-acid, Lithium-ion technology has a faster

recharge time, has a higher cycle count before failure, and can withstand higher temperatures [3].

Table 1.1 Lead Acid and Lithium-Ion Comparison [3].

	Lead Acid	Lithium-Ion
Nominal Cell Voltage (V)	2.0	3.7
Wh/kg	35-40	140-150
Wh/liter	70	400
Size of 1 kWh battery	14 Liters	2.5 Liters
Weight of 1 kWh battery	25 kg	6.7 kg
Size of 1 MWh battery	14,285 Liters	2,500 Liters
Weight of 1 MWh battery	25,000 kg	6,700 kg
Time between service	6 months	24-36 months
Replacement timeframe	1.5-2 years	5-7 years

A downside of Lithium-ion technology is that careful management is required because these batteries cannot withstand an over-charged state under the high-risk of explosion. In other words, battery equalizers are essential for energy storage systems that use Lithium-ion batteries. Because more complex circuitry is needed for Lithium-ion technology, another downside is the cost. Lithium-ion technology is generally 4 times more expensive per kWh than Lead-acid technology [3]. A123 Systems has produced a variation of the common Lithium-ion technology called Nanophosphate Lithium-ion [4]. This variation is supposed to improve safety performance as well as extend cycle life.

1.3 Vehicle Applications

In order to maintain high performance, a battery management system (BMS) is needed for Lithium-ion energy storage systems. The BMS is used to ensure that all cell voltages are kept in boundaries for safe operation and cycle life [1]. A BMS that is used for

an EV or HEV must be able to estimate the remaining range of the vehicle, to estimate the power capability, and to monitor the health of the batteries [5]. Health monitoring can include tracking the voltage, temperature, and charge and discharge current. These parameters can then be used to estimate the State of Charge (SOC) and State of Health (SOH) of a battery. The SOC is the percentage of charge that is left in the battery. The SOH definition can be somewhat arbitrary and can consider factors such as internal resistance, capacity, and cycle count. Inherent in the BMS are battery equalizers.

1.4 Grid Storage Applications

Energy storage is necessary to obtain the goal of creating smarter, cleaner and more reliable power grids. Grid energy storage can increase system efficiency, enable integration of renewable sources, improve grid stability and reliability, and increase energy security [4]. Grid storage also has promising applications in frequency regulation and dynamic stability support [4]. Figure 1.1 shows a future view of the power grid. The battery banks, and therefore battery equalizers, are essential to make this view reality.

1.5 Outline of Thesis

Chapter One has provided the most elementary background and applications for battery equalizers. Public Policy and needs in important areas such as electric vehicles and the power grid make battery equalizers and battery management systems necessary.

Chapter Two looks at the state of the art of battery equalizer technology and very high frequency converter design. Different equalization techniques are explored and the problems with some of the current solutions are mentioned.

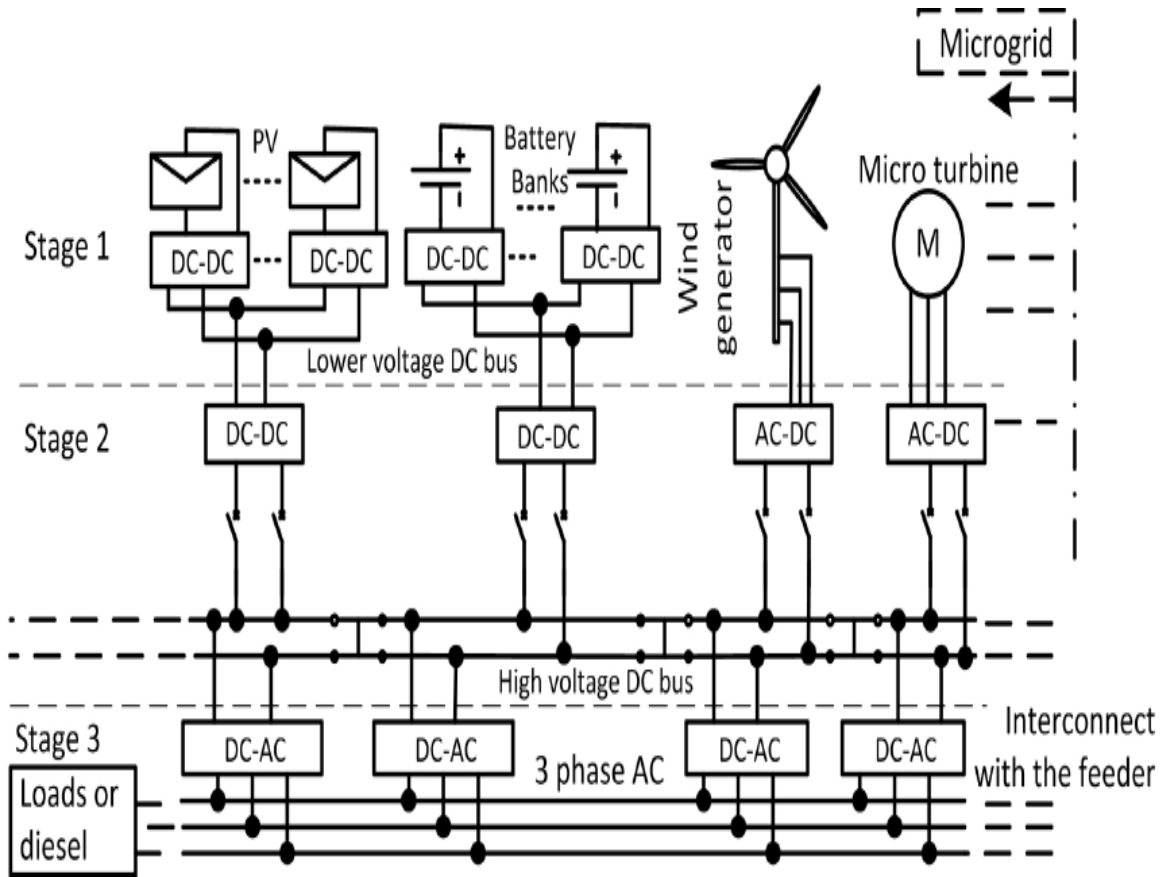


Figure 1.1 Future View of the Power Grid [55].

The benefits and challenges of very high frequency converter design are also explored.

Chapter Three introduces a new battery model and presents promising results that show the proposed model has better performance than the four most common electrical circuit models. These four models are explained and the method to formulate the model parameters is detailed and explained. It is thought that an accurate battery model will improve the performance of a battery management system, which needs to know information about the current state of the battery. In practice, the model should be complex enough for the BMS to describe the accurate status of all the cells. However, it should be simple enough

to reduce the computation time and characterization effort. It is believed that the proposed model provides a good balance between simplicity and accuracy.

Chapter Four introduces a high frequency resonant SEPIC bidirectional DC-DC converter that can be used for equalization and charger applications. Design, simulation and experimental results are presented.

Chapter Five concludes the thesis and proposes future work.

Chapter 2. **State of the Art Review**

2.1 Battery Equalizer

In the following subsections, the concepts of battery equalization are explored in more detail.

2.1.1 Importance of Battery Equalizers

As mentioned previously, battery equalizers are essential to the performance of EVs and HEVs and the integration of energy storage into the grid. This is especially true since the Lithium-ion technology is becoming more prevalent in new energy storage systems. It can be argued that battery equalization is the most important aspect of a BMS [12]. Due to manufacturing variance, aging degradation, imbalance between charging and discharging, and differences in thermal conditions, internal impedances, and the self-charging rate, the energy stored in any two batteries is not equal. These mismatches reduce the efficiency, capacity, and lifetime of the batteries and can cause safety concerns. Battery equalizers can reduce this mismatch and improve performance, thus making equalizers a very important aspect of future battery management systems [12-15].

2.1.2 Problems of Classic Equalizer Technology

Most types of equalizers suffer from slow equalization speed as the number of batteries connected in the system becomes large. For example, in hybrid electric vehicle applications with greater than 80 Lithium-ion series connected cells, the equalization time can approach 3 hours [14]. A traditional equalizer is shown in Figure 2.1. Notice that there is

only one way for energy to transfer from one battery to another. Therefore the equalization time is limited by the capacity of a single equalizer unit.

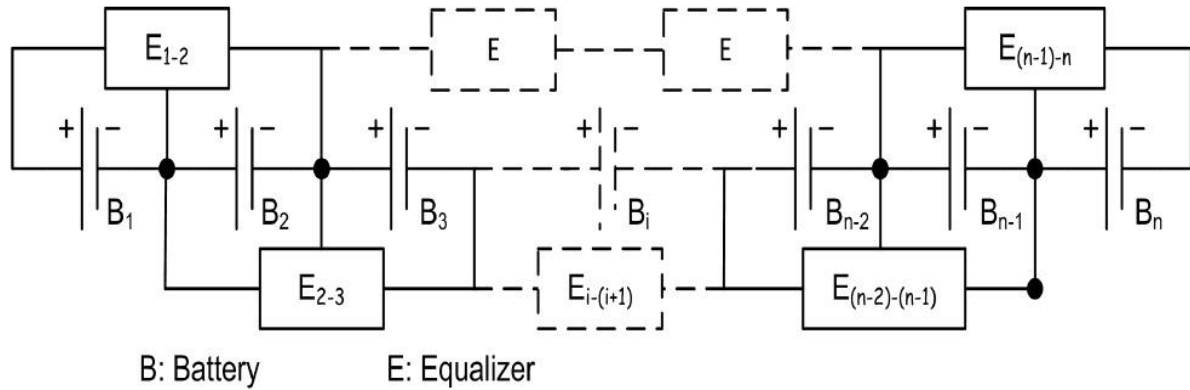


Figure 2.1 Traditional Battery Equalizer Architecture.

Another problem that current equalizer technology suffers from is bulky size. It is one of the goals of this research to reduce the size of battery equalizers. The proposed resonant circuit topology will allow for a reduction in size and cost because the size of the magnetic components can be reduced. This idea is explored in Section 2.2 in more detail.

2.1.3 Different Types of Equalization Methods

A comprehensive summary of different active balancing/equalization methods is done in [12]. The findings will be summarized in the following subsections.

2.1.4 Active vs. Passive Balancing

Passive balancing methods can only be used for Lead-acid and Nickel based batteries because these batteries can be brought into overcharge conditions without permanent damage. The excess overcharge energy is released in the form of heat or by gassing via the gassing valve. However, as the number of series cells grows, overcharge balancing is no

longer effective. Therefore, this method is a cost effective solution for low voltage Lead-acid and Nickel based battery applications.

Active balancing uses external circuitry to control the energy transfer among cells. This method can be used for most any battery technology because the characteristics of the battery chemistry have no effect on the performance of the external circuitry.

2.1.5 Shunting Active Balancing

This method involves the removal of excess energy from higher voltage cells in order to bring the voltage of these cells down to the level of the lower voltage cells. It is generally agreed that this is the most straightforward concept of battery equalization. Shunting equalization can either be dissipative or non-dissipative.

One of the most common balancing methods employs a dissipative resistor. Figure 2.2 shows the dissipative resistor equalizer circuit topology. The voltage monitors determine if the shunting resistor needs to be connected to the cell so that excess energy can be removed. The value of the resistor is chosen based on application and battery capacity. This is not a very efficient battery equalization method, but it provides a low cost solution.

Analog dissipative shunting is similar to dissipative shunting except that transistors are used instead of resistors as the shunting element. The topology is shown in Figure 2.3. When a cell reaches the maximum charge voltage, the current is proportionally shunted around the cell and through the transistor. This is done until the final cell reaches the maximum charge voltage. This method is more cost effective than the resistive dissipative shunting method because complex control is not needed.

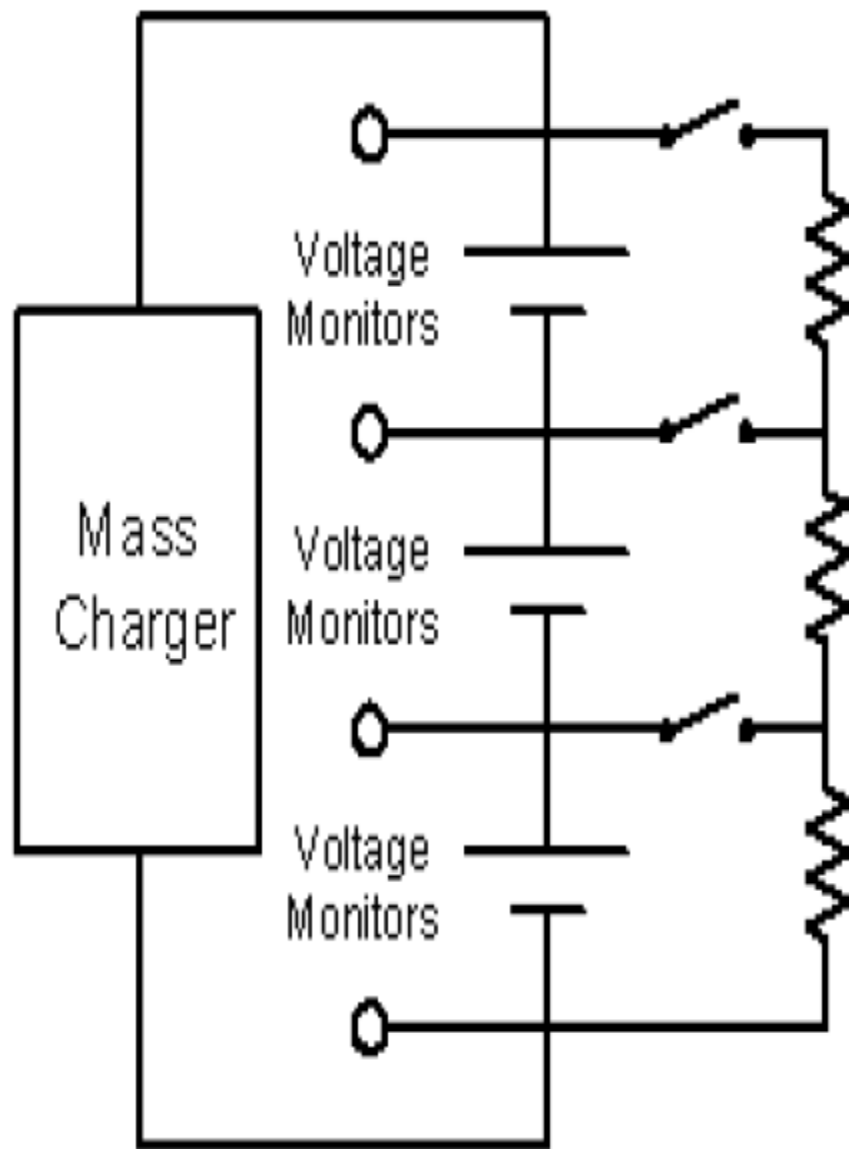


Figure 2.2 Dissipative Resistor Shunting [12].

PWM controlled shunting is a non-dissipative shunting method. The topology is shown in Figure 2.4. In this method, the BMS senses a voltage difference between neighboring cells and controls the PWM signals in order to affect the direction of average current flow. This is more efficient than dissipative equalization, but the control is complex. $2(n-1)$ switches and $(n-1)$ inductors are needed for n series-connected cells.

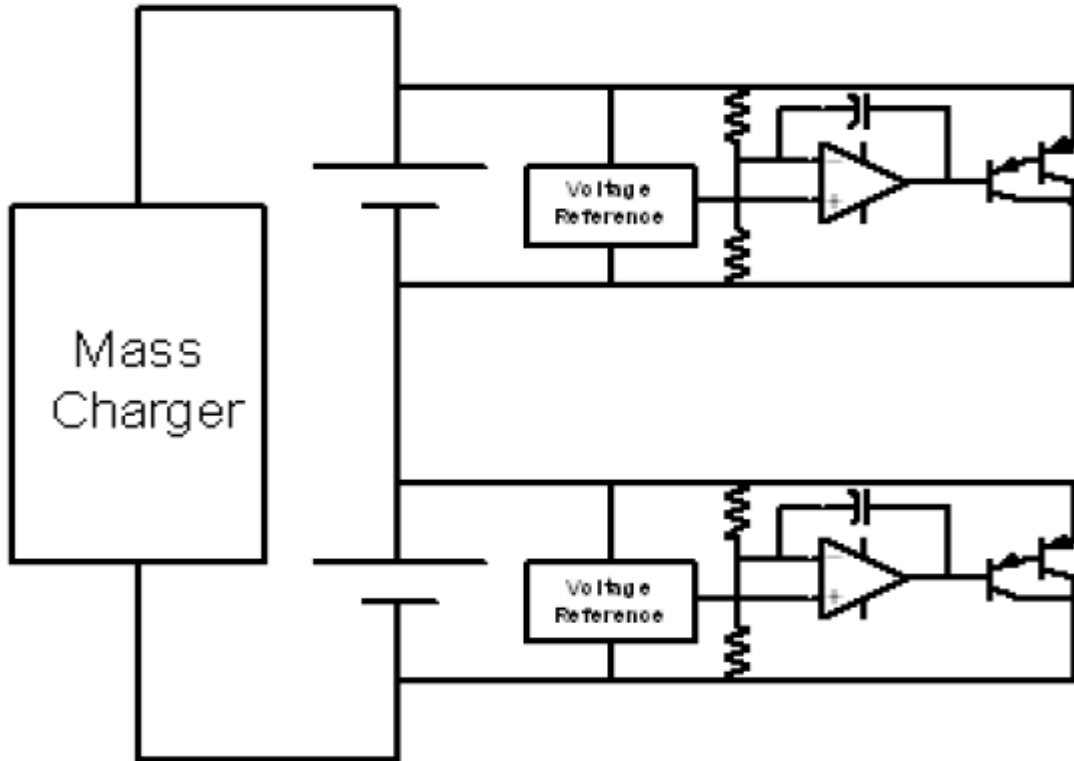


Figure 2.3 Dissipative Analog Shunting [12].

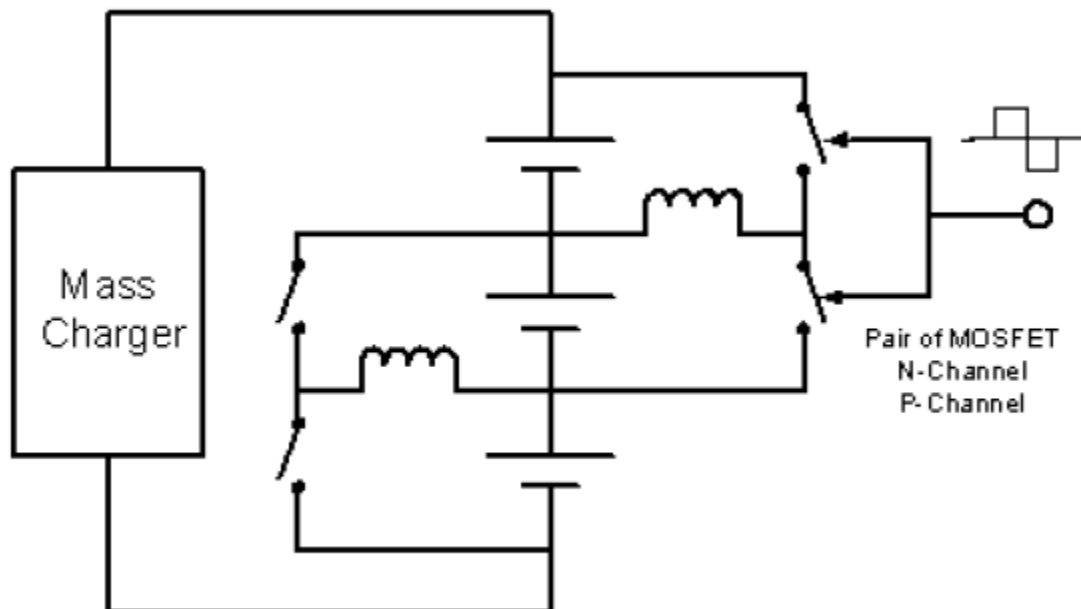


Figure 2.4 PWM Controlled Shunting [12].

The actual topology of PWM controlled shunting can vary. For example, Figure 2.5 shows the boost shunting method, which uses PWM to control the switches. The switch with the higher voltage will be activated. The circuit can be seen as a boost converter, which diverts the excess energy to other cells in the system.

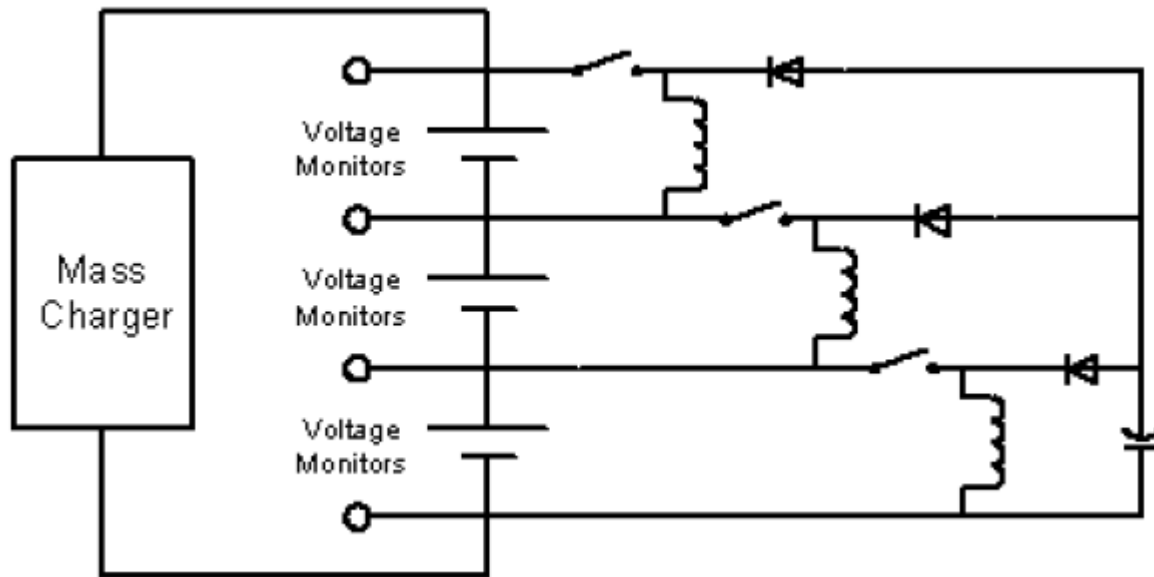


Figure 2.5 Boost Shunting [12].

2.1.6 Shuttling Active Balancing

Shuttling balancing utilizes energy storage devices, such as capacitors, to transfer energy among cells.

A switched capacitor topology is shown in Figure 2.6. In this setup, $2n$ switches and $(n-1)$ capacitors are required to balance n series-connected cells. Control is simple because there are only two states. For example, in the first state, C_1 and B_1 will be connected in parallel. C_1 will be charged or discharged to obtain the same voltage as B_1 . In the second

state, C1 will be connected in parallel with B2. The same thing happens in this state that happened in the first state. After this is complete, B1 and B2 will be equalized. The same thing happens with C2, B2, and B3. Intelligent control is not needed for this topology. It has applications in HEVs because it works well for both charging and discharging.

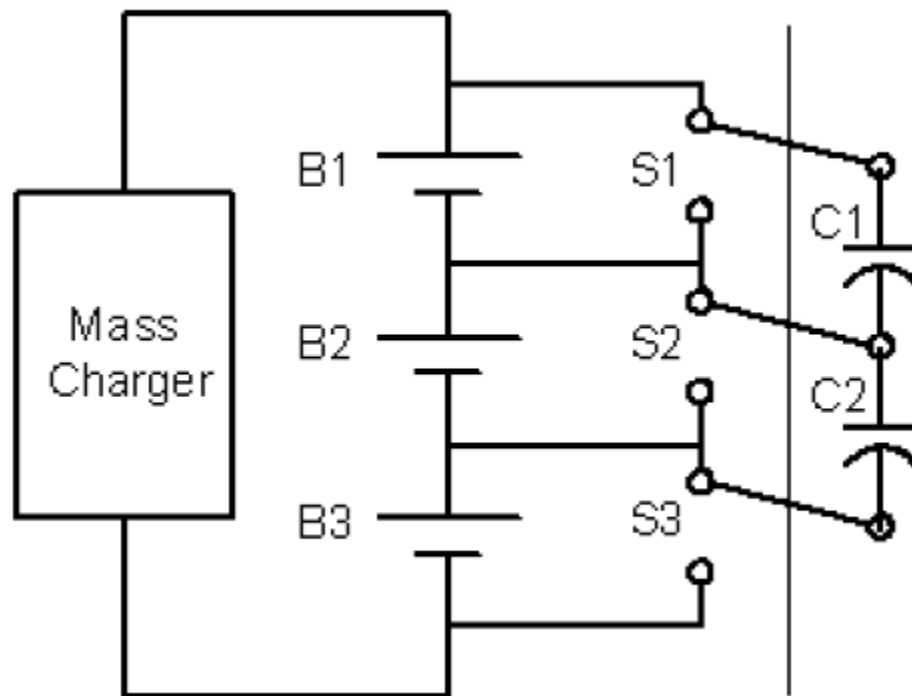


Figure 2.6 Switched Capacitor Shuttling [12].

There is a derivation of the topology shown in Figure 2.6 that involves using a single capacitor. In this setup, n switches are needed. As a result, the control can be much more complex.

2.1.7 Energy Converter Active Balancing

In this sense, energy converters are defined as isolated converters, meaning that the input and output sides have isolated grounds.

A step up converter balancing topology is shown in Figure 2.7. Isolated boost converters remove excess energy from cells and transfer it to the total pack. The input of each converter is connected to the each individual cell. Intelligent control is used to control the battery equalization. This method tends to be expensive due to complex control and the fact that each cell needs its own converter. It is suitable for modular design, but design can become tricky when the number of cells in the system increases. This is because each individual step up converter needs to boost a single cell voltage to the total pack voltage.

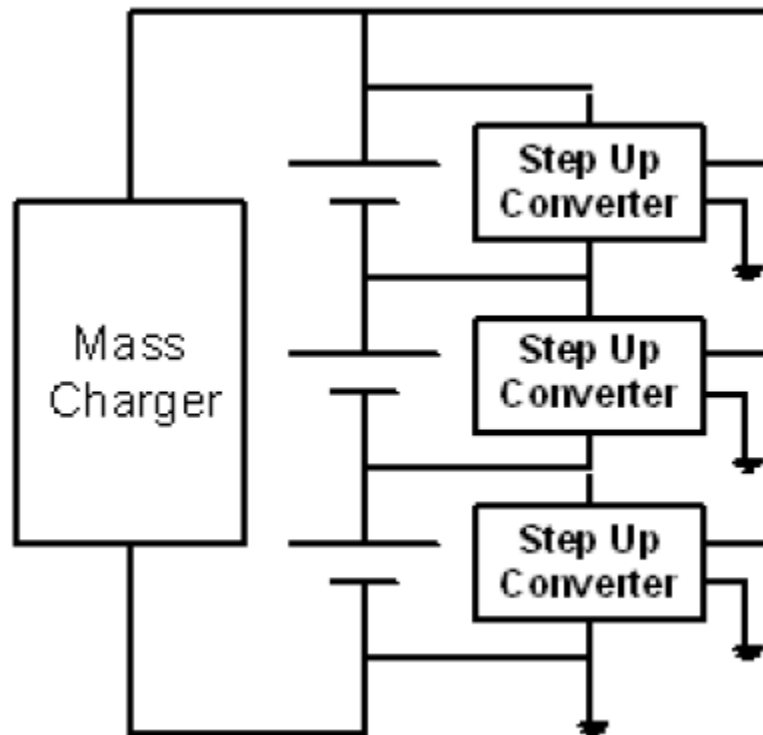


Figure 2.7 Step Up Converter Equalization [12].

There are various other converter topologies that involve the use of transformers. The topologies make use of multi-winding transformers, switched transformers, and multiple transformers in one topology. An example is shown in Figure 2.8.

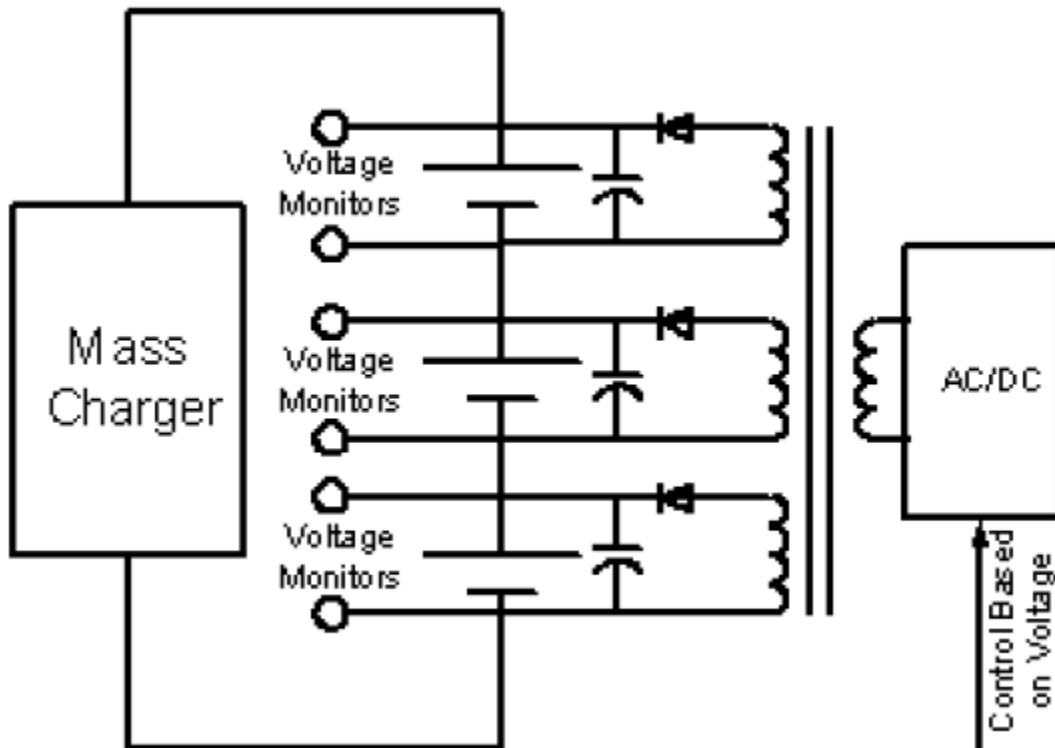


Figure 2.8 Transformer Based Converter Equalizer [12].

The multi-winding transformer of Figure 2.8 must be customized based on the number of cells. This makes it unable to be modularized. Also, the circuit is complex and costly, as are most equalizer designs that involve transformers.

2.1.8 Conclusions

There are several different techniques that are employed to achieve battery equalization. The different topology that is needed depends on application and budget. The effectiveness of each topology varies. As the equalization topology becomes more effective, the cost and complexity of control generally increases. Regardless of cost and complexity,

battery equalizers are needed to improve system performance, increase the lifetime, and increase the safety of battery systems.

2.2 Very High Frequency Power Electronics

2.2.1 Benefits of Increasing Switching Frequency

There is a growing need for power electronics to have greater compactness, manufacturability, and higher performance [6]. This leads to increased research that aims to increase switching frequencies. Increasing switching frequency reduces the energy storage requirements of converters, improves transient performance, and enables the miniaturization of passive components [6-7]. The miniaturization of passive components is especially important in this research. It is thought that a very high frequency equalizer would be able to be integrated within the battery modules themselves. Magnetics typically dominate the size and cost of power electronic systems.

2.2.2 Zero Voltage Switching

Frequency cannot be increased without penalty. As the switching frequency of a converter is increased, it is necessary to consider increased loss mechanisms, including switching loss and gating loss. To counteract the increased loss, zero voltage switching (ZVS) can be used to reduce capacitive discharge loss and voltage/current overlap losses at the switching transitions [8]. Figure 2.9 shows this overlap that results in unnecessary power dissipation. Switching loss is directly proportional to switching frequency.

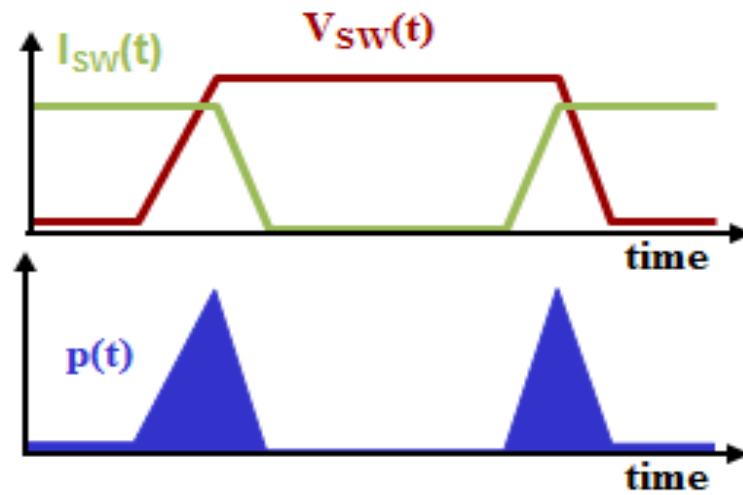


Figure 2.9 Hard Switched Power Dissipation [55].

ZVS can help eliminate the power dissipation shown in Figure 2.9 because the switch transitions occur when the voltage across the switch is zero and thus the power is zero. An example of ZVS is shown in Figure 2.10. Notice how the switching turn on and turn off occurs when the drain-source voltage of the switch is zero. The switching frequency in the simulation is 5 MHz.

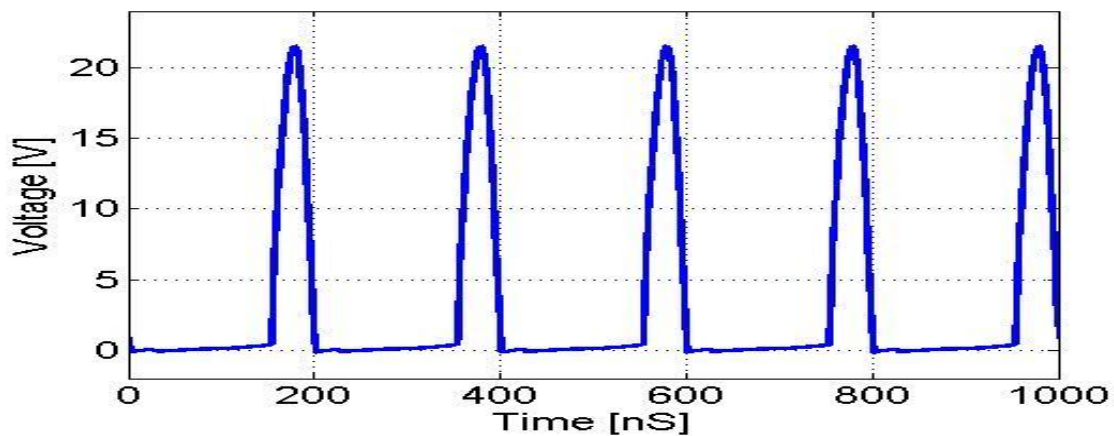


Figure 2.10 Example of ZVS for 5 MHz Design.

2.2.3 Resonant Gating

Resonant gating can also be used to help reduce gating loss. This loss is directly proportional to switching frequency when hard gating is used, as seen in (2.1).

$$P = Q_G * V_{GS} * f_s \quad (2.1)$$

Q_G is the charge on the gate, V_{GS} is the voltage used to drive the gate, and f_s is the switching frequency. It is shown in [8] that soft gating realized by a resonant gate drive strategy results in power dissipation shown in (2.2).

$$P = 2\pi^2 * f_s^2 * C_{iss}^2 * R_G * V_{G,ac}^2 \quad (2.2)$$

R_G is the gate resistance, C_{iss} is the input capacitance, and $V_{G,ac}$ is the magnitude of the sinusoidal voltage swing at the gate. Figure 2.11 shows normalized power loss for hard gating and soft gating. It is clear that resonant gating can offer improved performance.

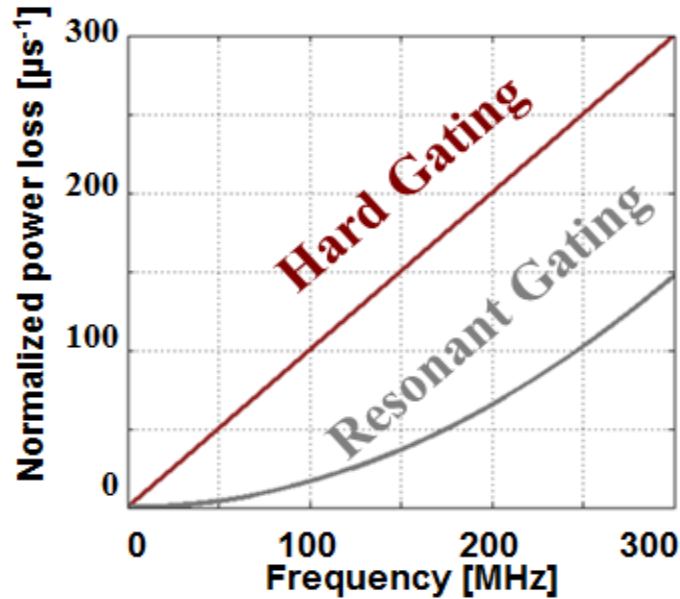


Figure 2.11 Normalized Gating Loss for Different Gating Techniques [55].

The resonant gating loss is proportional to the square of the switching frequency while the hard gating loss is directly proportional to the frequency. Resonant gating can only diminish losses if the gate time constants are short compared to the desired switching time [8]. There are many strategies and topologies for implementing the resonant gate drive in actual applications [8-11].

2.2.4 Very High Frequency Magnetics

Although inductance values scale inversely with frequency, the scaling of the actual size of magnetic components is much more complex. The size depends on winding loss, core loss, permeability, and heat transfer [6]. For reference, a single winding core is shown in Figure 2.12.

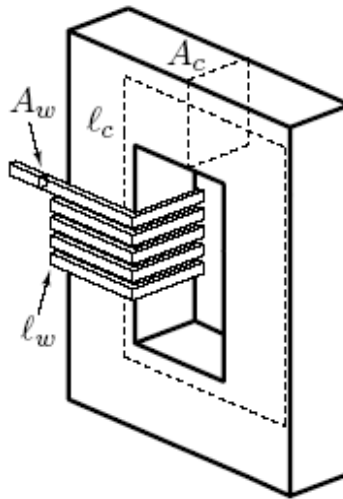


Figure 2.12 Typical Single Winding Magnetic Core [55].

Increased core loss is another reason why frequency cannot be increased without bound. Loss due to the magnetic core is given in (2.3).

$$P_{core} = V_{core} C_M f_s^\alpha B_{ac}^\beta \quad (2.3)$$

V_{core} is the core volume, B_{ac} is the sinusoidal flux density in the core, and C_M , α , and β are parameters dependent on core material. Winding loss, another consideration that needs to be taken into account when increasing switching frequency, is given in (2.4).

$$P_w = \frac{1}{2} I_{ac}^2 R_{ac} \quad (2.4)$$

I_{ac} is the sinusoidal ac current amplitude. R_{ac} is given in (2.5).

$$R_{ac} = \frac{l_w}{w_w} \sqrt{\pi \rho \mu_o f_s} \quad (2.5)$$

R_{ac} takes skin depth, which decreases as frequency is increased, into account. l_w is the length of the winding. w_w is the width of the conductor. (2.3), (2.4), and (2.5) were taken from [6].

It is shown in [6] that cored ac inductors have a frequency limit in terms of achieving miniaturization. Increasing the switching frequency past this point requires that the core size increase in order to maintain a given amount of power loss. Improvements can be made in size if low permeability RF magnetic materials are used [6]. These materials are used when making the inductors that are present in the bidirectional resonant SEPIC converter that is introduced later in this report.

Coreless magnetics can also offer great improvements in magnetic size due to the fact that winding loss is the only major loss mechanism. When temperature rise is not taken into effect, it is shown in [6] that inductor volume scales as $f^{-3/2}$ when the quality factor (Q) is

held constant. When temperature is taken into effect, the inductor volume scales as $f^{-1/2}$.

During this case, it is shown that the quality factor scaling improves as $f^{1/3}$.

Fig 2.13 shows a study performed in [6] that compares volume of a conventional high permeability magnetic material (3F3), an RF material (Ferronics P), and a coreless inductor. The study is for a given maximum loss budget ($Q \geq 100$) and a temperature rise limit ($T \leq 40^\circ$). Figure 2.6 shows that the coreless inductor will always outperform a cored inductor beyond a certain frequency.

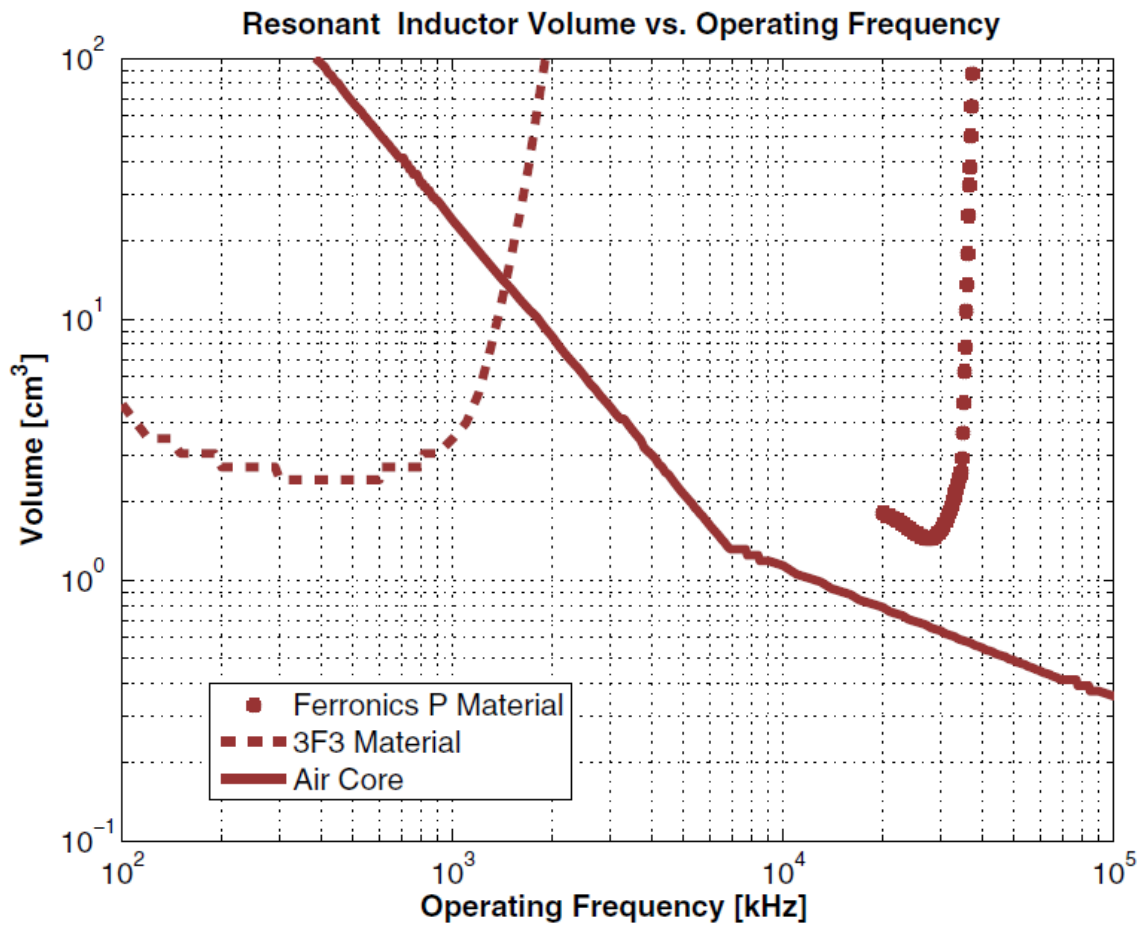


Figure 2.13 Comparison of Inductor Volume for Different Cores [6].

Air cores were also tested when designing the inductors used in the bidirectional resonant SEPIC converter.

2.2.5 Other Considerations for Very High Frequency Design

When switching frequency is increased, semiconductor device parasitics can no longer be ignored. Device loss, such as gating loss (discussed previously) and displacement loss, includes terms that depend on the square of device capacitance and switching frequency. Therefore, VHF power conversion can be improved through semiconductor device improvements [6]. SiC devices and GaN devices offer exciting opportunities in this area. The parasitics of these are reduced because device capacitance is greatly reduced when compared to typical silicon semiconductors [6]. Because of this, device loss can be reduced. These types of devices can be operated at much higher temperatures but generally are still very expensive and hard to obtain.

Other challenges for VHF power converter design are explored in [6]. These challenges include:

- 1) Difficulty of achieving high efficiency over a wide load range
- 2) Difficulty of maintaining good performance across wide input and output voltage ranges
- 3) Existing VHF topologies have high component stresses

There are many challenges associated with VHF power converter design, but many advantages as well.

Chapter 3. Battery Modeling

3.1 Introduction to Battery Modeling

Since batteries are very complex, accurate battery modeling can improve the performance of a battery management system. There are many different models that are employed in hopes of obtaining an accurate representation of the battery. Electrical circuit models are prevalent [16-35], but there are also models that are solely based on battery chemistry [36]. The electrical circuit models will be the focus of this section.

Formulation of model parameters involves performing tests on the battery that is to be modeled. These tests can range from taking impedance spectroscopy measurements [38] to submitting the battery to a pulsed discharge load profile [17]. Tests like these usually take a long time (multiple days) to complete. Also, it is very hard to perform these measurements because accuracy is of paramount importance. Precise voltage and current sensing is required. Optimization software is generally used to generate model parameters based on the measured data [35]. These parameters vary with state of charge, temperature, and state of health, among other things. Since the parameters are nonlinear, the complexity of the models increases as the number of circuit components in the model increases.

In this section, a model is proposed that is easy to generate and very simple. Optimization software is not needed and the tests required to formulate the model do not take long to complete. The performance of the proposed model exceeds that of the four most common electrical circuit models used today.

3.2 Overview of Various Battery Models

The following subsections will examine the most popular battery models.

3.2.1 Rint Model

The Rint model is shown in Figure 3.1. This model is one of the simplest battery models used. The voltage source and series resistance vary based on a number of factors such as state of charge, temperature, and whether the battery is in a charging state or a discharging state [34].

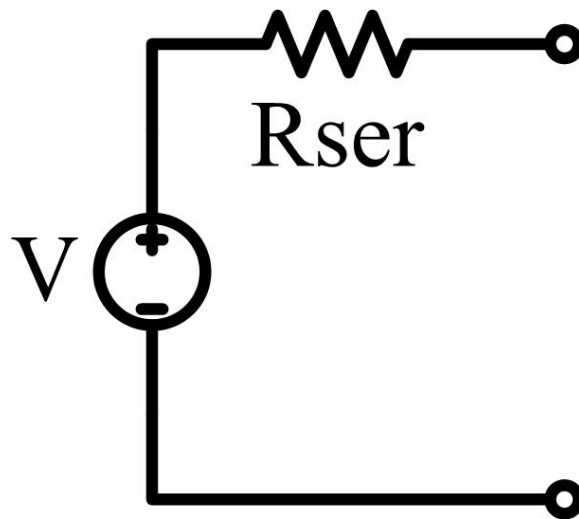


Figure 3.1 Rint Battery Model Circuit Topology.

3.2.2 1 RC Model

The 1 RC model is shown in Figure 3.2. This model includes an RC element that aims to better simulate transient performance [23].

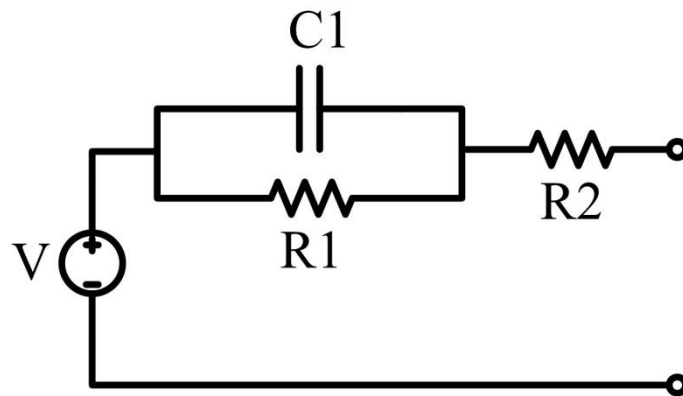


Figure 3.2 1 RC Battery Model Circuit Topology.

3.2.3 2 RC Model

The 2 RC model is shown in Figure 3.3. This model includes 2 RC elements which makes it similar to the 1 RC model but the 2 RC model aims to have better model accuracy at the end of charge and discharge [23].

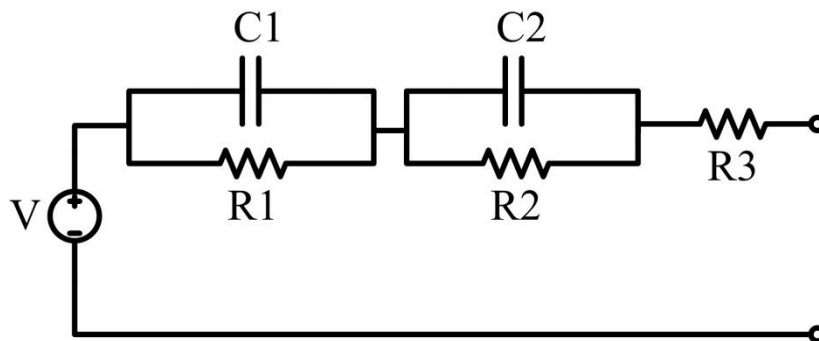


Figure 3.3 2 RC Battery Model Circuit Topology.

3.2.4 SAFT Model

The SAFT model is shown in Figure 3.4. This model is different from the previous three in that the main element is a capacitor, not a voltage source. The bulk capacitor (C_b) is

used to estimate the SOC while the surface capacitor (C_c) is used to simulate the surface effects of the battery [23]. This model can be seen then as a combination of the electrical models and chemistry models. The resistances change with SOC. However, the two capacitors remain generally constant for a given temperature [35].

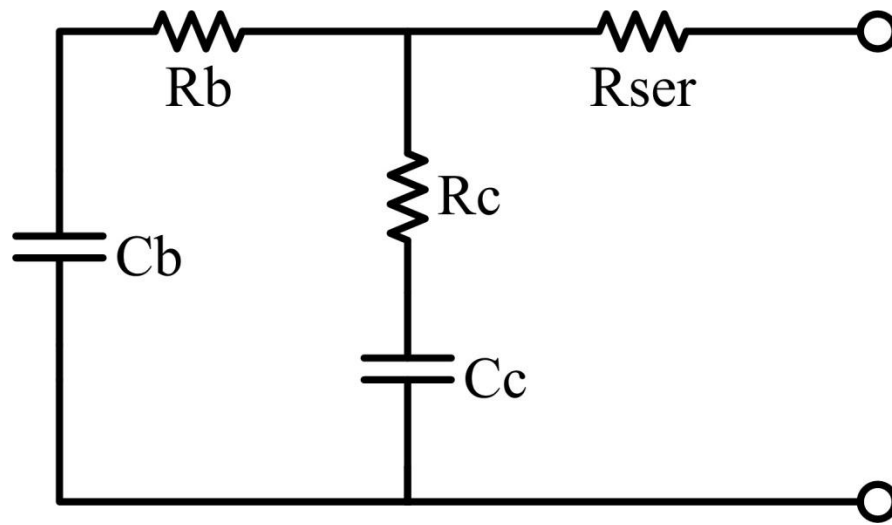


Figure 3.4 SAFT Battery Model Circuit Topology.

3.2.5 UW Model

The proposed UW model is shown in Fig 3.5. It can be seen as a variation of the SAFT model of Fig 3.4. A capacitor is the main element but the component number is reduced to two. The main elements of the SAFT model (the bulk capacitor and the series internal resistance) make up this simplified representation. As mentioned previously, this model has minimal complexity due to the small number of components. However, the bulk capacitor of this model is not constant for a given temperature as it is in the SAFT model shown previously.

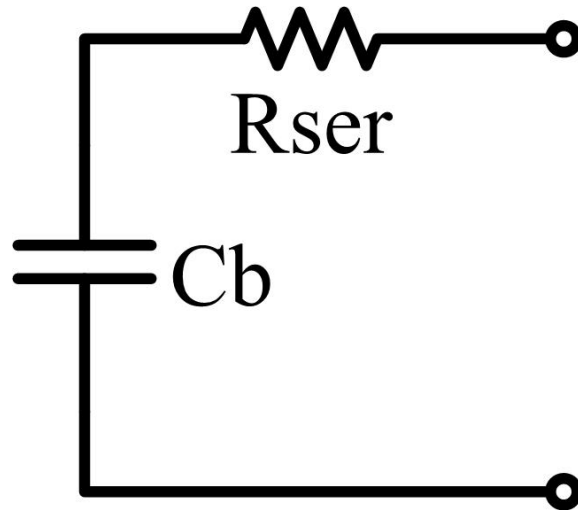


Figure 3.5 Proposed UW Battery Model Circuit Topology.

3.3 Battery Measurements

3.3.1 A123 Cylindrical Cell

An A123 (now part of Johnson Control, Inc.) ANR26650 Cylindrical Cell was chosen as the battery of interest. This Lithium-Ion cell was chosen because the primary applications of this cell include grid stabilization energy storage systems and commercial truck and bus hybrid electric vehicles [37]. The nominal voltage of a cell is 3.3 V and the measured capacity is 2.3 Ahr. The dimensions of a single cell are shown in Figure 3.6.

3.3.2 Battery Test Setup

As mentioned previously, it is important that all measurements are very accurate. To perform all desired measurements, the test stand described in [38] is used. Figure 3.7 shows the test stand setup. The test stand can discharge or charge the battery according to any arbitrary test profile that the user desires. It is also capable of performing impedance

spectroscopy measurements. The measurement results are conveniently exported to Microsoft Excel.

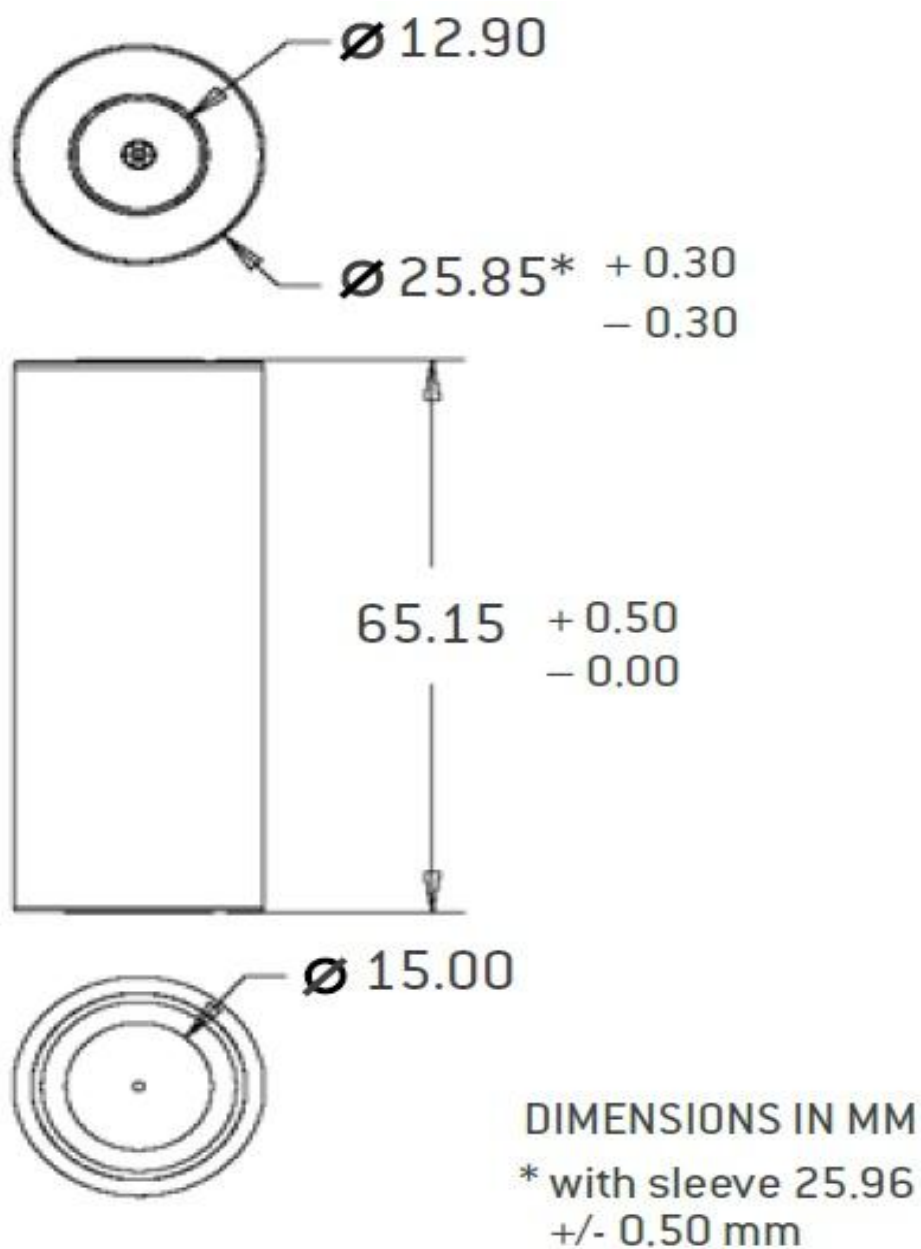


Figure 3.6 A123 Cell Dimensions [37].

Different types of measurements were performed in order to be able to formulate all five models.

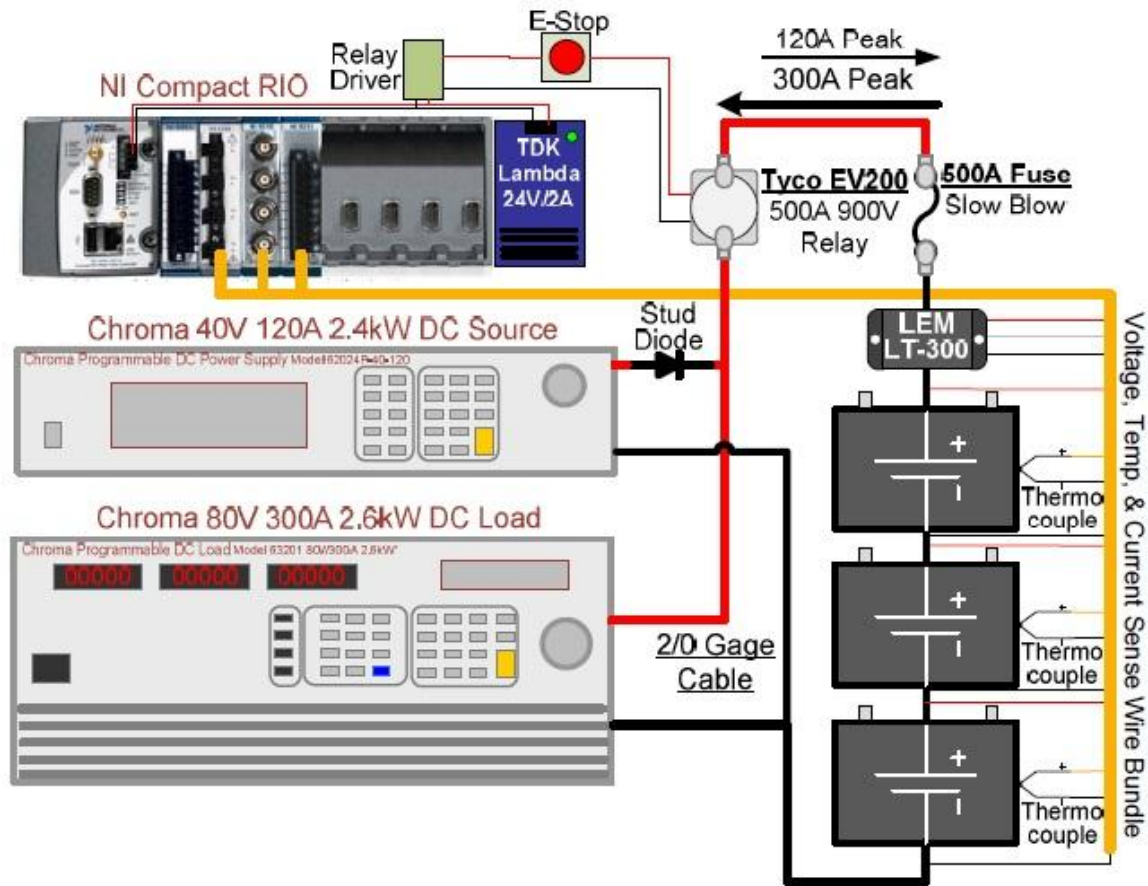


Figure 3.7 Test Stand Setup [38].

3.3.3 Constant Current Discharge

Constant current discharge measurements, shown in Figure 3.8, were taken at a number of discharge rates in order to determine the series resistance. Three of the models have a series resistance (R_{ser}) so this test was used for those models. In order to determine the series resistance, the approach used in [16] was employed, except that a resistance value was calculated for more than one SOC value. The series resistance is equal to the difference in voltages of the two discharge characteristics in Figure 3.8 divided by the difference in discharge current.

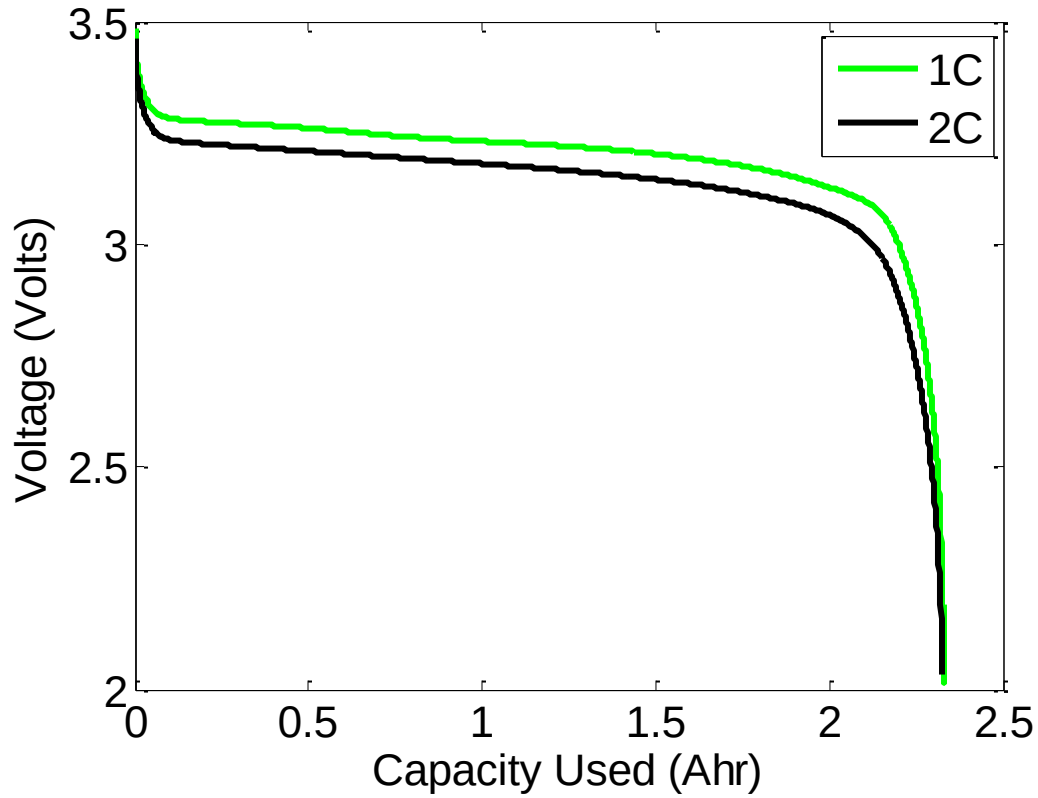


Figure 3.8 Constant Current Discharge Measurements for A123 Cell.

The 1C and 2C discharge characteristics were used for this analysis. The series resistance is defined by (3.1).

$$R_{ser} = \frac{V_{1c} - V_{2c}}{i_{1c} - i_{2c}} \quad (3.1)$$

The denominator of (3.1) is a constant value (2.3 A) since both discharge currents are constant.

The series resistance calculations are shown in Table 3.1.

Table 3.1 Series Resistance for Different SOC Values.

SOC (%)	R _{int} (mΩ)
0	46.1
10	31.7

20	25.6
30	25.2
40	24.3
50	23
60	21.3
70	21.3
80	22.6
90	22.2
100	9.3

3.3.4 Open Circuit Voltage Measurements

Open circuit voltage (OCV) measurements were taken at the 11 different SOC values shown in Table 1. In order to get these measurements, the battery was discharged at a rate of 1C and a voltage reading was taken after an hour to allow the battery to reach equilibrium [32]. The OCV measurements were used to describe the voltage source behavior in all of the models that have a voltage source as the main component. The OCV measurements were also used in the UW model and the SAFT model. Figure 3.9 shows the measurement results.

3.3.5 Impedance Spectroscopy

Impedance spectroscopy is a very common electrochemical technique that is used to obtain impedance data in the frequency domain. A voltage or current source is needed to inject a sinusoidal waveform into the battery. The current or voltage response of the battery is measured. This data is then used to determine the battery impedance at each frequency.

Impedance Spectroscopy (IS) measurements were performed on the battery at the 11 different SOC values of Table 1 for 19 different frequencies ranging from 0.02 Hz to 5 kHz. As a result, this testing took about one week to complete.

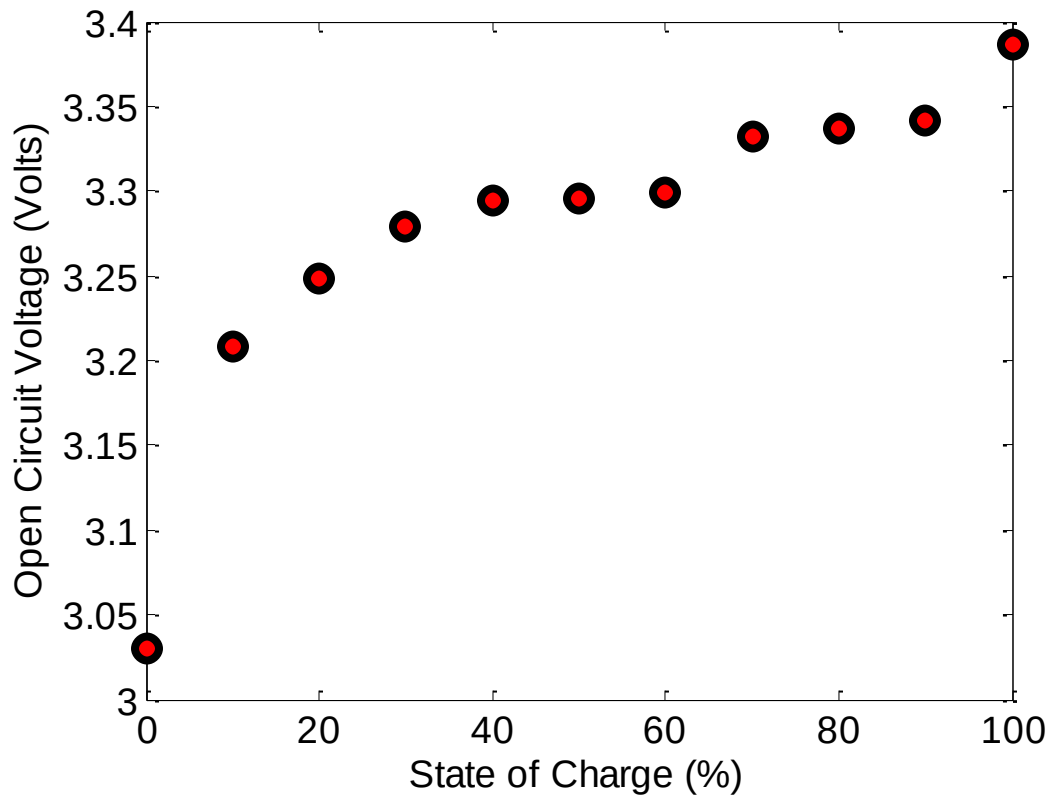


Figure 3.9 Open Circuit Voltage Measurements for A123 Cell.

The IS data was imported into a least-square fitting program named Kumho MESIP, which is available for download at [39]. This software was used to estimate the R and C parameters for the 1 RC and 2 RC models. It is thought that using IS as a method to determine the R and C parameters instead of a pulsed discharge method, such as the one used in [17], is preferable since IS is time independent and not specific to a defined load profile.

An example of the results gained from the MESIP software is shown in Figure 3.10. The graph depicts data that is valid when the battery is at 80% SOC. The measured data is directly from the test station. The 1 RC model (resistance and one RC element) and 2 RC (resistance and 2 RC elements) Nyquist plots in Figure 3.10 are generated by using the R and

C parameters that are given by the MESIP software for a given SOC. Table 3.2 and Table 3.3 show the output of the least-square fitting program for a given SOC for the 1 RC model and the 2 RC model, respectively. To summarize, using the R and C parameter values at 80% SOC shown in Table 3.2 and Table 3.3, the Nyquist plots for the 1 RC and 2 RC models of Figure 3.10 are generated. Figure 3.11 – Figure 3.20 show the Nyquist plots for the other conditions.

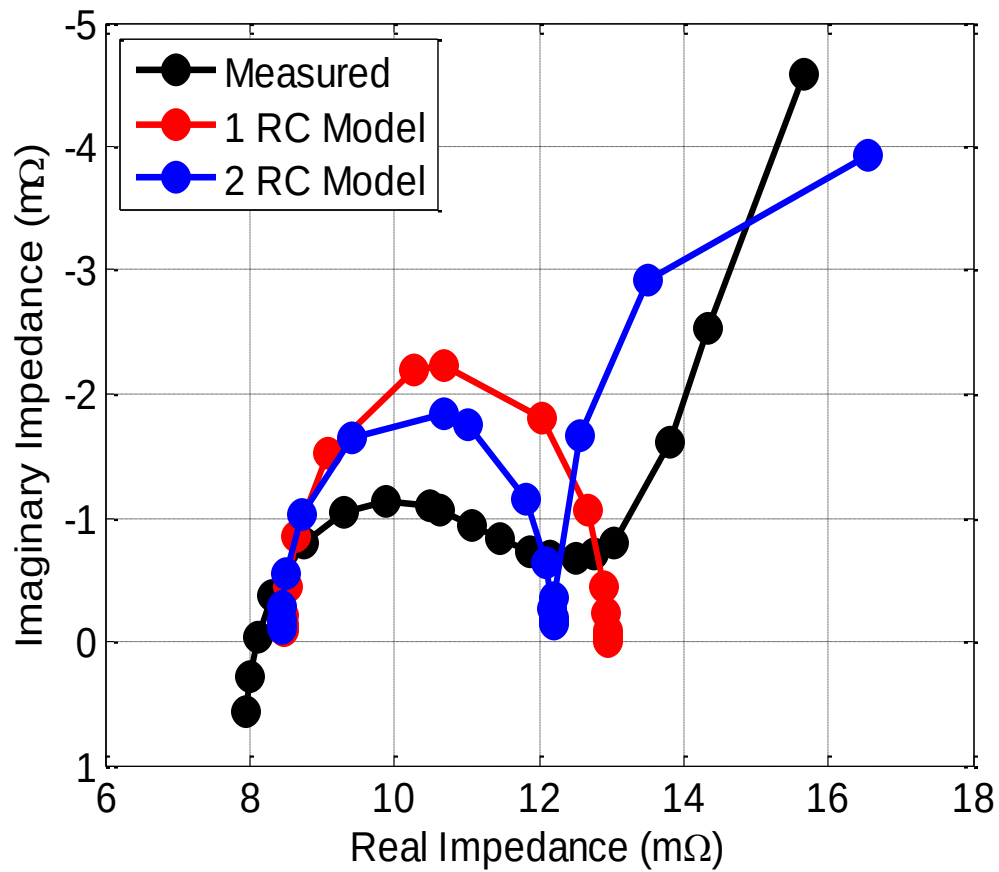


Figure 3.10 Measured and Model Nyquist Plots at 80% SOC.

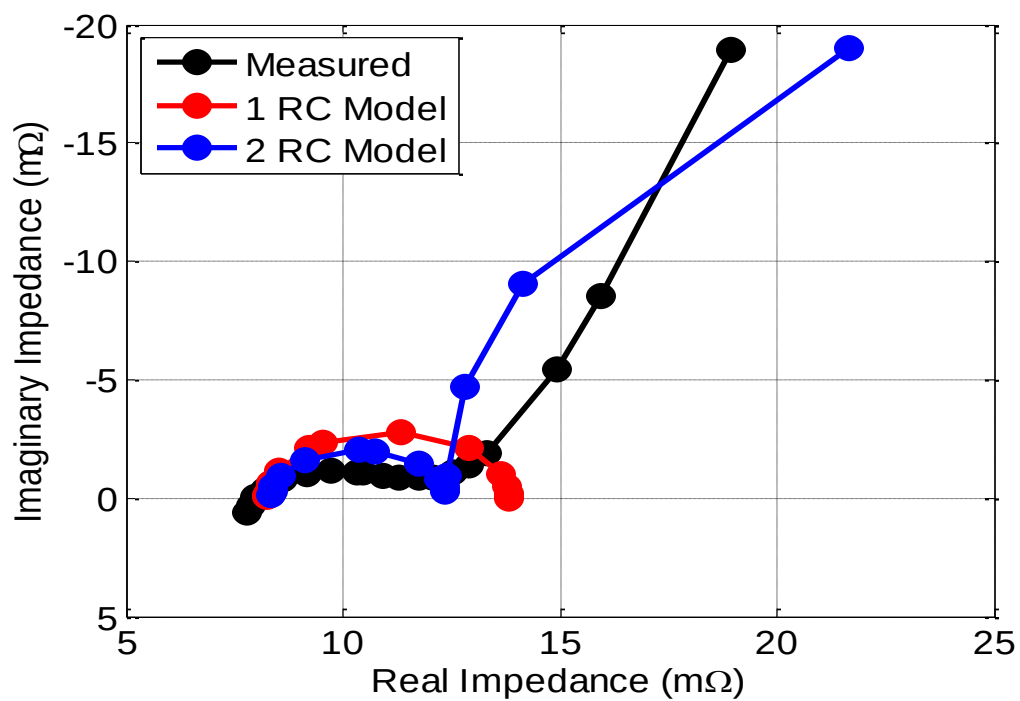


Figure 3.11 Measured and Model Nyquist Plots at 100% SOC.

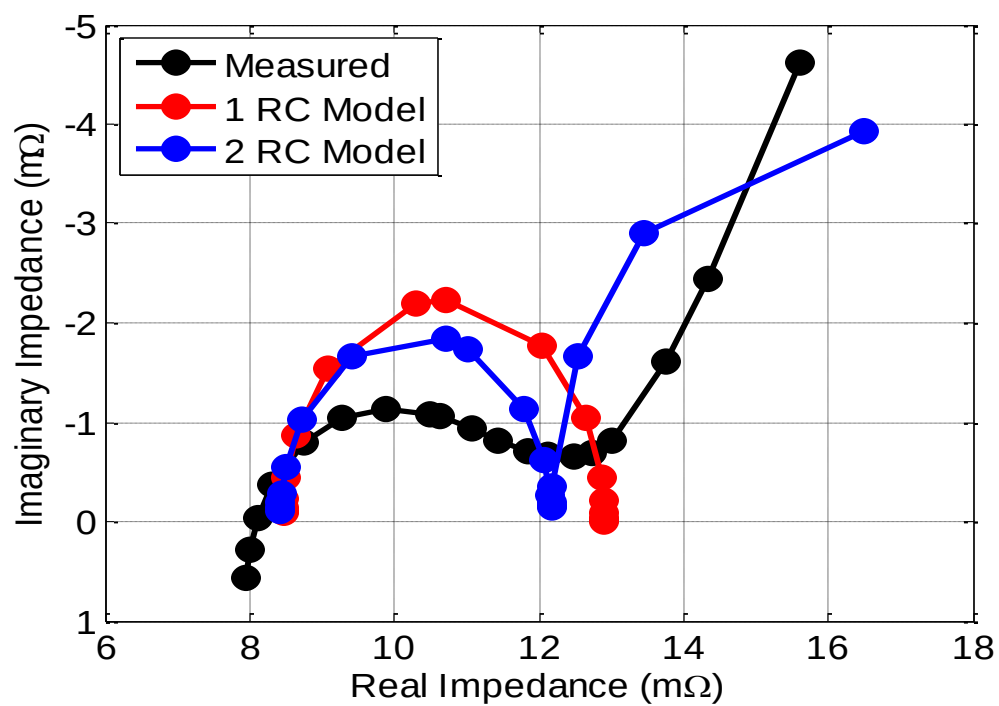


Figure 3.12 Measured and Model Nyquist Plots at 90% SOC.

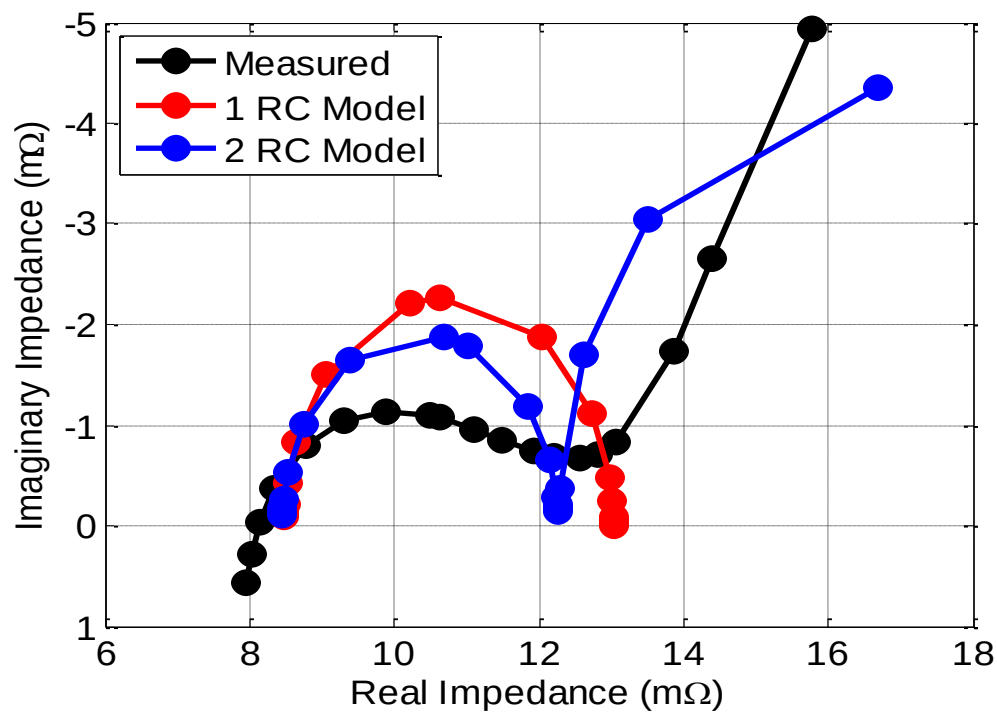


Figure 3.13 Measured and Model Nyquist Plots at 70% SOC.

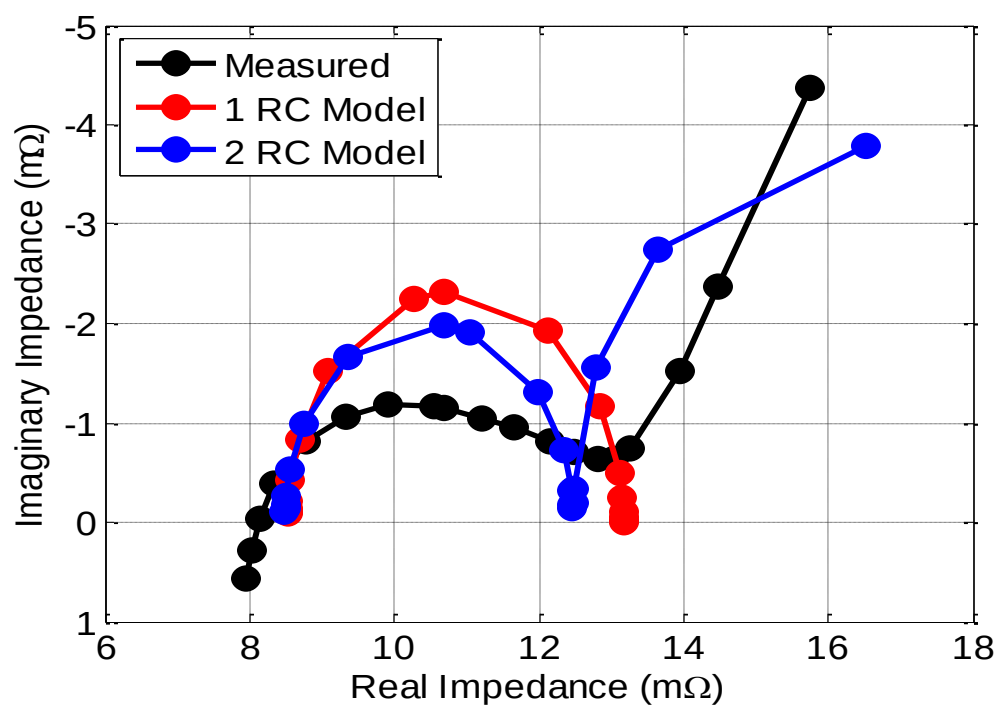


Figure 3.14 Measured and Model Nyquist Plots at 60% SOC.

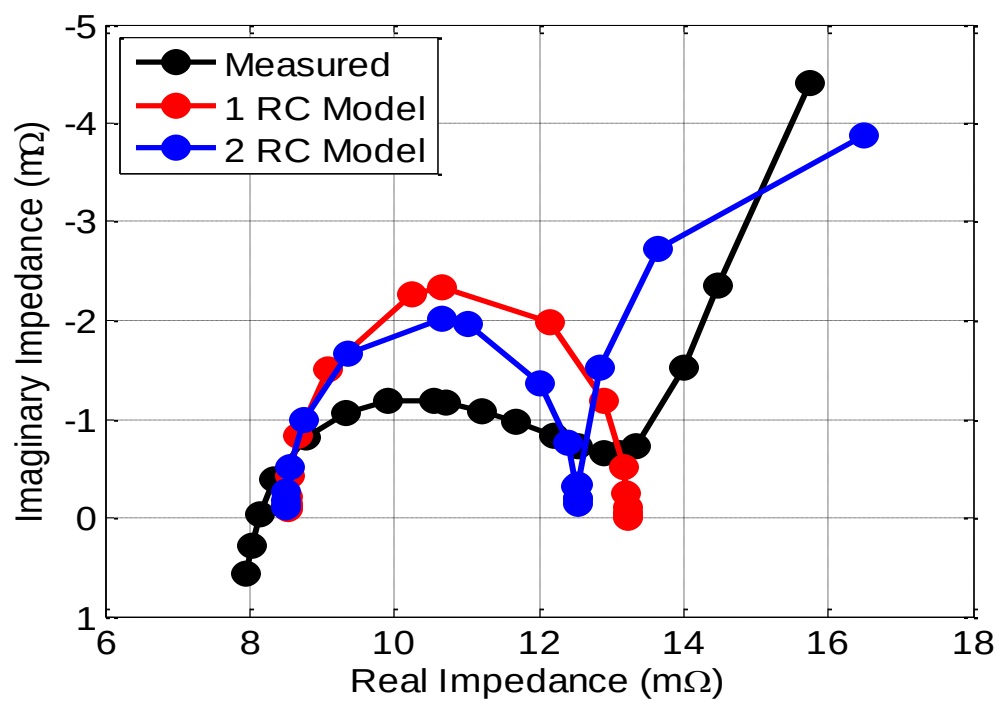


Figure 3.15 Measured and Model Nyquist Plots at 50% SOC.

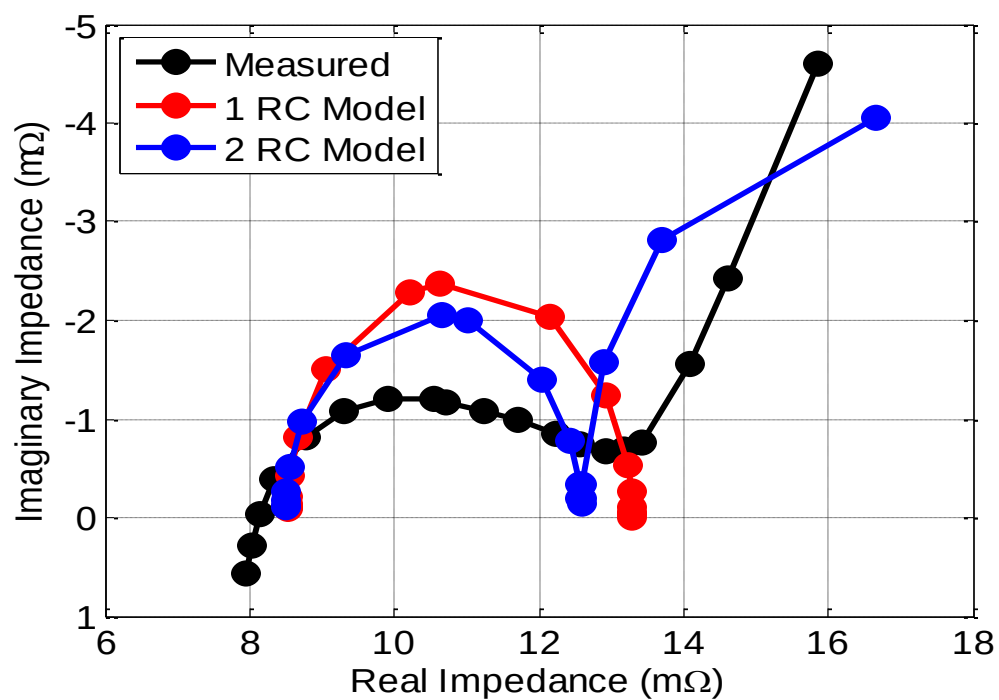


Figure 3.16 Measured and Model Nyquist Plots at 40% SOC.

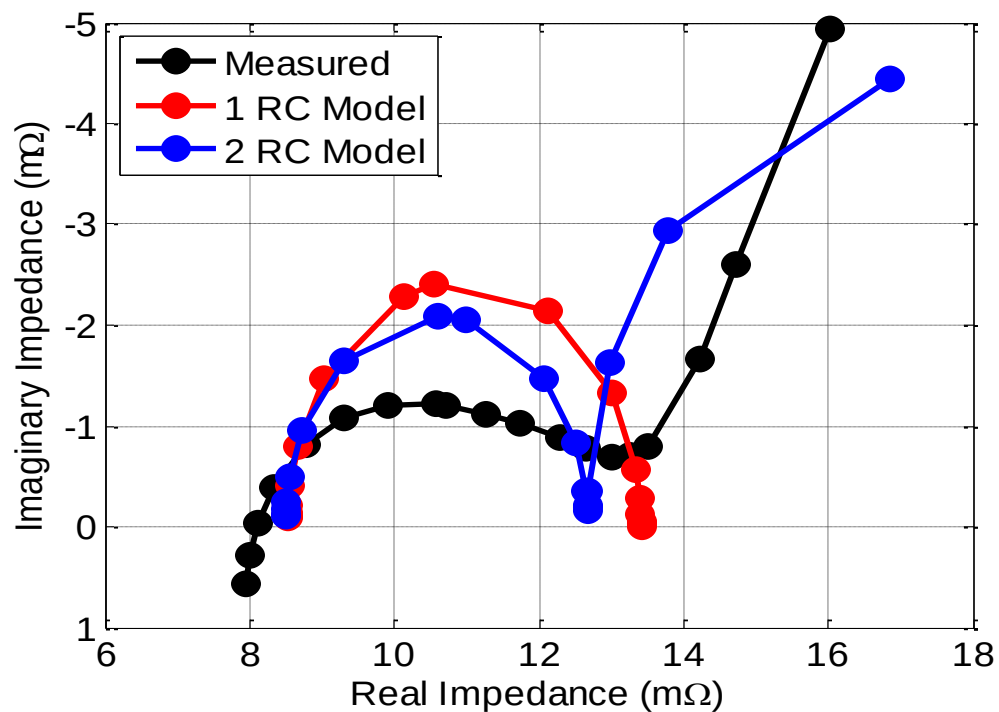


Figure 3.17 Measured and Model Nyquist Plots at 30% SOC.

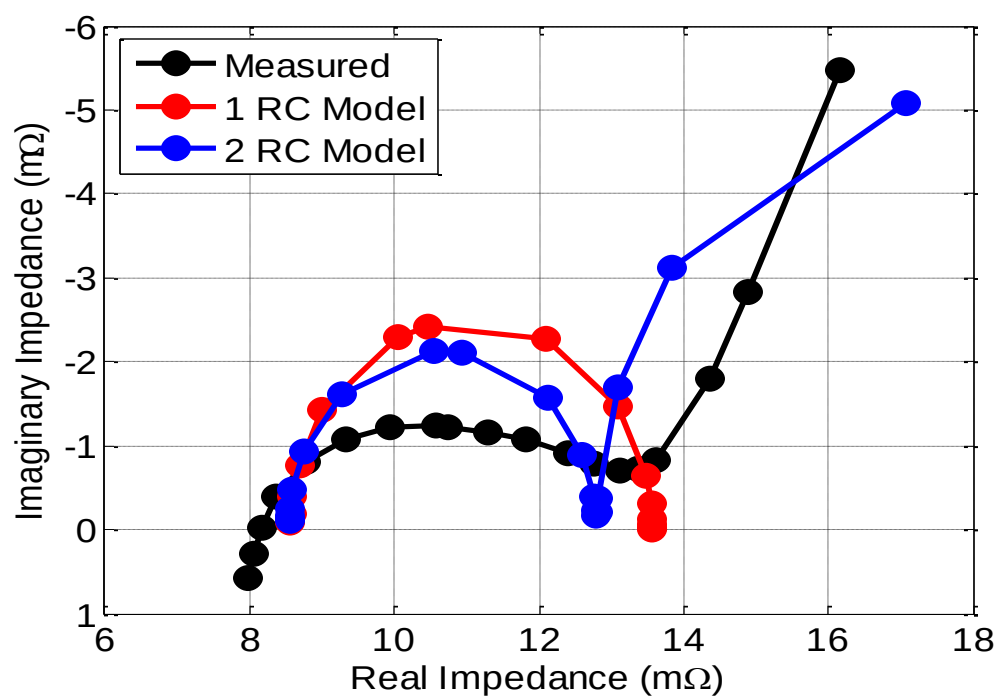


Figure 3.18 Measured and Model Nyquist Plots at 20% SOC.

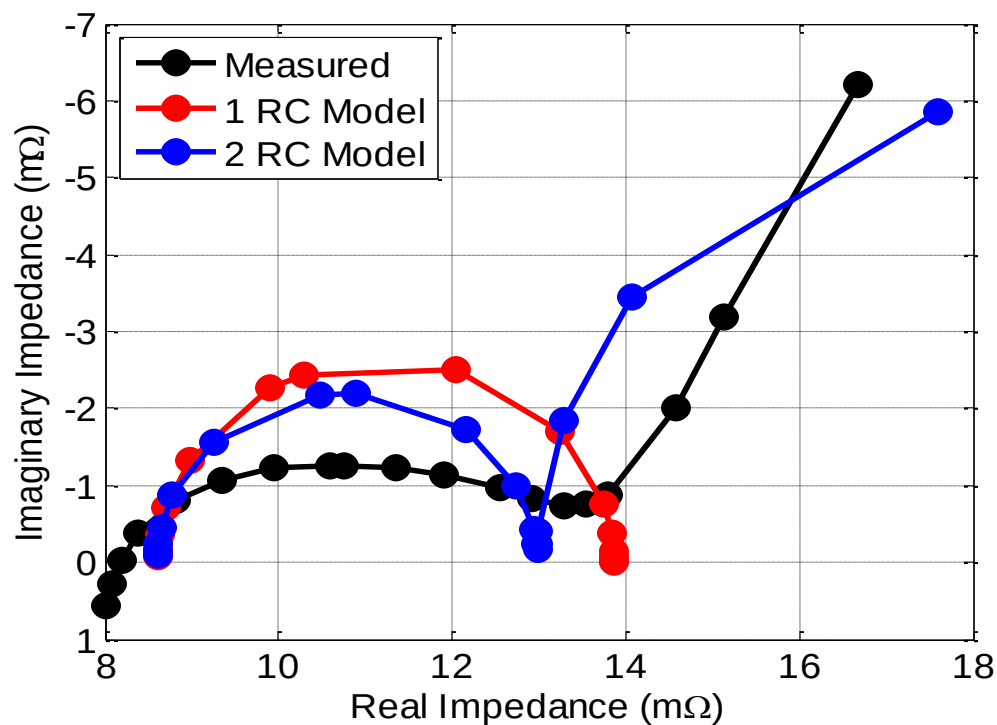


Figure 3.19 Measured and Model Nyquist Plots at 10% SOC.

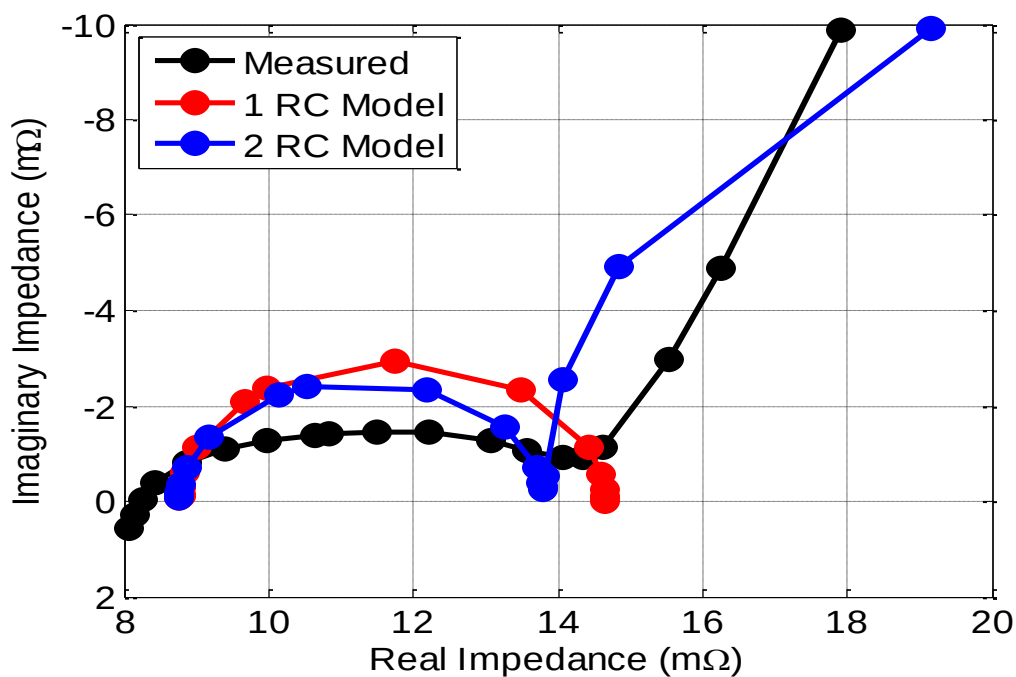


Figure 3.20 Measured and Model Nyquist Plots at 0% SOC.

Table 3.2 1 RC Model Element Values Output by MESIP Software.

SOC (%)	R1 (mΩ)	C1 (F)	R2 (mΩ)
0	5.87	0.535	8.77
10	5.25	0.443	8.62
20	5.01	0.409	8.57
30	4.88	0.388	8.53
40	4.77	0.377	8.53
50	4.69	0.371	8.53
60	4.64	0.368	8.53
70	4.54	0.370	8.48
80	4.47	0.361	8.47
90	4.45	0.354	8.46
100	5.57	0.519	8.26

Table 3.3 2 RC Model Element Values Output by MESIP Software.

SOC (%)	R1 (mΩ)	C1 (F)	R2 (mΩ)	C2 (F)	R3 (mΩ)
0	5.03	0.431	23.7	620	8.77
10	4.39	0.348	12.0	839	8.60
20	4.26	0.328	10.3	917	8.54
30	4.18	0.313	8.91	947	8.50
40	4.10	0.306	8.10	976	8.49
50	4.05	0.303	7.75	1003	8.49
60	3.98	0.298	7.60	980	8.48
70	3.81	0.293	8.71	897	8.45
80	3.77	0.287	7.88	909	8.43
90	3.75	0.283	7.90	919	8.42
100	4.01	0.330	48.3	337	8.35

3.4 MODEL FORMULATION

3.4.1 Rint Model Formulation

The parameters of the Rint model of Figure 1 were generated using the OCV measurements for the voltage source and the CC discharge measurements for the series resistance.

3.4.2 1 RC Model Formulation

The parameters of the 1 RC Model of Figure 2 were generated using the OCV measurements for the voltage source and the IS measurements for the RC element and resistance.

3.4.3 2 RC Model Formulation

The parameters of the 2 RC Model of Figure 3 were generated using the OCV measurements for the voltage source and the IS measurements for both RC elements and resistance.

3.4.4 SAFT Model Formulation

The parameters of the SAFT Model were generated using the OCV measurements for the bulk capacitor and the CC discharge measurements for the series resistance. The model shown in Figure 3.4 was simplified to the same topology shown for the UW model in Figure 3.5. This was done because the bulk capacitor is generally much larger than the surface capacitor [34-35]. As mentioned previously, the bulk capacitor is constant. In order to determine this constant value, the OCV measurements were used. The capacitor value is defined in (3.2).

$$C_{SAFT} = \frac{dQ}{dV} \quad (3.2)$$

Therefore, the inverse of the slope of a best fit line to the OCV data in Figure 3.9 would yield the capacitor value. The OCV data along with the best fit line is shown in Figure 3.21. The 0% and 100% SOC data points were ignored since they represent the two extreme conditions of the battery (completely empty and completely full, respectively). It is important

to note that special attention must be paid to the units. The graph has units of percentage for charge and this must be changed to A*s in order to get an accurate capacitor value.

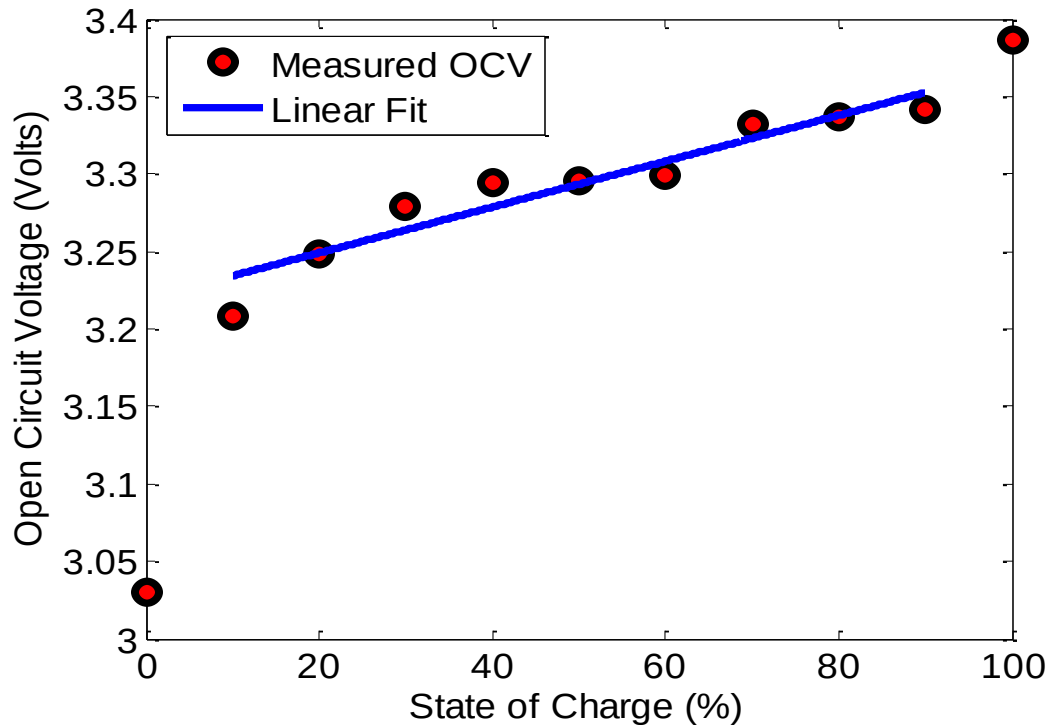


Figure 3.21 OCV Data with Best Fit Line Used for the SAFT Model Formulation.

The capacitor value that results is 55.3 kF.

3.4.5 UW Model Formulation

The parameters of the proposed UW Model were generated using the OCV measurements for the bulk capacitor and the CC discharge measurements for the series resistance. The difference between this model and the SAFT model is that the bulk capacitor is not constant for the UW model. The equation given in (3.2) is still used except that for this model, the inverse slope of the line between two adjacent points is used to determine the capacitor value during that range of SOC. Since there are 11 data points for the OCV

measurement, there will be 10 capacitor values. Figure 3.22 shows the OCV measurements with lines connecting adjacent points. The inverse of the slope of each line yields the capacitance value during each SOC range.

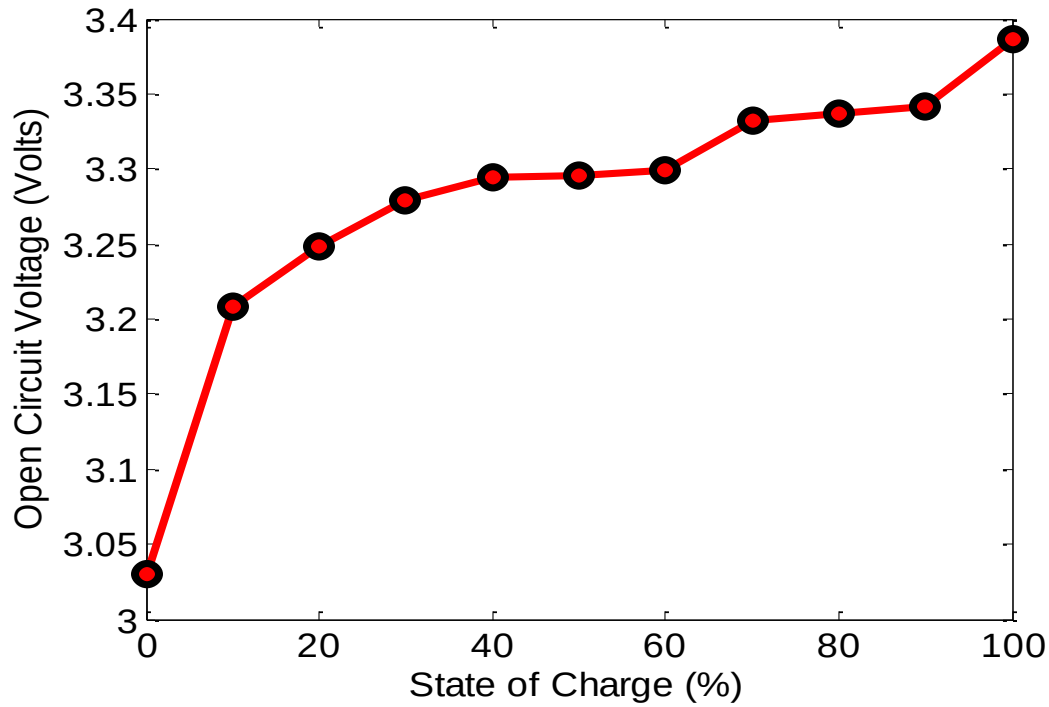


Figure 3.22 OCV Data with Lines that were Used for the UW Model Formulation.

The capacitance values that result are given in Table 3.4.

Table 3.4 Bulk Capacitor Values for the UW Model.

SOC Range (%)	C_B (kF)
0-10	17.250
10-20	258.75
20-30	85.361
30-40	29.361
40-50	267.1
50-60	552.0
60-70	40.588
70-80	27.692

80-90	22.198
90-100	4.913

It is important to note that during simulation of the UW model, the capacitance value was held constant until there was 10% left in the range. At this time, the capacitance then increased linearly to the next value. This was done so that there wasn't a sudden increase in capacitance. It was thought that this would help maintain the integrity of the simulation.

3.4.6 Model Formulation Conclusions

Table 3.5 shows a summary of what measurements were needed to formulate each model. It can be estimated that OCV measurements took about 15 hours to complete, the CC measurements took about 5 hours to complete and the IS measurements took about one week to complete. Therefore, the 1 RC and 2 RC model formulations took much longer to formulate the model for a given battery. Optimization software was also needed for the 1 RC and 2 RC models, increasing the complexity of the formulation task.

Table 3.5 Measurements Needed to Formulate Each Model.

Model	OCV	CC	IS
Rint	X	X	
1 RC	X		X
2 RC	X		X
SAFT	X	X	
UW	X	X	

3.5 Experimental and Model Prediction Results

In order to quantify the effectiveness of the models, each model was submitted to a load profile that is shown in Figure 3.23. The amplitude of the current is the 1C rate (2.3 A).

It was desired to have the battery completely discharge in 6 pulses, as in [17], so the duration of each pulse was set to 600 seconds. The battery was allowed to rest for one hour to reach equilibrium before the next pulse. PLECS Standalone was used to carry out the simulations of all the models. Standard lookup tables were used to vary the parameter values as a function of SOC. The temperature was assumed constant. The simulation results were then compared to actual measured results. While under test, the battery temperature was held at room temperature using the chamber that is attached to the test stand that is detailed in [38]. The PLECS setup for each of the five models is shown in Figure 3.24 – Figure 3.28. The waveform in Figure 3.23 is generated by the function generator and controllable current source. A current sensor and integrator is used to keep a running tally of the battery SOC. The lookup tables use the SOC values according to their definition. Note that for the series resistance, PLECS requires that the variable resistance be connected in parallel with a capacitor. This is because the PLECS solver must be described by Ordinary Differential Equations so the resistor is decoupled by a state variable (a capacitor in this case) to avoid an algebraic loop. For simulation purposes, the value of this capacitor was chosen to be on the order of pico Farads so it has minimal effect on the overall system performance.

3.6 Battery Modeling Simulation Results Conclusions

In order to quantify the effectiveness of each model, error calculations were performed according to (3.3).

$$error = \frac{V_{measured} - V_{model}}{V_{measured}} \quad (3.3)$$

The average error over the entire load profile is shown in Table 3.6.

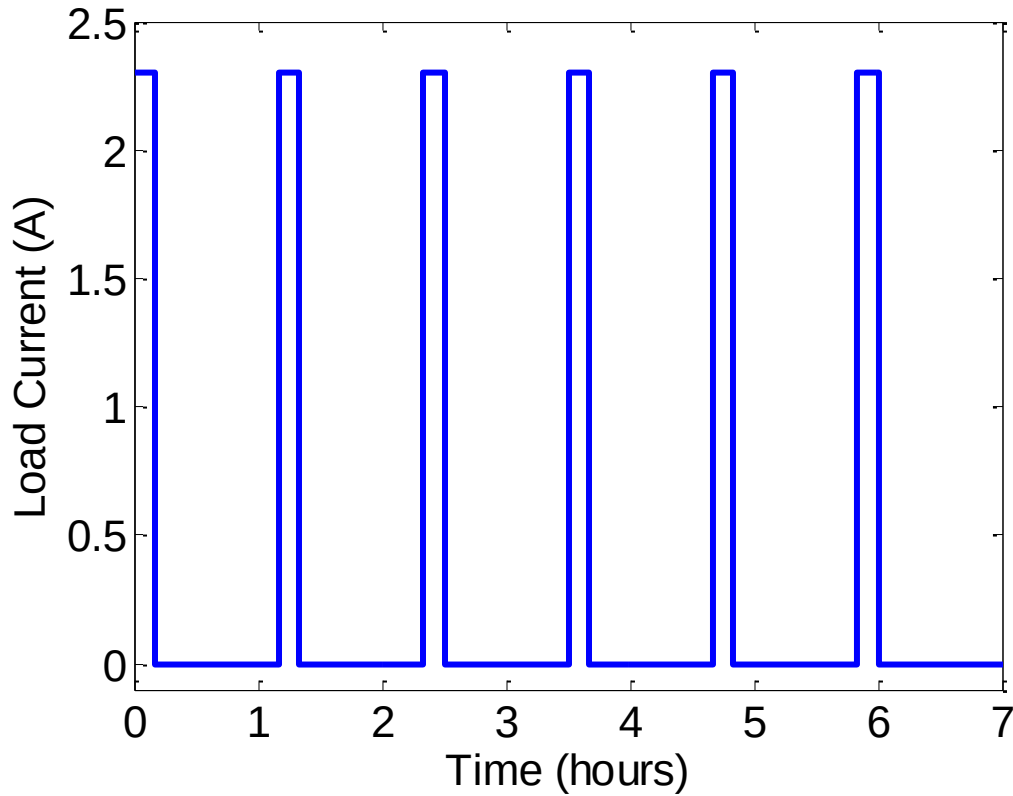


Figure 3.23 Load Current Profile Used to Verify Performance of Models.

Based on these results, the UW model performs the best. The average error values of the Rint and SAFT models are consistent with those reported in [34]. It is also evident from Figure 14 that the SAFT model does a poor job at predicting the steady state voltage value of the battery that occurs when the battery is resting for an hour after each pulse. The voltage source based models have an inherent advantage over the capacitor based models in this regard because the steady state value of the voltage based models is controlled directly by the lookup table of OCV values (when there is no current, the output voltage of the voltage based models is simply the value of the voltage source). The UW model performance, however, is right on line with that of the voltage based models.

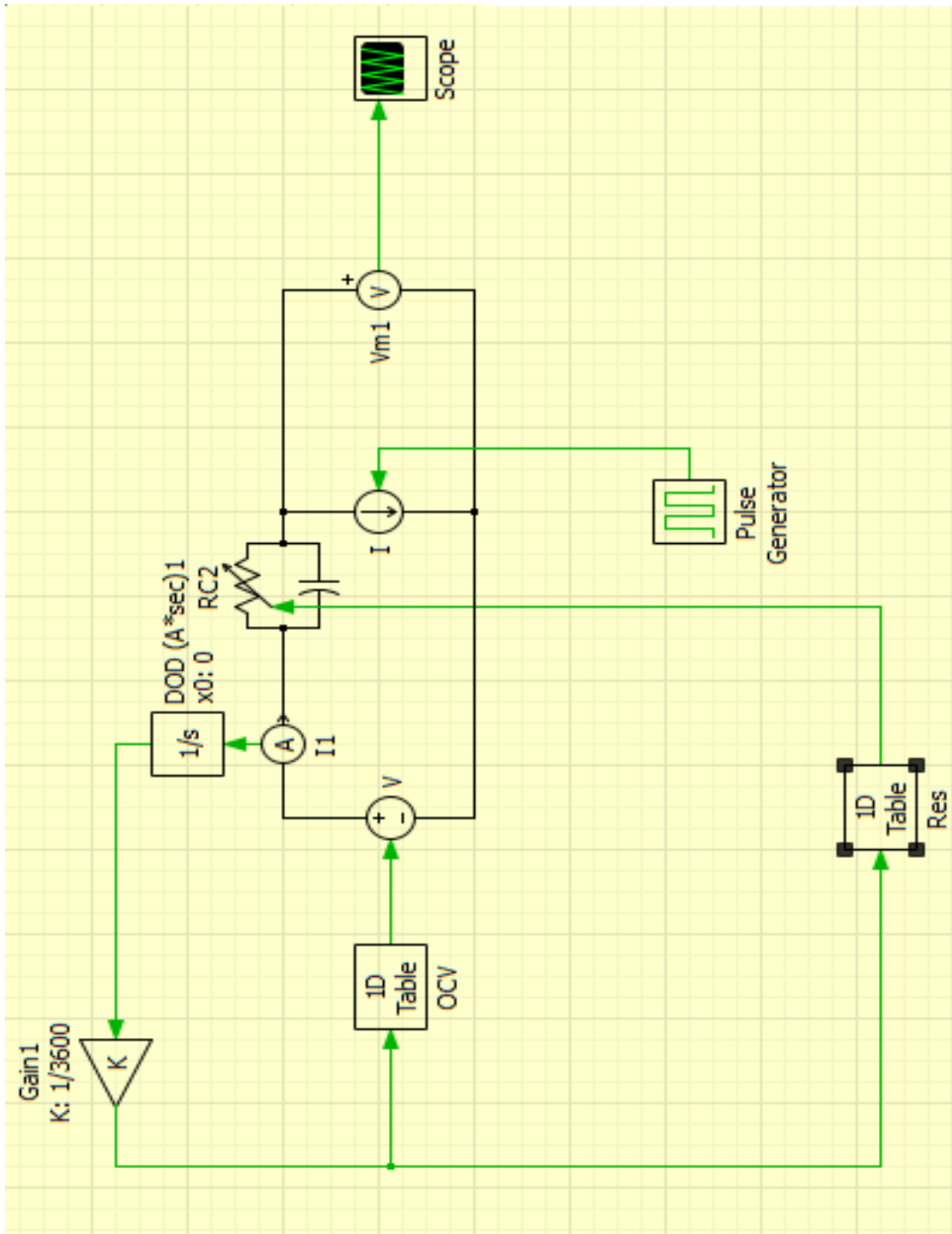


Figure 3.24 Rint Model PLECS Setup.

Figure 3.25 1 RC Model PLECS Setup

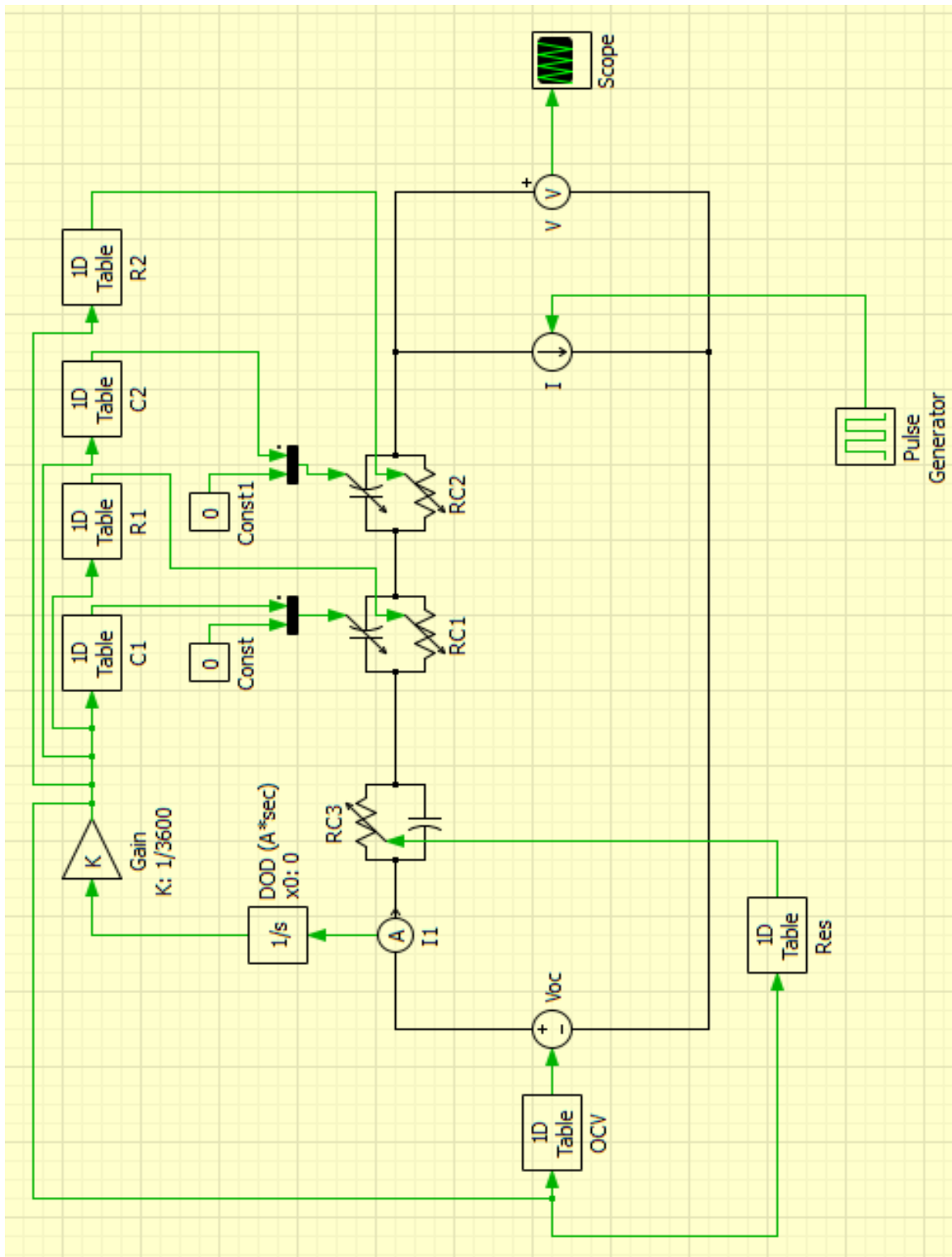


Figure 3.26 2 RC Model PLECS Setup.

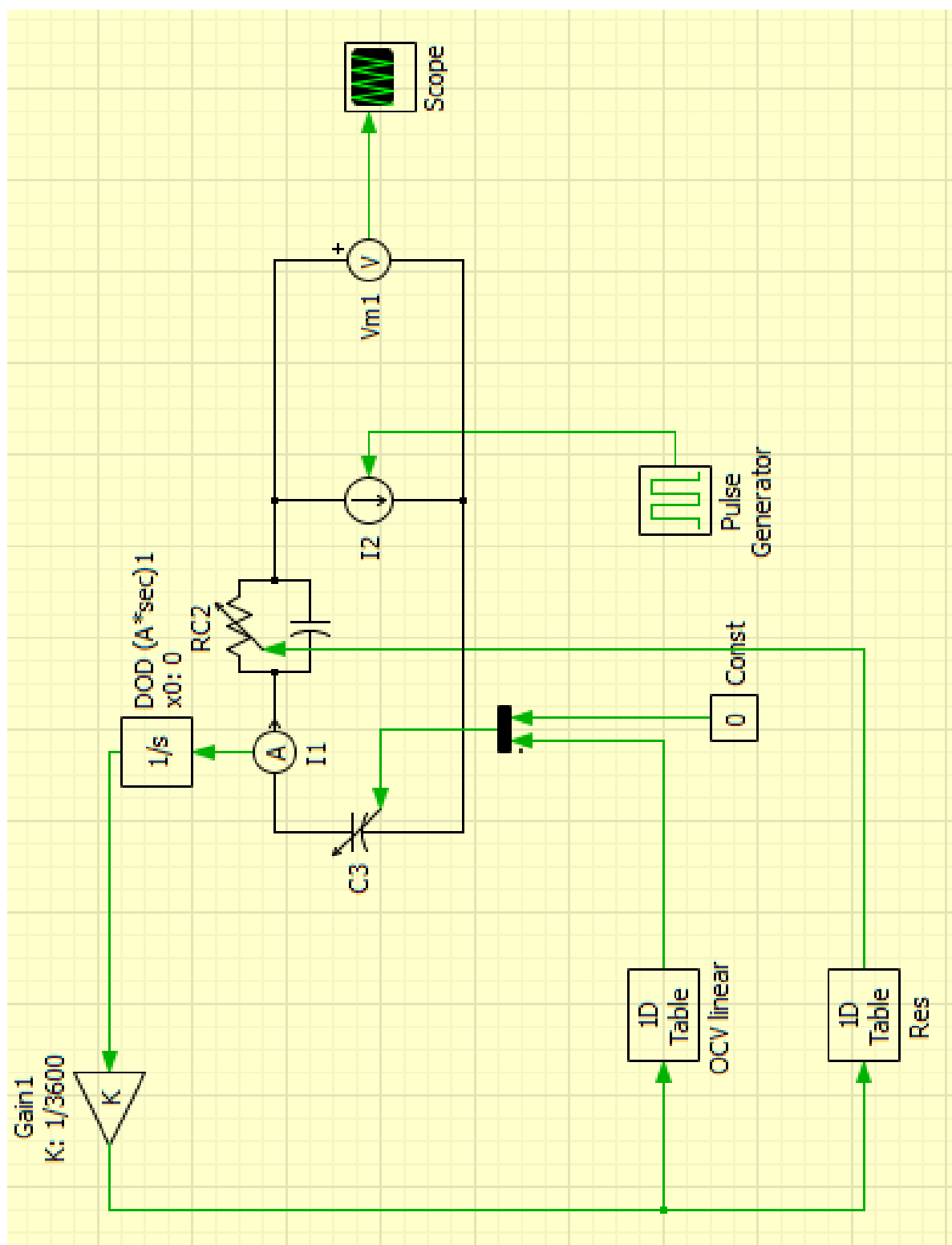


Figure 3.27 SAFT Model PLECS Setup.

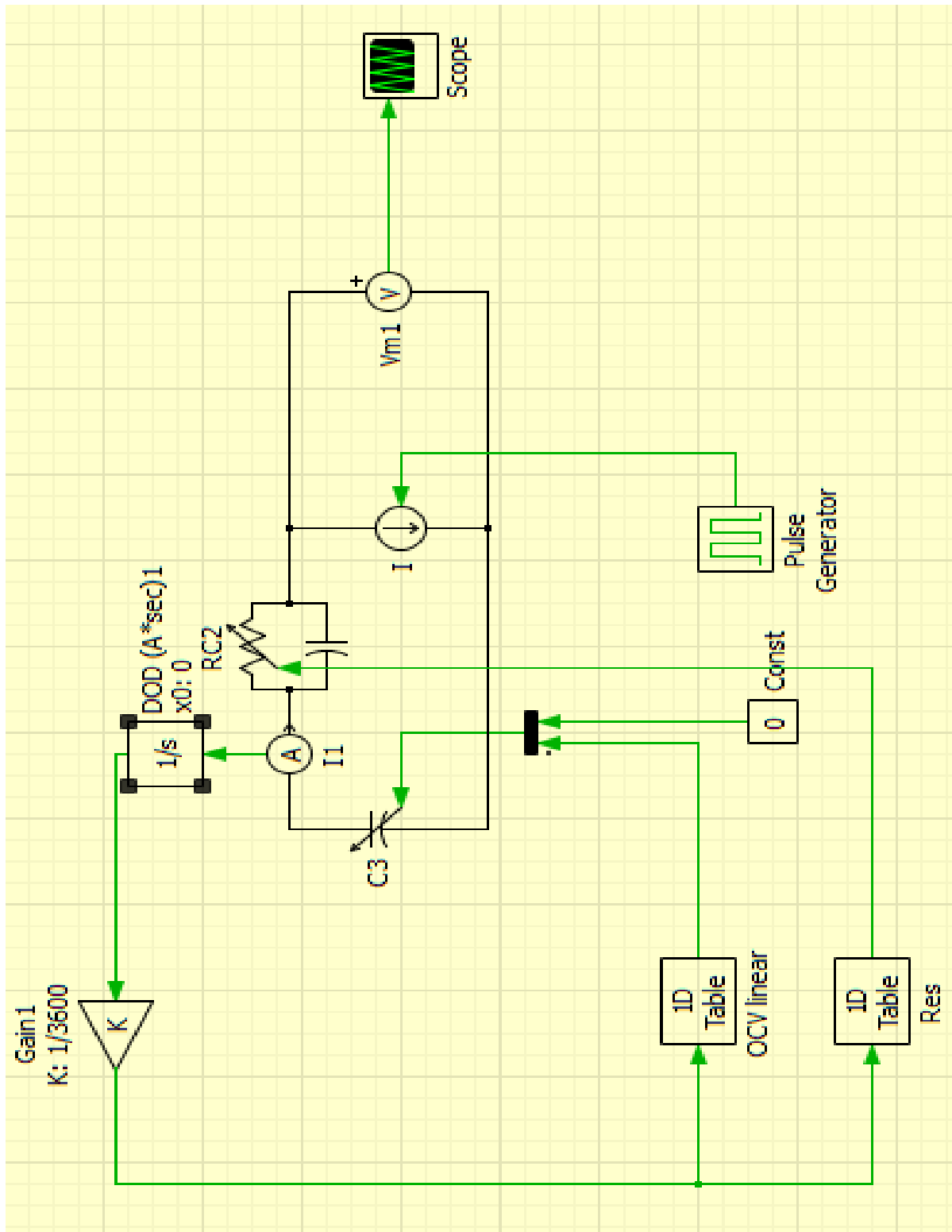


Figure 3.28 UW Model PLECS Setup.

The simulation results along with the measured results are shown in Figure 3.29.

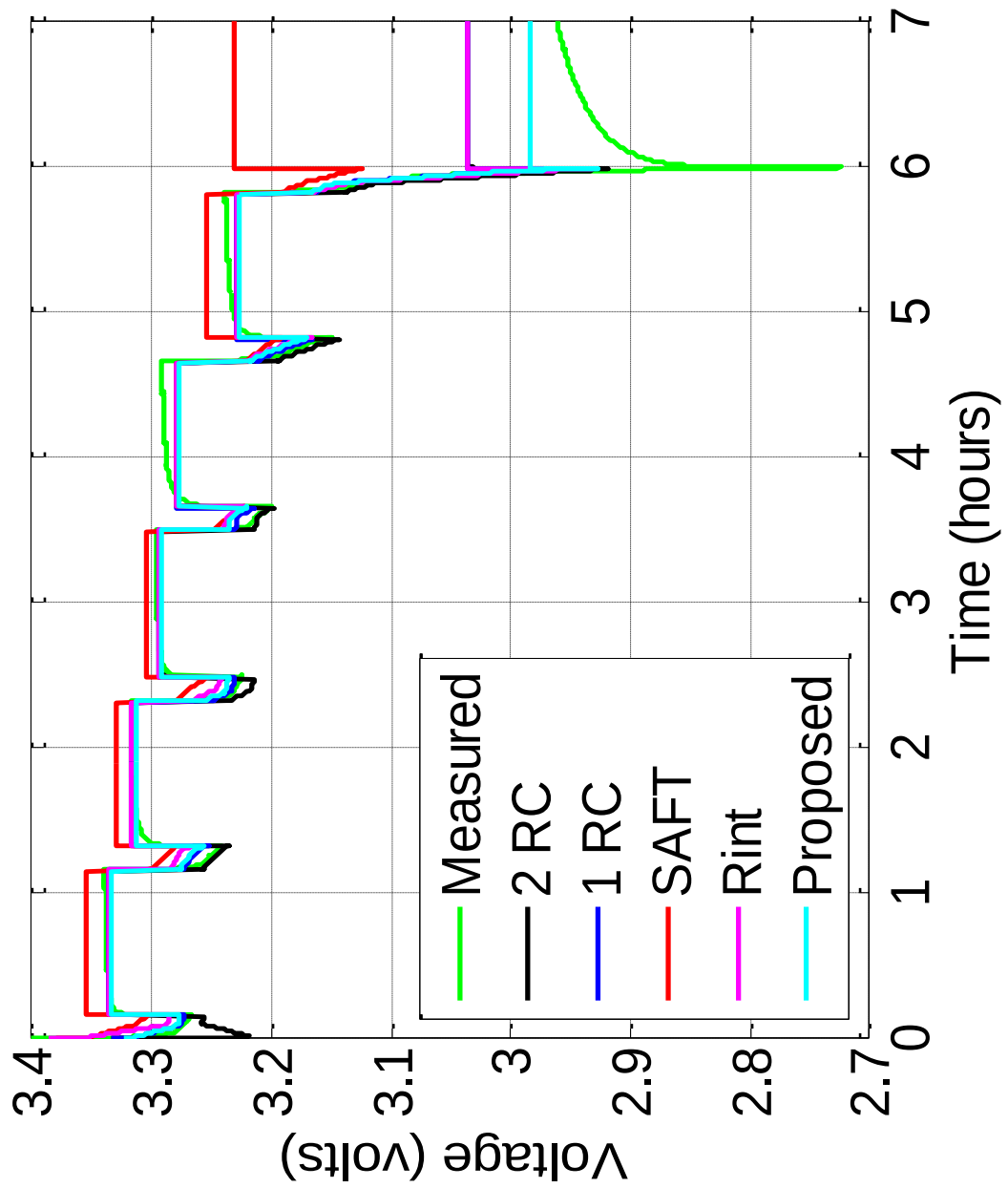


Figure 3.29 Simulation Results of All Battery Models.

Table 3.6 Average Error for Each Battery Model.

Model	Average Error (%)
Rint	0.7024%
1 RC	0.674%
2 RC	0.69%
SAFT	1.96%
UW	0.452%

The average error during the transient pulses is shown in Table 3.7

Table 3.7 Average Error During the Transient Pulses for Each Battery Model.

Model	Average Error (%)
Rint	0.7217%
1 RC	0.52%
2 RC	0.68%
SAFT	1.41%
UW	0.56%

The SAFT model performs the worst during the transient pulses. However, the Rint model does not have as poor performance as depicted in [23] and [34]. This could be related to how the parameter values were determined or the difference in load profiles that were used.

The average error during the extreme operating points (SOC>80% and SOC<20%) is shown in Table 3.8. Battery behavior when full or empty is very different when compared to 20-80% SOC operation. This is evidenced by the fact that parameter values for the model components, as shown in Table 3.1, Table 3.2, Table 3.3, and Figure 3.9, are generally very different during the extreme operating conditions compared to the rest of the battery

operating points. The UW model has the best performance during the extreme operating conditions. It is able to show the rapid voltage decrease at the end of the test profile (see Figure 3.29) as well as come close to predicting the final steady state value. The UW model performs much better than the SAFT model because the SAFT model bulk capacitance is not able to take into account the sudden voltage drop because the capacitance is constant. A sudden drop in capacitance is needed to capture this event, as seen in bulk capacitance values of the UW model listed in Table 3.4. It should also be noted that the 2 RC model is normally better than the 1 RC. However, in this case, the 2 RC model does a poor job predicting in the 80%-100% range, as evidenced in Table 3.8 and Figure 3.29. The 2 RC model performance is better than the 1 RC model performance when you consider only 0%-80% operation.

Table 3.8 Average Error During Extreme Operating Points for Each Battery Model.

Model	Average Error – Pulse #1 (%)	Average Error – Pulse #6 (%)
Rint	0.82%	3.2%
1 RC	0.33%	3.24%
2 RC	1.22%	3.26%
SAFT	1.29%	9.3%
UW	0.38%	1.631%

3.7 Battery Modeling Conclusions and Future Work

The proposed UW model performs the best out of the models that were investigated. The method for determining the parameter values is straightforward. There is no need for optimization software or tests such as impedance spectroscopy. The model exhibits good performance in both transient and steady state conditions. Overall, the model is very simple as it consists only of a single capacitor and a single resistor.

There is future work that can be performed. Currently, a current sensor is needed in order to obtain the SOC estimation in all of the battery models. It would be interesting to see how big of a drop in performance would result if the current sensor was swapped out for a voltage sensor at the terminals. Also, there can be more analysis done to verify the performance of the battery models. Different load profiles, such as those in [23] and [34] can be used to see if the average error remains consistent with the results when the load profile shown in Figure 3.23 is used.

Chapter 4. DC-DC Converter Design

4.1 Introduction to DC-DC Converter Design

In this chapter, a bidirectional DC-DC converter is proposed that can be employed for various uses in a battery management system, including equalization and charging/discharging. Bidirectional operation is necessary in equalization applications because power needs to be transferred in either direction, depending on which battery voltage is higher.

Series-connected batteries are widely used in high voltage and high power applications, including energy storage devices for renewable energy systems, electric vehicles (EV), and hybrid electric vehicles (HEV). Due to manufacturing variance, aging degradation, imbalance between charging and discharging, and differences in thermal conditions, internal impedances, and the self-charging rate, the energy stored in any two batteries is not equal. These mismatches reduce the efficiency, capacity, and lifetime of the batteries and can cause safety concerns. Battery equalizers can reduce this mismatch and improve performance, thus making equalizers a very important aspect of future battery management systems [12-15]. Another promising application could be the use of second life batteries taken from HEVs and EVs in other stationary applications. These batteries will need equalizers because each battery will degrade differently as cycle count is increased.

The proposed resonant circuit topology will allow for a reduction in size and cost because the size of the passive components, especially magnetics, will be greatly reduced. The very high frequency (VHF) technology will make it possible to integrate the

equalizer/charger on or in battery modules themselves, which would make the installation of such a system much easier for a customer.

4.2 Topology Selection

A SEPIC converter was chosen as the topology for the DC-DC converter. It is similar to the topology in [43] except that the rectifier portion of the circuit is different. The resonant rectifier for the proposed topology was chosen so that the circuit is symmetrical, leading to easy implementation of bidirectional operation. The proposed topology is shown in Figure 4.1

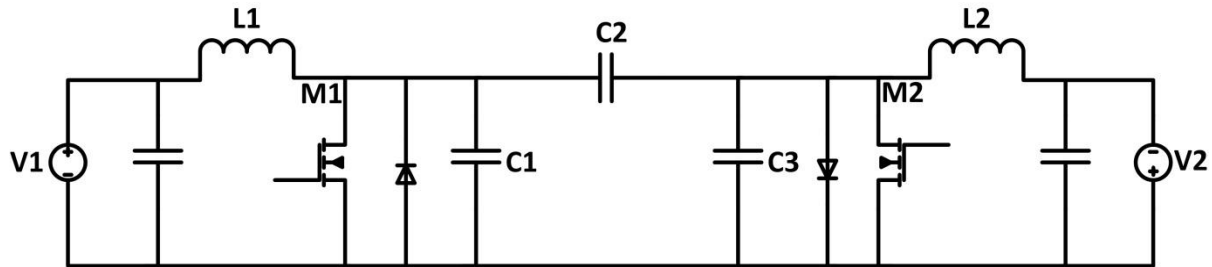


Figure 4.1 Schematic of the proposed resonant SEPIC converter topology.

Other topologies such as a Class E rectifier, as in [44]-[45], were considered. It was ultimately determined that the topology in Fig 4.1 provided the easiest way to achieve bidirectional operation. Assuming zero voltage switching (ZVS) can be obtained, the symmetry allows output power for left to right operation to equal the output power for right to left operation.

The proposed circuit represents an improvement over the solution in [43] because input and output voltages are clamped by batteries and low frequency (≈ 170 kHz) ON-OFF

modulation to regulate the output is not necessary for this design. The direction of power flow is what needs to be controlled, and once the converter is on, the quantity of power flow will accumulate to the mismatched energy of two battery cells. Therefore, the proposed circuit is smaller and the bandwidth is larger because the input and output filter components do not need to be sized for a lower ON-OFF modulation frequency as in [43].

Energy must be able to be transferred from V_1 to V_2 and vice versa. Because of the symmetry, L_1 and L_2 are equal in value and C_1 and C_3 are equal in value. The circuit is designed such that it operates at Zero Voltage Switching (ZVS). The batteries are modeled as ideal voltage sources (V_1 and V_2) with filter capacitors.

If M_1 is switching (with M_2 always OFF), power flows from left to right. L_1 , C_1 , and C_2 are resonant in this case. For equalization applications, M_1 will be switching if the V_1 is greater than V_2 .

If M_2 is switching (with M_1 always OFF), power flows from right to left. L_2 , C_3 , and C_2 are resonant in this case. For equalization applications, M_2 will be switching if V_2 is greater than V_1 .

4.3 Converter Power Rating

Before the component values of the DC-DC converter can be determined, the power rating of the circuit should be defined. In order to do this, some assumptions must be made about the input and output voltages, V_1 and V_2 .

For the initial design, it was assumed that the converter would be used to equalize batteries on a module level. Assuming a module is four cells connected in series, the input

and output voltage for the converter is nominally 13.2 V. This value is obtained when you consider the cell to be the A123 3.3 V batteries mentioned in the previous chapter. In order to boost the power consumed by a theoretical module, it is assumed that two parallel connected batteries make up a single cell. Therefore, the nominal power of a module is $(2 \times 2.3 \text{ A}) \times 3.3 \text{ V} \times 4$, which is approximately equal to 60 W.

Equation (4.1), taken from [46], shows the relation between energy mismatch of a cell (mis_{cell}), energy mismatch of a module (mis_{mod}), and the number of cells within a module (n).

$$mis_{mod} = \frac{mis_{cell}}{\sqrt{n}} \quad (4.1)$$

If a worst case cell mismatch is assumed to be 20%, the module mismatch would be 10%, meaning that the DC-DC converter should be rated for 6 W operation. The switching frequency is defined to be 5 MHz. This is an improvement over typical equalizers on the market today, which usually max out at a frequency of 1 MHz. It is important to note that the approach of designing for a theoretical module level equalizer is not the only application for this converter (the module design was simply used as a starting point). This topology can be used for a range of voltages and power levels. It is just necessary to determine a power rating before one can set component values.

4.4 Converter Component Value Selection

To select component values, an approach similar to the one in [43] was used. The inverter portion and the rectifier portion were originally tuned separately to get a starting point for component values. Assuming M1 is switching and power is flowing from left to

right, the inverter is boxed in red in Figure 4.2 and the rectifier is boxed in green. Assuming M2 is switching and power is flowing from right to left, the inverter is boxed in red in Figure 4.3 and the rectifier is boxed in green. The resonant components depend on which switch is active, as described in Section 4.1.

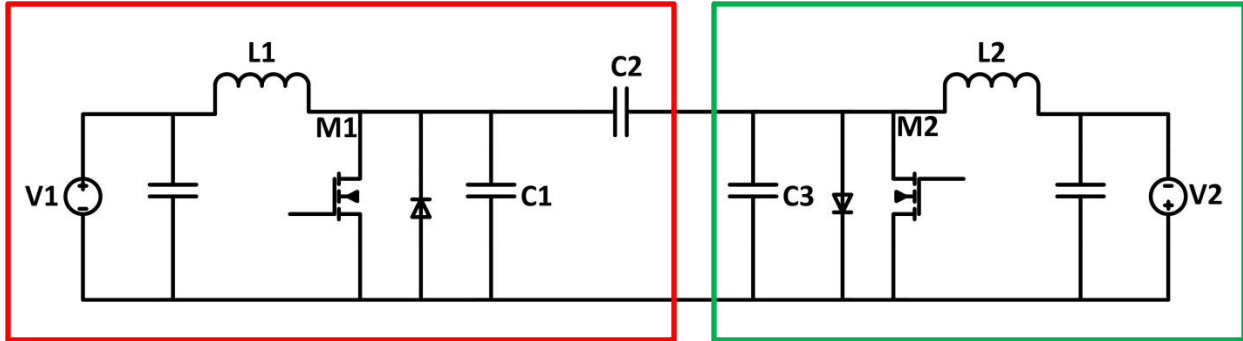


Figure 4.2 Inverter (red) and Rectifier (green) circuits for M1 Switching.

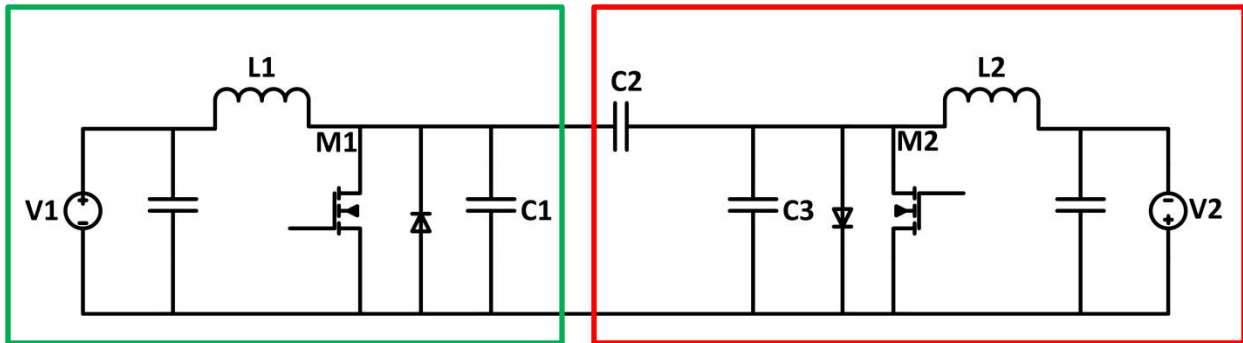


Figure 4.3 Inverter (red) and Rectifier (green) circuits for M2 Switching.

However, the process to determine the component values is not exactly the same as in [43]. There is no symmetry requirement in [43]. Therefore, after initial values for components were determined, a good deal of trial and error using LT Spice was necessary in order to meet the symmetry, ZVS, and power rating requirements.

Slightly adjusting the duty ratio was the main tool used to achieve ZVS. It was found that adjusting C1 and C3 within a small range could be used as a means to control the power

throughput while maintaining ZVS. The sensitivity of the circuit to changes in passive component values will be explored in a later section. The component values and duty ratio that were determined with this process are shown in Table 4.1. The output power was deemed acceptable if it was within 20% of the desired 6 W.

Table 4.1 Component Values for 6W DC-DC Converter.

Parameter	Value
L1	120 nH
L2	120 nH
C1	2650 pF
C2	1800 pF
C3	2650 pF
Duty Ratio	47%

4.5 Converter Simulation Results

As mentioned previously, simulation was performed using LT Spice. The setup is shown in Figure 4.4.

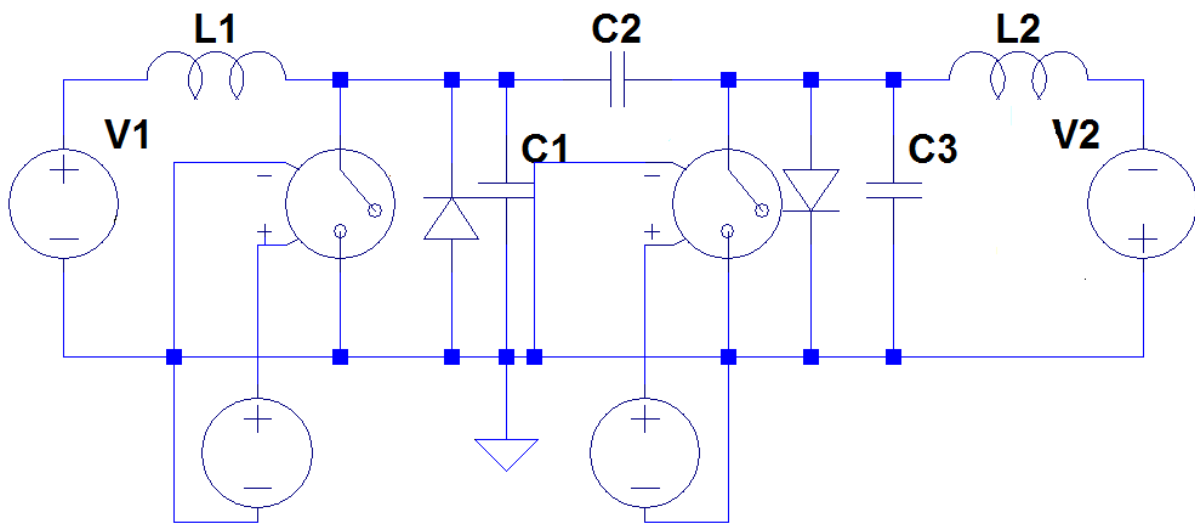


Figure 4.4 LT Spice Simulation Setup.

After components are selected, series resistances can be estimated for the inductors and switches and forward voltage drops can be determined for the diodes. This will allow for an efficiency estimate. The component values listed in Table 4.1 were used. Simulation results are shown in Figure 4.5 and Fig 4.6. ZVS is achieved for power flowing in either direction.

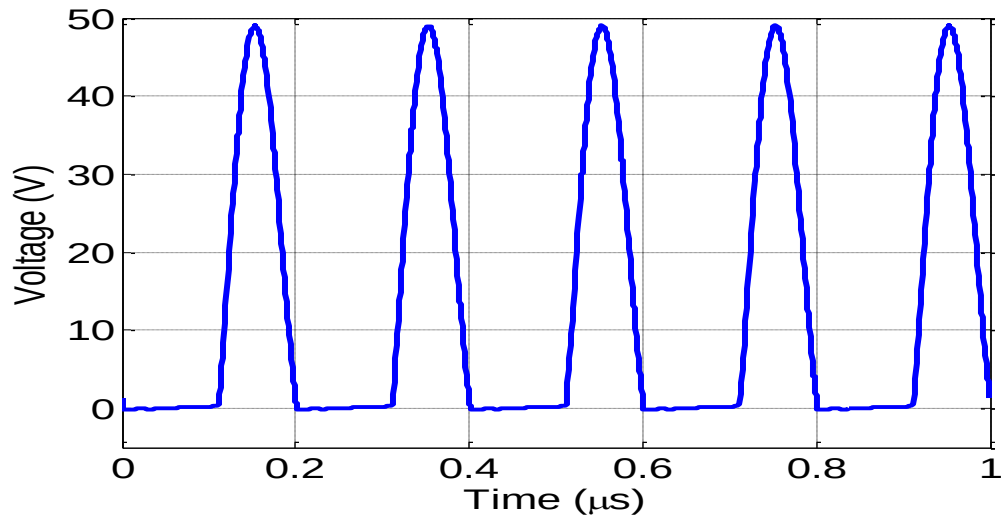


Figure 4.5 Converter Simulation Results for M1 Switching.

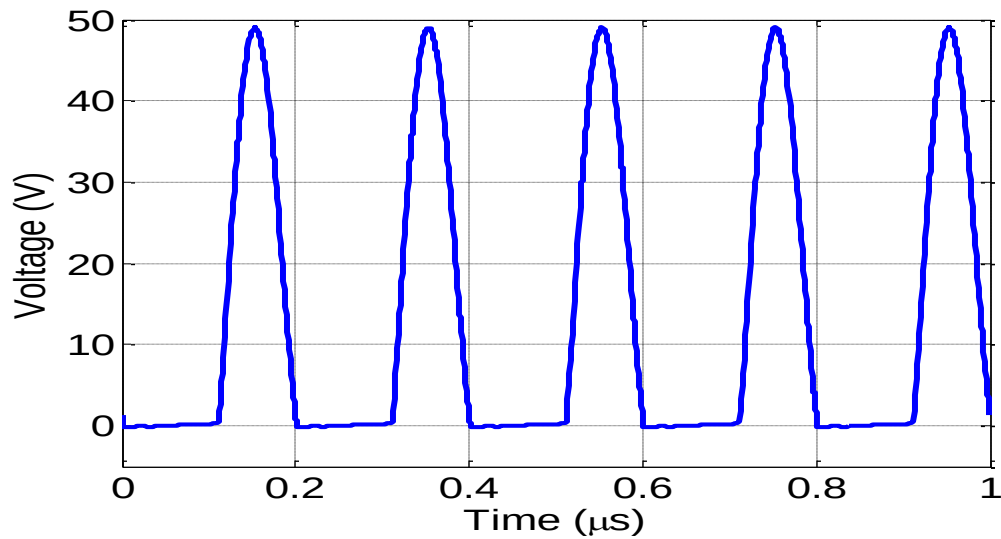


Figure 4.6 Converter Simulation Results for M2 Switching.

The simulation results can also give an estimate for the necessary voltage and current ratings for the various components. These ratings are listed in Table 4.2. Note that Table 4.2 lists the minimum quantities that the components must be able to withstand. In a real design, the ratings of the components will be higher than these minimum ratings. To obtain the ratings, simulation was performed with M1 switching. The components that are part of the inverter withstand the highest stress in every case except the diode. The voltage peak for the diode is from the inverter diode but the average current rating comes from the rectifier diode. The inverter diode never turns on so there is never current through it. Since the circuit can operate in either direction, all five sets of components (L1/L2, C1/C3, M1/M2, and both diodes) must be rated assuming the component will be part of the inverter (or the rectifier in the case of the diode). For example, with M1 switching, the RMS current of L1 is 4.58 A and the RMS current of L2 is 3.02 A. However, L2 must be rated for the 4.58 A since this value will be the RMS current of L2 when M2, not M1, is active.

Table 4.2 Necessary Voltage and Current Ratings for Circuit Components.

Component	Rating
L1/L2	$I_{L,rms} = 4.58 \text{ A}$
C1/C3	$V_{pk} = 48.8 \text{ V}$
C2	$V_{pk} = 70 \text{ V}$
M1/M2	$V_{pk} = 48.8 \text{ V}, I_{DS,rms} = 2.13 \text{ A}$
Diodes	$V_{pk} = 48.8 \text{ V}, I_{D,ave} = 590 \text{ mA}$

4.6 Converter Component Selection

Before accurate simulation can be performed, components must be defined so component loss can be estimated. When applicable, the dimensions of each component will

be provided because size of the converter is very important in this research. Also, figures from component data sheets will be provided to aid in the design process.

4.6.1 Inductor Selection

RF air core inductors (2222SQ) from Coilcraft were selected because of their high current and high frequency operation. The dimensions are shown in Figure 4.7.

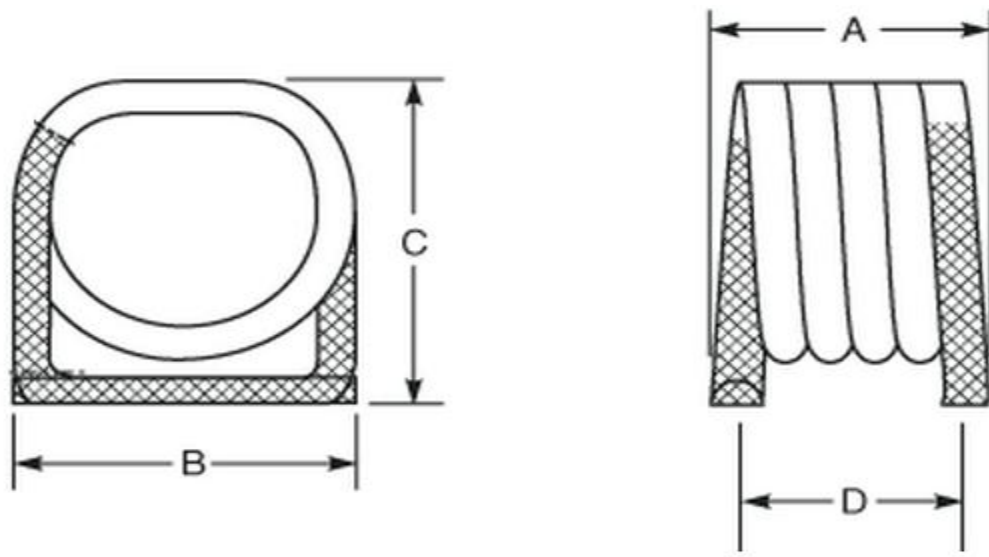


Figure 4.7 Coilcraft RF Inductor Dimensions [47].

Table 4.3 summarizes the dimensions shown in Figure 4.7.

Table 4.3 Coilcraft RF Inductor Dimensions.

A	6.35 mm
B	5.59 mm
C	5.69 mm
D	5.84 mm

The dimensions of this inductor are very small, which is important because the size of the magnetics is the main determining factor of the size of the converter.

Figure 4.8 shows how the quality factor of these inductors varies with frequency.

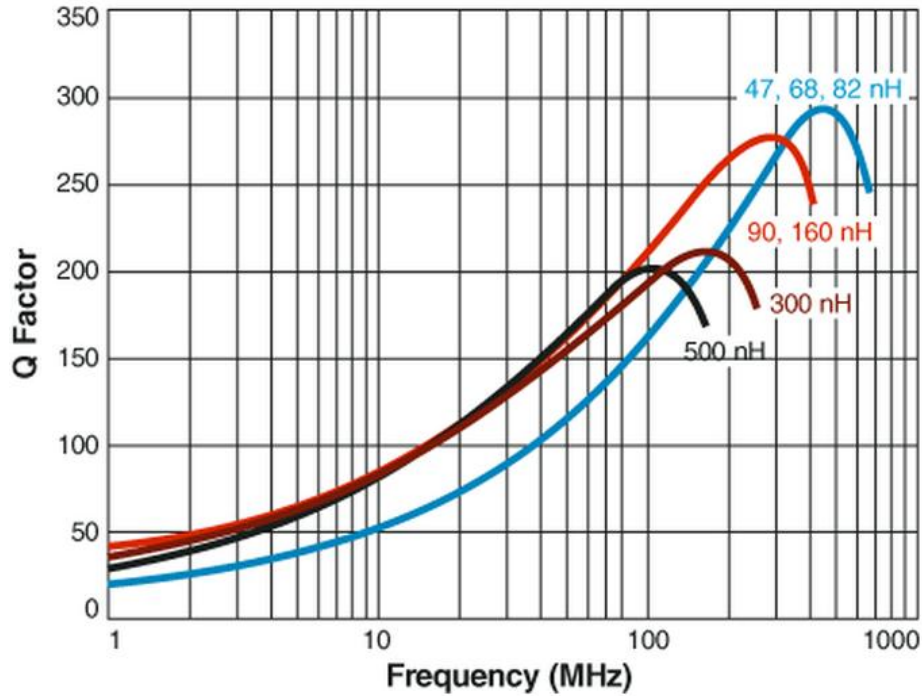


Figure 4.8 Quality Factor vs. Frequency for Coilcraft Inductors [47].

Therefore, these inductors will have a quality factor of approximately 60 for a switching frequency of 5 MHz. The RMS current rating of the inductors is 5.7 A, which meets the requirements from Table 4.2.

4.6.2 Diode Selection

A Schottky diode (STPS5L60) from STMicroelectronics was chosen for the two diodes shown in Figure 4.1. The dimensions of the diode are shown in Figure 4.9. The units are inches, with millimeters in parentheses. The diode is rated for 60 V and an average forward current of 5 A. Therefore, this diode meets the requirements shown in Table 4.2.

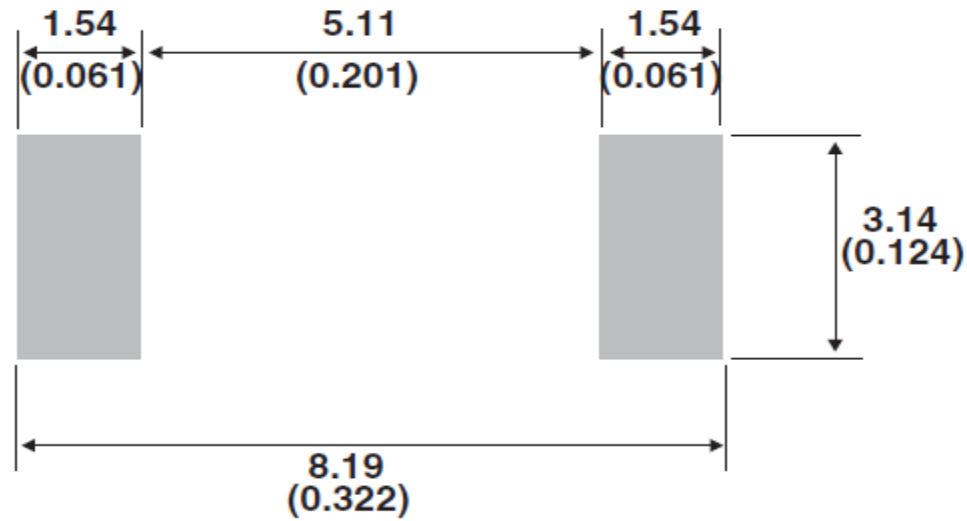


Figure 4.9 STMicronelectronics Diode Dimensions [48].

The device capacitance will also be of interest when it comes time to layout the printed circuit board. The device capacitance vs. forward current is shown in Figure 4.10.

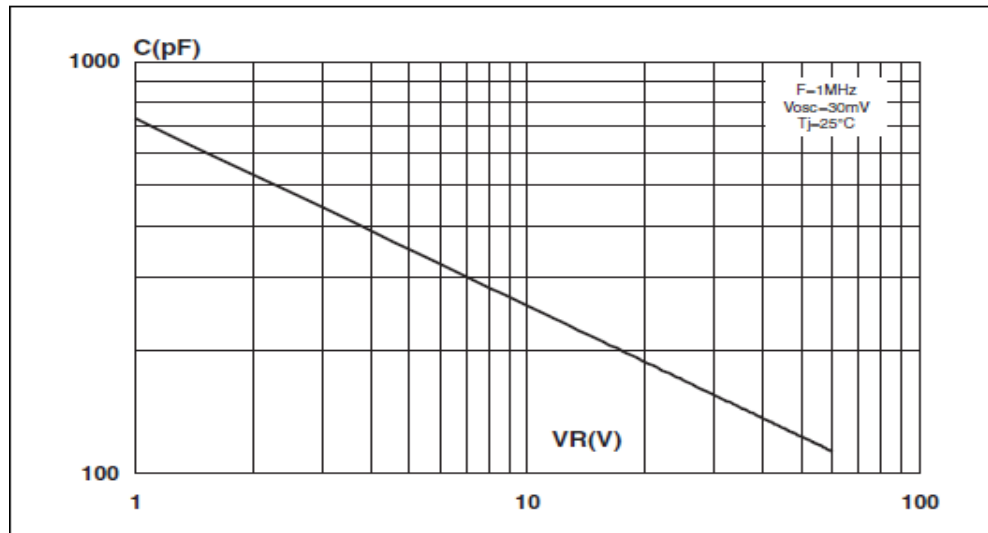


Figure 4.10 STMicronelectronics Diode Capacitance vs. Forward Current [48].

The diode forward voltage versus current is shown in Figure 4.11. This value is necessary when trying to make an efficiency estimate.

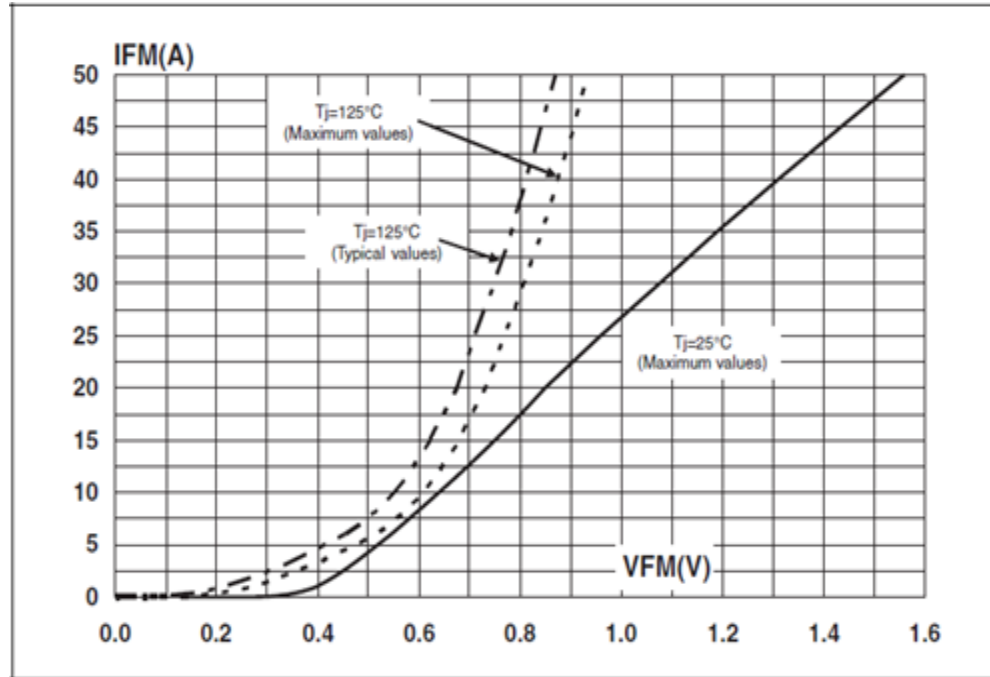


Figure 4.11 STMicroelectronics Diode Forward Current vs. Forward Voltage [48].

4.6.3 MOSFET Selection

An N-Channel Power Trench MOSFET (FDS86106) from Fairchild Semiconductor was chosen for M1 and M2. This part comes in a standard SO-8 package, shown in Fig 4.12. This switch is rated for a voltage of 100 V and a current of 3.4 A, which meet the requirements of Table 4.2. The other figures of interest for the MOSFET are shown in the next two figures. Figure 4.13 shows the normalized on resistance versus gate voltage and drain current, which is needed for efficiency estimates. For the MOSFET, the normal on resistance is 105 m Ω . Figure 4.14 shows the relationship between device capacitance and drain-source voltage, which is needed for PCB layout.

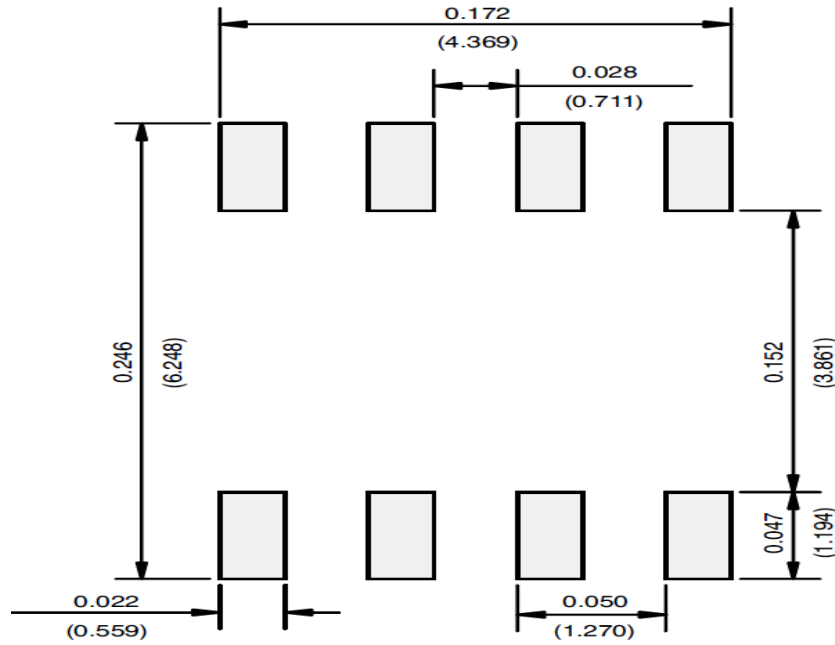


Figure 4.12 Fairchild Semiconductor MOSFET Dimensions [49].

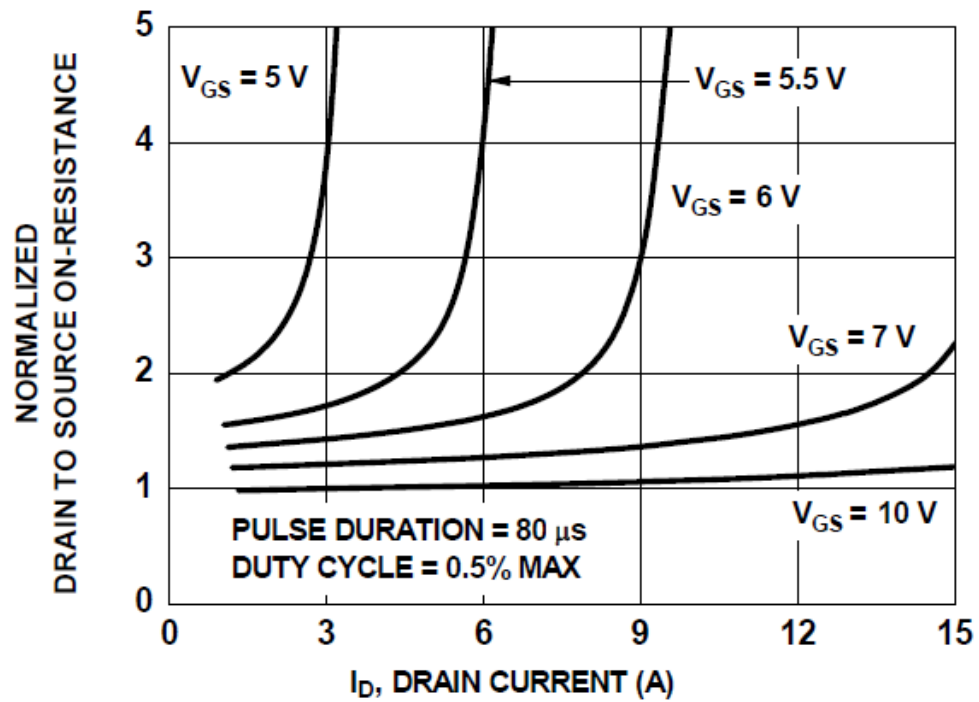


Figure 4.13 Fairchild Semiconductor On Resistance vs. Drain Current [50].

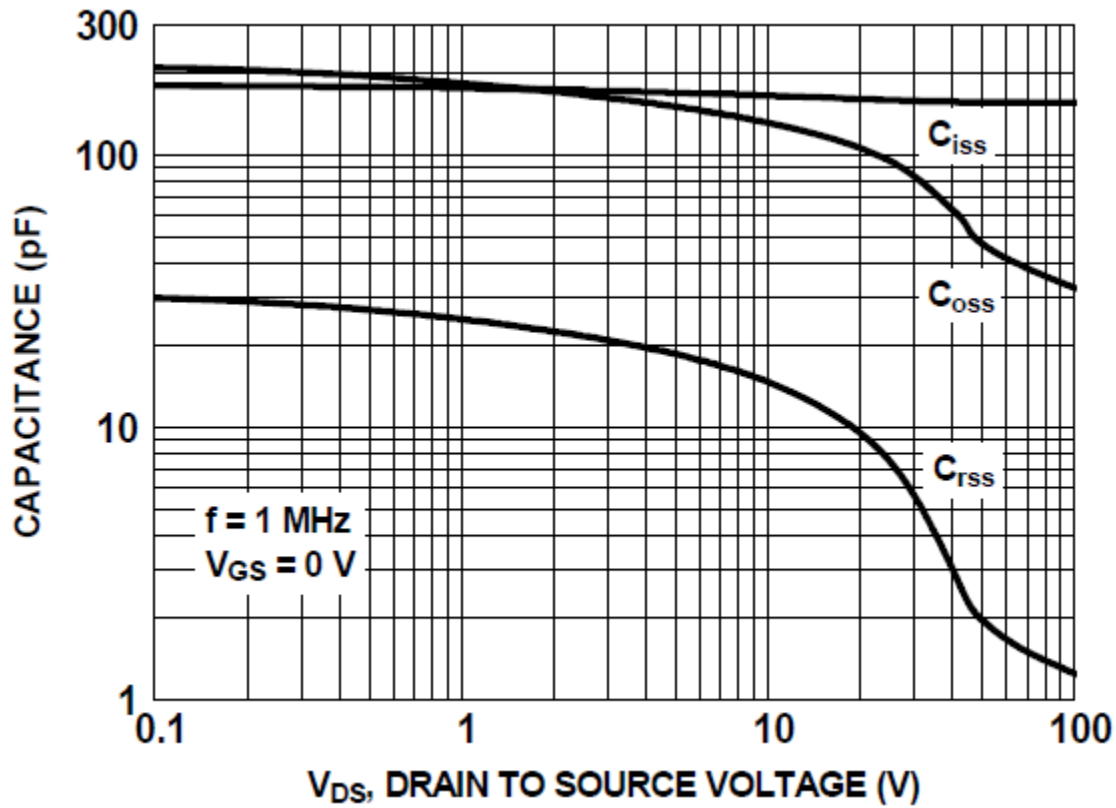


Figure 4.14 Fairchild Semiconductor Capacitance vs. Drain-Source Voltage [50].

The drain-source capacitance can be obtained by subtracting C_{rss} from C_{oss} .

4.6.4 Capacitor Selection

Capacitors (700B series) from American Technical Ceramic (ATC) were used. These capacitors are rated for 1500 VDC, which exceeds the ratings in Table 4.2. The 700B capacitors have a very low equivalent series resistance so capacitor loss was ignored during efficiency calculations. The recommended solder pad dimensions for a single capacitor are shown in Figure 4.15. Table 4.4 summarizes the dimensions shown in Figure 4.15.

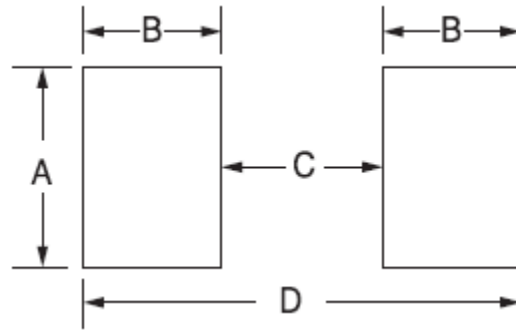


Figure 4.15 ATC Capacitor Pad Dimensions [51].

Table 4.4 ATC Capacitor Pad Dimensions.

A	3.30 mm
B	1.27 mm
C	1.90 mm
D	4.45 mm

4.6.5 Gate Driver Selection

A SiLabs (Si8235AB-C-IS1) gate driver was used. The driver can switch up to 8 MHz and can be referenced to a floating voltage. This last fact will be useful when connecting the driver for M2. The pad dimensions are shown in Figure 4.16. The package is a 16 pin Narrow Body SOIC PCB. Table 4.5 summarizes the dimensions shown in Figure 4.16.

Table 4.5 SiLabs Gate Drive Dimensions.

C1	5.4 mm
E	1.27 mm
X1	0.6 mm
Y	1.55 mm

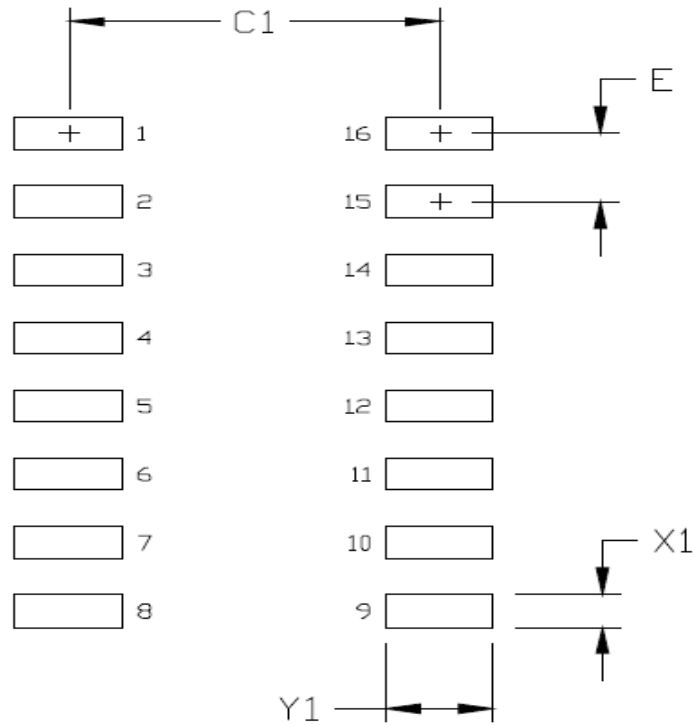


Figure 4.16 SiLabs Gate Drive Dimensions [52].

4.6.6 Component Selection Conclusions

All components that were selected meet the ratings requirements listed in Table 4.2.

Table 4.6 shows the component listing.

Table 4.6 Component Listing.

MOSFETS	Fairchild FDS86106
DIODES	STMicroelectronics STPS5L60
INDUCTORS	Coilcraft Air Core 2222SQ
CAPACITORS	ATC Multilayer 700B
GATE DRIVER	SiLabs Si8235

4.7 Efficiency Estimation

LT Spice was used to estimate the efficiency of the converter. The figures from section 4.5 were used to estimate the loss parameters of each component. The series

resistance of the inductors, the series resistance of the MOSFETs, and the forward voltage of the diodes were considered for this estimate.

Figure 4.8 shows that the quality factor of the Coilcraft inductor is approximately 60. Using (4.2), the series resistance can be estimated to be 57.6 m Ω for an inductance of 110 nH.

$$Q = \frac{\omega L}{R} \quad (4.2)$$

Table 4.1 lists the inductance value as 120 nH. However, the closest inductance that Coilcraft has is 110 nH. As a result, the other component values had to be slightly adjusted to maintain ZVS. The new values are listed in Table 4.7.

Table 4.7 Converter Component Values after Inductance was Adjusted.

Parameter	Value
L1	110 nH
L2	110 nH
C1	2600 pF
C2	1800 pF
C3	2600 pF
Duty Ratio	52.5%

LT Spice shows that the average forward current of the diode that is part of the rectifier is approximately 450 mA. Figure 4.11 shows that the forward voltage can be estimated to be approximately 0.3 V.

LT Spices shows that the RMS current through the MOSFET is approximately 2.57 A. For a drive voltage of around 10 V, Figure 4.13 shows that the MOSFET on resistance can

be estimated to be 105 m Ω . To decrease the resistance, two MOSFETS were connected in parallel, which decreases the on resistance by a factor of two.

Table 4.8 summarizes the loss components for the different components.

Table 4.8 Loss Values for Different Components.

Component	Loss Parameter
L1/L2	R=57.6 m Ω
Diodes	V=0.3 V
M1/M2	R=52.5 m Ω

Using these component values, LT Spice estimate that the efficiency is 72.2% with a total power loss of 2.28 W. Table 4.9 shows the loss by component as well as the percentage of total loss.

Table 4.9 Estimated Power Loss by Component.

Component	Power Loss	% of Total Loss
L1	1.42 W	62.3
L2	0.42 W	18.4
MOSFET	0.35 W	15.4
Diode	89.1 mW	3.90

Table 4.9 shows that the inductors make up 80.7% of the total loss. If improvements to the inductors can be made, the efficiency can be significantly improved. This will be addressed in a future section.

4.8 Printed Circuit Board Design

Eagle was used to design the printed circuit board. In order to minimize parasitics that result from PCB traces, all of the power circuit components were placed as closely as

possible to each other. The schematic and PCB layout is shown in Figure 4.17 and Figure 4.18, respectively.

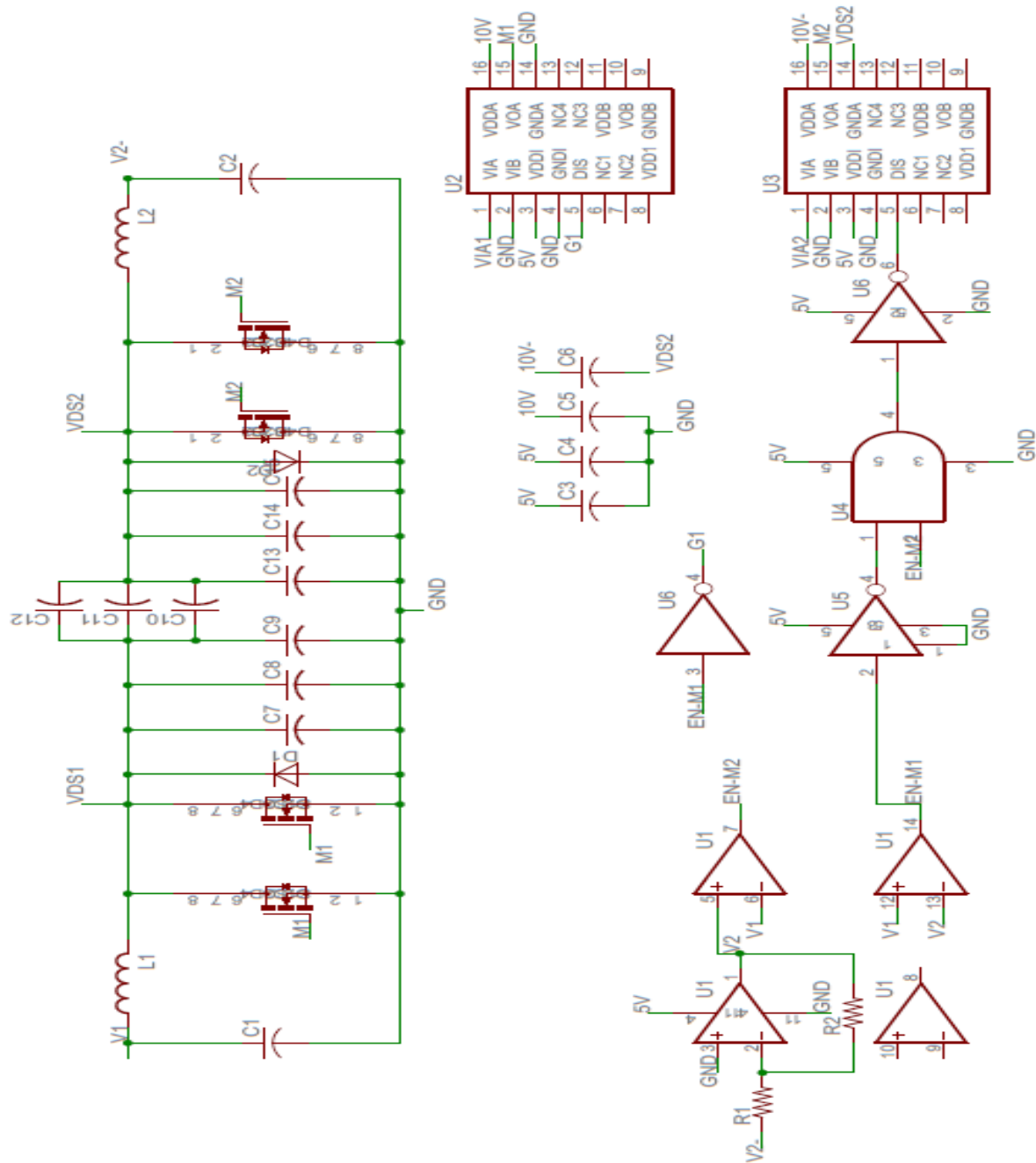


Figure 4.17 Eagle Schematic for DC-DC Converter.

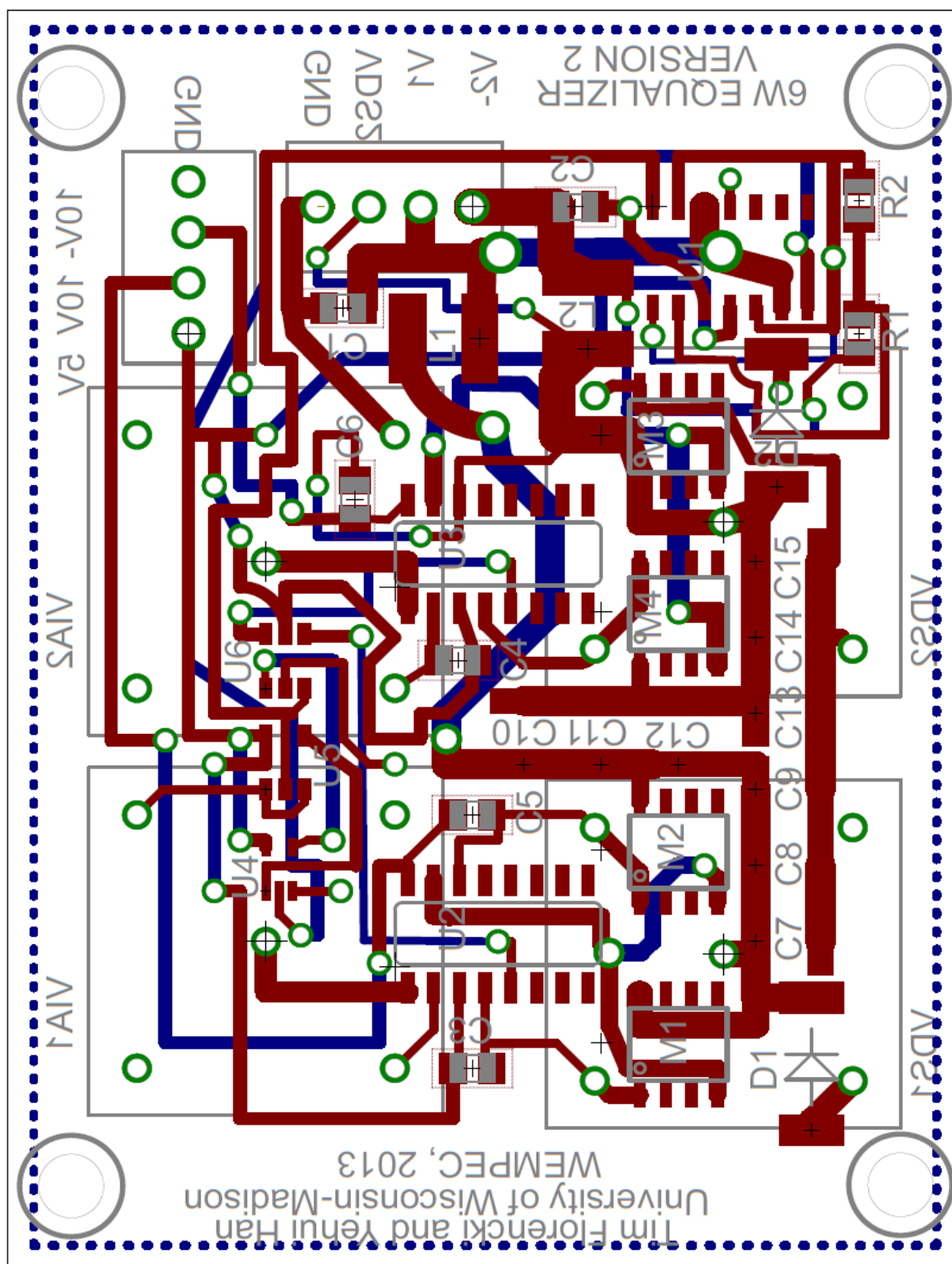


Figure 4.18 Eagle PCB for DC-DC Converter.

There are a few things to note about these figures.

- 1) Extra capacitors were placed on the PCB to make debugging easier. C7, C8, and C9 of Figure 4.17 make up C1 in Figure 4.1. Likewise, C10, C11, and C12 make up C2 in Figure 4.1. Lastly, C13, C14, and C15 make up C3 in Figure 4.1.
- 2) Filter capacitors (C1 and C2 in Figure 4.17) were placed at the input and output. The size of these capacitors was chosen to be 10 μF . An 0805 package was used in order to keep the size small.
- 3) As mentioned previously, two MOSFETs were connected in parallel to decrease the on resistance. In a final design, a better MOSFET can be chosen so that only one is needed, further decreasing the size of the power circuit.
- 4) C3, C4, C5, and C6 in Figure 4.17 are input and output capacitors necessary for the gate drivers (U2 and U3) to operate properly. Their value is 0.1 μF . An 0805 package was used in order to keep the size small.
- 5) Also pictured in Figure 4.17 is one potential solution to sense which voltage is greater, which is important in determining which switch should be active. U1 inverts V2, which has an opposite polarity of V1 (see Figure 2.1 and Figure 4.1). V1 and V2 are then compared to see which voltage is greater. This is also done by U1. If V1 is greater than V2, the signal EN-M1 is high, G1 is low, which activates M1. EN-M2 is low, which makes the signal into pin 5 of U3 high, which deactivates M2. The opposite of this scenario is true if V2 is greater than V1.
- 6) U4 and U5 serve as protection circuitry that prevents M1 and M2 from being active at the same time.

- 7) Pin 5 of the gate drivers (U2 and U3) is the disable pin. If high, the gate driver is not active. If low, the gate driver is active.
- 8) Vertical BNC connectors were used as a means for an easy way to obtain the drain-source voltages of each switch.
- 9) A signal generator was used to generate the PWM signal fed into pin 1 of the gate drivers. BNC connectors were also used as the interface. It is very difficult to find a PWM chip that generates VHF PWM signal and allows for precise duty cycle control. It is believed that a chip with these capabilities will be available in the future. Conversely, an FPGA could accomplish this but it is thought that a solution like this is too complicated and bulky for this design.
- 10) The output of U2 is referenced to the source of M2. This is possible because the SiLabs gate driver can be referenced to a floating voltage.
- 11) Wire connectors were used to connect the voltage sources (V1 and V2) and the 5 V and 10 V supplies to the PCB.

Table 4.10 lists the component part numbers that were used for U1, U4, U5, and U6. They are just standard logic integrated circuits that can be found on a site like Digikey.

Table 4.10 Voltage Sensing Circuitry Part Numbers.

Component	Part #
U1	LMC660
U4	SN74LVC1G08DCKR
U5	SN74LVC1GU04DBVR
U6	SN74LVC2GU04DBVR

4.9 Device Parasitics

As mentioned previously, device parasitics must be considered when populating the PCB. Parasitics from the switches and diode were considered. Figure 4.10 shows the diode capacitance as a function of reverse voltage. From this figure, it can be estimated that the diode capacitance is approximately 160 pF. Figure 4.14 shows the MOSFET capacitance as a function of drain-source voltage. From this figure, it can be estimated that the MOSFET capacitance is approximately 66 pF.

According to Table 4.7, the desired value of C1 and C3 is 2600 pF. The total device capacitance is approximately 226 pF. This means that the device capacitance makes up approximately 7% of the total desired capacitance of 2600 pF.

4.10 Experimental Results

To gain experimental results, a voltage source was used for the input voltage and a Zener diode was used at the output. The results are shown in Figure 4.19 (M1 switching) and Figure 4.20 (M2 switching). The results are promising because almost perfect ZVS is achieved. It is believed that the results are not identical due to some parasitics on the PCB. The measured efficiency was 72.9%. The next section will address how to improve the efficiency.

4.11 Efficiency Improvement

As mentioned previously, 80.7% of the loss is due to inductors. If the quality of the inductors are improved, the efficiency can be improved. There will be a tradeoff between size and efficiency improvement because it is still desirable to have the inductors as small as possible.

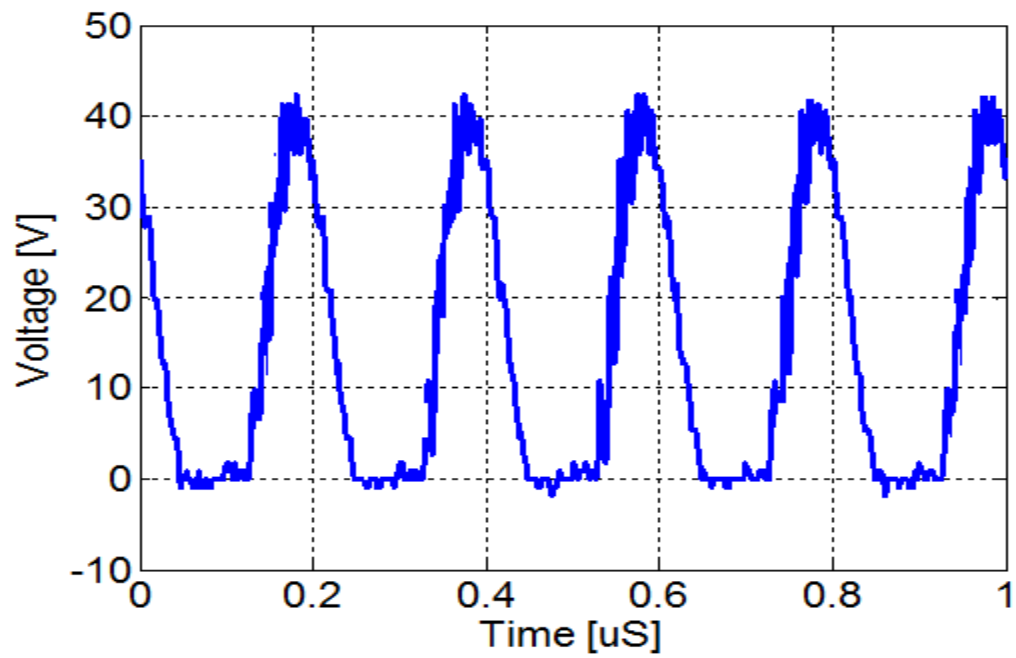


Figure 4.19 Experimental Results. M1 Active.

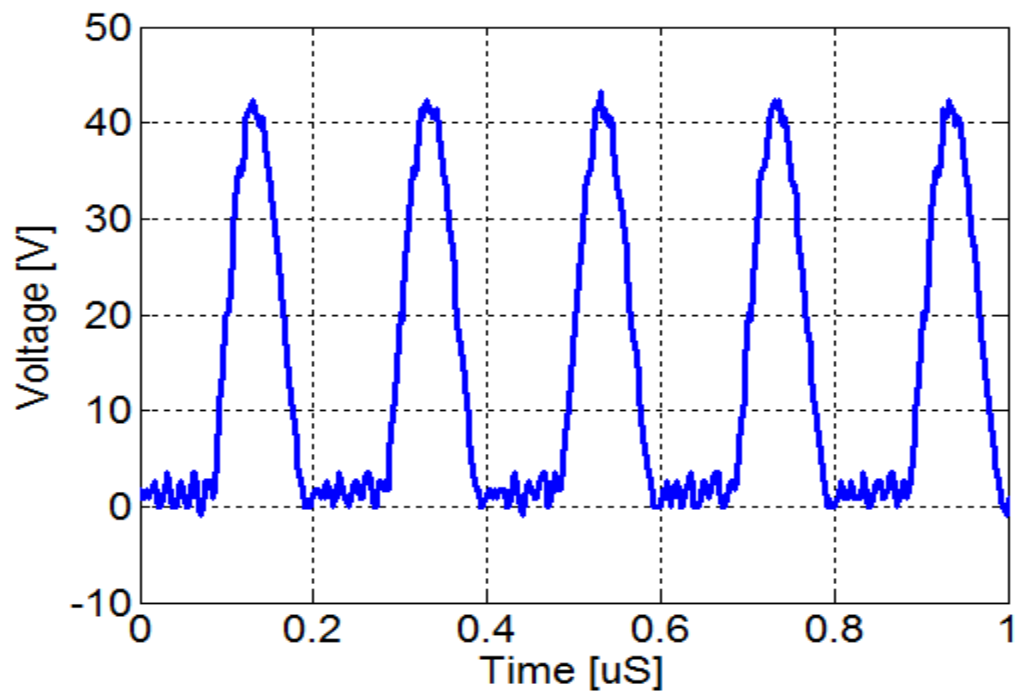


Figure 4.20 Experimental Results. M2 Active.

Doubling the quality factor of the Coilcraft inductors will yield an efficiency of approximately 80%.

Low permeability magnetic and non-magnetic toroids from Micrometals and Ceramic Magnetics were explored. The design approach detailed in [7] was used. Copper foil was used for the windings. A toroidal core is shown in Figure 4.21. The outer diameter, inner diameter, height, and material permeability can be used to estimate the inductance, as seen in (4.3).

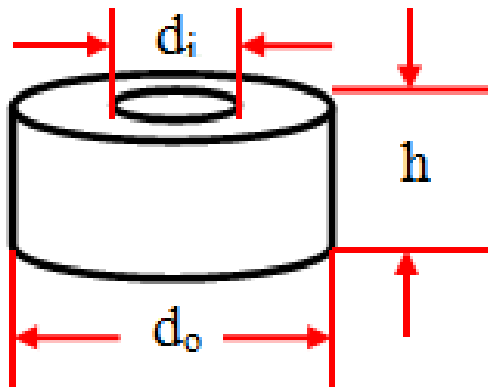


Figure 4.21 Basic Toroidal Core.

$$L = \frac{N^2 h \mu_r \mu_o}{2\pi} \ln \left(\frac{d_o}{d_i} \right) \quad (4.3)$$

The following are considerations for copper foil design.

- 1) The thickness of the copper foil must be much greater than the skin depth, as seen in (4.4). This will mitigate the effects of the skin effect. For 5 MHz, the skin depth is approximately 0.0296 mm.

$$t_{cu} \gg \delta = \sqrt{\frac{\rho_{cu}}{\pi \mu_o f_s}} \quad (4.4)$$

- 2) The width of the copper foil must be sized so the appropriate number of turns can fit along the circumference of the inner diameter, as seen in (4.5).

$$w_{cu} = \frac{\pi d_i}{N} \quad (4.5)$$

- 3) The length of the copper foil can be estimated according to (4.6).

$$l_{cu} = N(2h + d_i + d_o) \quad (4.6)$$

- 4) The resistance of the copper foil can be calculated according to (4.7) as long as (4.4) is observed.

$$R_{cu} = \rho \frac{l_{cu}}{w_{cu} \delta} \quad (4.7)$$

Two magnetic materials were tested. The first was NiZn material from Ceramic Magnetics, which has a permeability of 15. The quality factor information for this material is shown in Figure 4.22. The quality factor can be approximated using (4.8).

$$Q = \frac{\mu'}{\mu''} \quad (4.8)$$

At 5 MHz, μ' is 15 and μ'' is approximately 0.18. Therefore the quality factor of this core when disregarding copper loss is approximately 83.

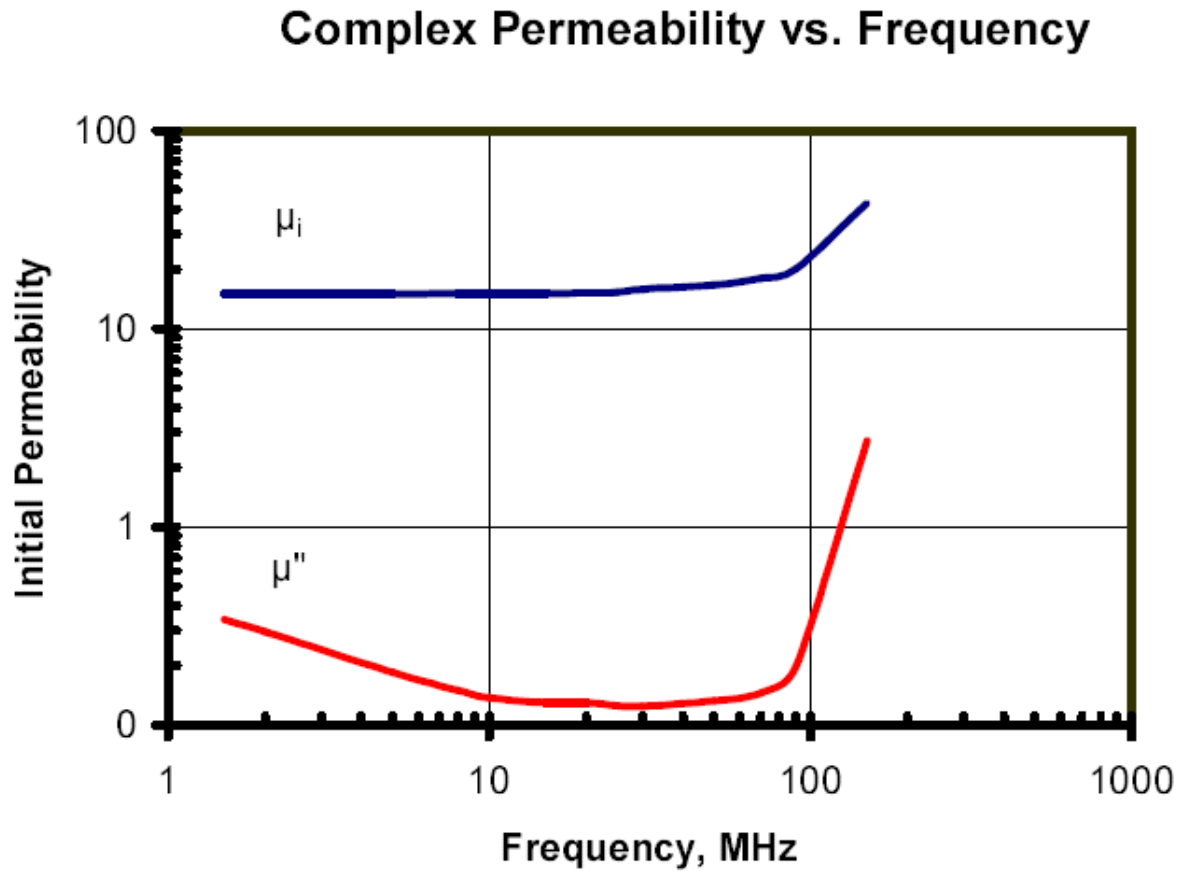


Figure 4.22 Complex Permeability vs. Frequency for N40 Material [54].

The dimensions and other relevant information for the N40 material inductor info are shown in Table 4.11.

Table 4.11 N40 Inductor Information.

Material	μ_r	d_o	d_i	h	N	L (nH)
N40	15	9.53	4.85	3.30	4	106.9

According to (4.2), the resistance due to the core is 40 m Ω . Assuming a copper width of approximately 3 mm, the resistance due to the copper can be estimated to be 8.6 m Ω .

Therefore the total resistance is approximately 48.6 m Ω . Using (4.2), the quality factor of the

Ceramic Magnetic inductor can be estimated to be 68.6. Therefore, it is expected that the efficiency of the converter will be slightly better with these inductors as opposed to the Coilcraft inductors. The measured efficiency was 74.5%. Figure 4.23 shows the PCB with the CMI inductors.

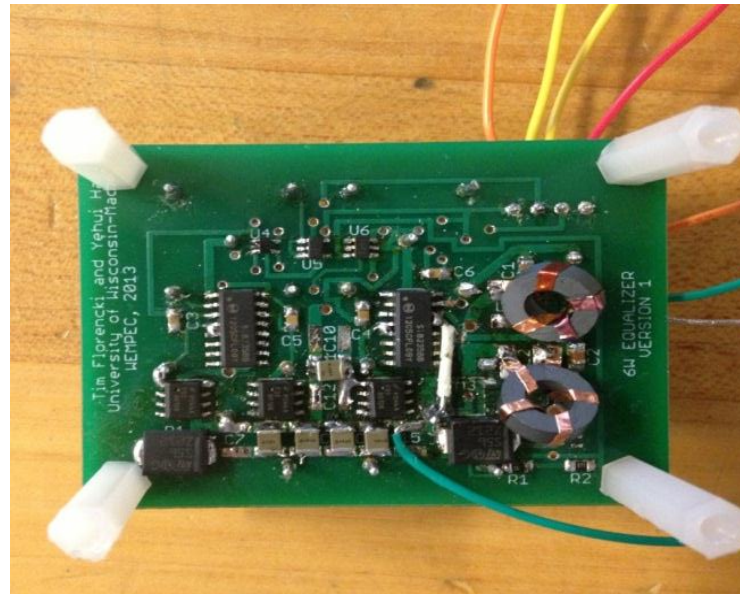


Figure 4.23 PCB with CMI Inductors.

It can be seen from this Figure that the size of the inductors is still very small.

A similarly sized core from Micrometals was explored. The material is called -2, which has a Carbonyl E iron power coating. The permeability is 10. The quality factor information for this material is shown in Figure 4.24. The specific core is T37-2. The quality factor neglecting copper loss can be estimated to be 185.

The dimensions and other relevant information for the -2 material inductor info are shown in Table 4.12. The size of this material is almost the exact same as the N40 material.

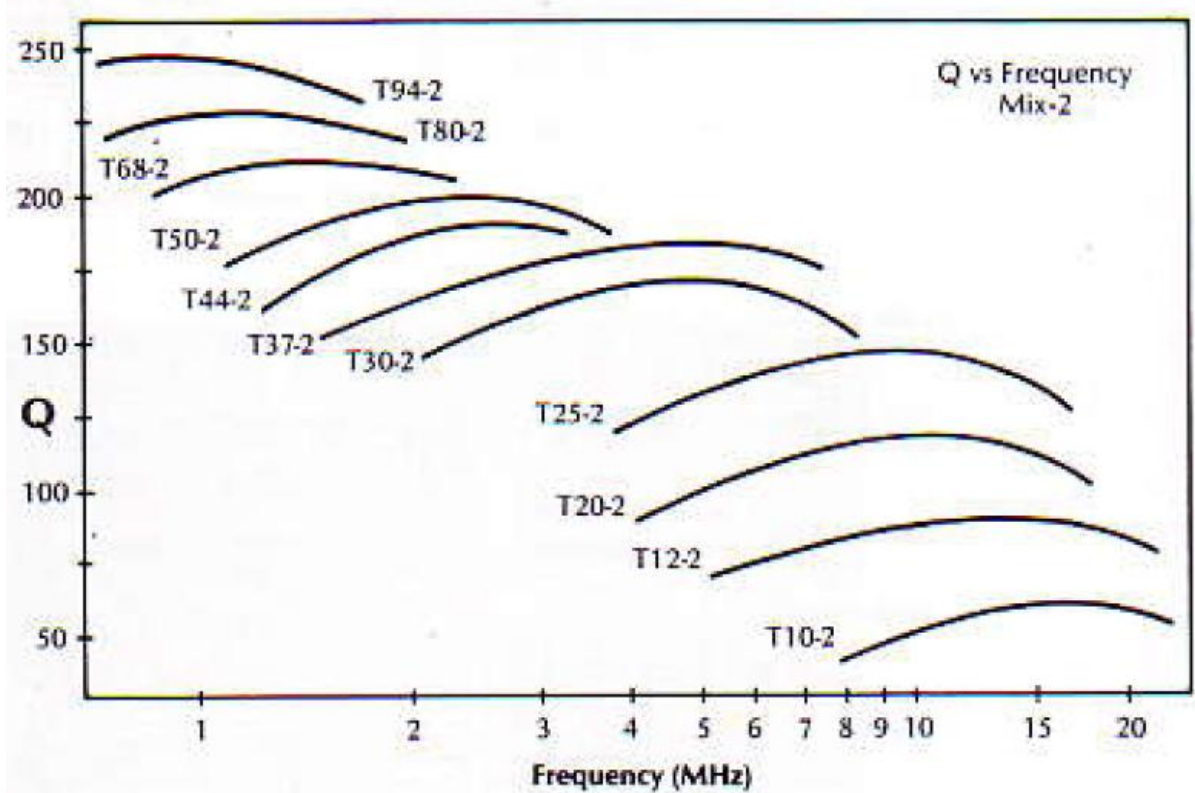


Figure 4.24 Quality Factor vs. Frequency for -2 Material [53].

Table 4.12 -2 Inductor Information.

Material	μ_r	d_o	d_i	h	N	L (nH)
-2	10	9.53	5.21	3.25	5	105

From (4.2), the resistance due to the core is 17.8 m Ω . Assuming a copper width of approximately 3 mm, the resistance due to the copper can be estimated to be 9.6 m Ω . Therefore the total resistance is approximately 24.4 m Ω . Using (4.2), the quality factor of the Micrometal inductor can be estimated to be 120.4. The measured efficiency was 81.1%, which is a very promising result. Figure 4.25 shows the PCB with the Micrometals inductors.

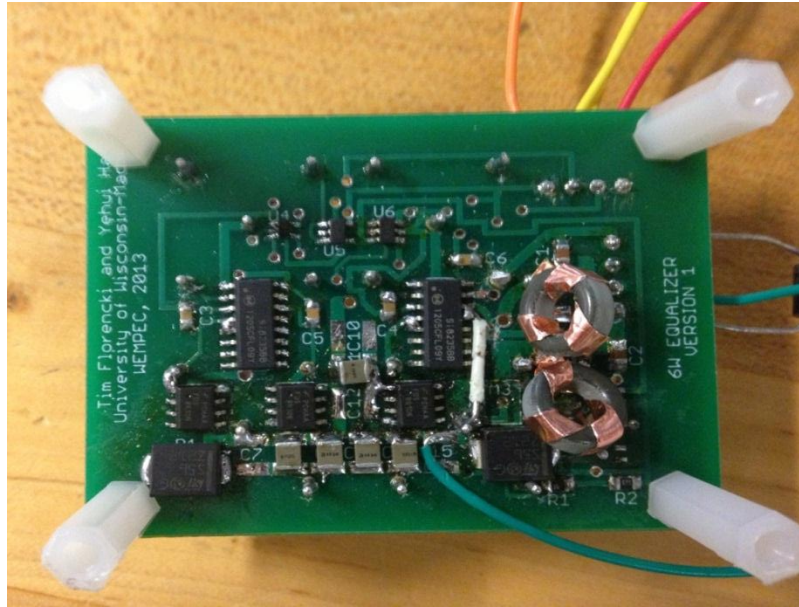


Figure 4.25 PCB with Micrometals Inductors.

As a reference, a similarly sized non-magnetic toroid from Micrometals (-0 material) was considered. The relevant information is shown in Table 4.13. Because the Q is so low, this toroid was not even tested.

Table 4.13 -0 Toroid Information.

Material	μ_r	d_o	d_i	h	N	L (nH)	Q
-0	1	9.53	5.21	3.25	15	110.3	31.9

4.12 Passive Component Variation

For certain topologies, circuit operation is very sensitive to component values changing. Slight variation in component values from the nominal values can cause problems, such as the circuit losing its zero voltage switching properties. For example, Class E

topologies are usually sensitive to slight variations in component values. Passive component aging can be seen as a big concern for some topologies. As the components age, their value decreases.

To get an idea of the robustness of the proposed topology, LT Spice was used to find how much each passive component value can be decreased until ZVS is lost. Each passive component was examined, while holding all other components at their nominal value. For example, when the inductors were examined, three scenarios were considered. They are:

- 1) Decrease L1 until ZVS is lost while keeping all other components, including L2, at their nominal value.
- 2) Once this value is obtained, simulate for power and efficiency measurements for M1 active and then M2 active.
- 3) Lastly, decrease L2 to the same value as L1 and simulate for power and efficiency measurements.

In this way, you are able to see how the circuit behaves if one of the inductors or both suffer from component variation. In the first two steps, the quantity of power flow will be different based on which switch is active, but ZVS will still be maintained. For simulations, it was assumed that the inductors are the Coilcraft inductors that were used in the initial design process.

4.12.1 Inductor Variation

The results of simulating inductor variation are shown in Table 4.14.

Table 4.14 Inductor Variation Simulation.

L1 (nH)	L2 (nH)	Active Switch	Power (W)	Efficiency (%)
65 nH	110 nH	M1	22.4	69.7
65 nH	110 nH	M2	9.85	76.8
65 nH	65 nH	M1 or M2	14.4	62.6

The inductors can be decreased by approximately 41% and still maintain zero voltage switching. In order to maintain ZVS, the diode that is in parallel with the active switch is on for a short period of time. In nominal operation, this diode is never on.

4.12.2 Capacitor Variation

The results of simulating C1/C3 variation are shown in Table 4.15.

Table 4.15 Capacitor Variation Simulation.

C1 (pF)	C3 (pF)	Active Switch	Power (W)	Efficiency (%)
1500	2600	M1	21.3	82.1
1500	2600	M2	10.9	79.3
1500	1500	M1 or M2	21.1	82.9

C1 and C3 can be decreased 42.3% and still keep zero voltage switching. In order to maintain ZVS, the diode that is in parallel with the active switch is on for a short period of time. In nominal operation, this diode is never on.

The variation of C2 was also investigated. It was found that C2 can be theoretically reduced to just above zero and ZVS will still be maintained. However, the output power will also go to zero.

4.12.3 Component Variation Conclusions

The preliminary results show that the proposed circuit topology has some robustness because it is not extremely sensitive to changes in component values. While the power level rises above the desired value, the predicted efficiency is similar to or higher than the predicted efficiency for the proposed topology. While the parallel diode turning on to maintain the switch is not desired for this application, it can be acceptable in other applications. Nevertheless, the robustness of the circuit topology displayed in this section is promising because it shows that there is some room to improve the current design. As mentioned previously, improving the quality of the inductors was a method used to improve the efficiency. This section shows that efficiency can potentially be improved by changing the power level of the converter. The future work should include finding the optimum voltage and power range to maximize the efficiency of the proposed topology.

Chapter 5. Conclusions and Future Work

A high frequency resonant SEPIC DC-DC converter is proposed, analyzed, simulated, and an experimental prototype is built and tested. The prototype achieved an experimental efficiency that is greater than 81%, which is considered very promising.

As mentioned previously, there is still room to improve on this design. A great deal of improvement was gained by improving the quality of the inductors. Optimizing magnetic design even further can lead to even more improvements on the efficiency front. The procedure for determining the output and input voltages and the power rating was just used as a starting point. More work needs to be done to find the optimum voltage and power levels for this topology. Even the preliminary work done in the component variation section shows that an efficiency improvement can be gained by adjusting the component values and power level. The power level will depend on application and equalization architecture that is used. A resonant gate drive should also be developed to help decrease size and increase efficiency.

An accurate and simple electrical circuit battery is also proposed and experimentally verified. The proposed model is simpler and easier to generate than the common models used today. The performance of the proposed model exceeds that of the common models.

Work needs to be done to further verify the performance and improve the model. Other load profiles should be tested to see if the results are consistent. Other batteries should also be tested. Expanding the component count should also be explored to see if accuracy is gained. The tradeoff between accuracy and complexity would need to be closely considered

in this case. Battery modeling should then be applied for use in a battery management system, which is the end goal of research such as this.

References

- [1] H. Qian, J. Zhang, J. Lai, and W. Yu, "A High-Efficiency Grid-Tie Battery Energy Storage System," *IEEE Transactions on Power Electronics*, vol. 26, no. 3, pp. 886-896, March 2011.
- [2] Press release, <http://www.whitehouse.gov/the-press-office/2012/08/28/obama-administration-finalizes-historic-545-mpg-fuel-efficiency-standard>.
- [3] "Li-Ion vs. Lead Acid," ultralifecorporation.com/download/275/.
- [4] a123systems.com
- [5] L.W. Juang, P.J. Kollmeyer, T.M. Jahns, and R.D. Lorenz, "Improved Nonlinear Model for Electrode Voltage –Current Relationship for Consistent Online Battery System Identification," *IEEE Energy Conversion Congress and Exposition (ECCE)*, pp. 2628-2634, September 2011.
- [6] D.J. Perreault, J.Hu, J.M. Rivas, Y. Han, O. Leitermann, R.C.N. Pilawa-Podgurski, A.Sagneri, and C.R. Sullivan, "Opportunities and Challenges in Very High Frequency Power Conversion," *IEEE Applied Power Electronics Conference and Exposition (APEC)*, pp. 1-14, February 2009.
- [7] Y. Han, G. Cheung, A. Li, C.R. Sullivan, and D.J. Perreault, "Evaluation of Magnetic Materials for Very High Frequency Power Applications," *IEEE Transactions on Power Electronics*, vol. 27, no. 1, pp. 425-435, January 2012.

- [8] J.R. Warren III, K.A. Rosowski, and D.J. Perreault, "Transistor Selection and Design of a VHF DC-DC Power Converter," *IEEE Transactions on Power Electronics*, vol. 23, no. 1, January 2008.
- [9] W. Bowman, J. Balicki, F. Dickens, R. Honeycut, W. Nitz, W. Strauss, W. Suiter, and N. Zeisse, "A resonant dc-to-dc converter operating at 22 Megahertz," *IEEE Applied Power Electronics Conference and Exposition (APEC)*, pp. 3-11. February 1988.
- [10] J.M. Rivas, J. Shafran, R.S. Wahby, and D.J. Perreault, "New architectures for radio-frequency dc/dc power conversion," *IEEE Transactions on Power Electronics*, vol. 21, no. 2, pp. 380-303, March 2006.
- [11] J.R. Warren III, "Cell modulated dc/dc converter," M.Eng. Thesis, Department of Electrical Engineering and Computer Science, Massachusetts Institute of Technology, Cambridge, May 2005.
- [12] J. Cao, N. Schofield, and A. Emadi, "Battery Balancing Methods: A Comprehensive Review," *IEEE Vehicle Power and Propulsion Conference (VPPC)*, pp. 1-6, September 2008.
- [13] P. Cassani and S. Williamson, "Significance of battery cell equalization and monitoring for practical commercialization of plug-in hybrid electric vehicles," *IEEE Applied Power Electronics Conference*, pp. 465-471, 2009.
- [14] C. Kim, M. Kim, D. Cho, and G. Moon, "A modularized charge equalizer using battery monitoring for series connected Li-Ion battery strings in an electric vehicle,"

- IEEE* Energy Conversion Congress and Exposition (ECCE), pp. 3923-3928, September 2010.
- [15] J.W. Kimball, B.T. Kuhn, P.T. Krein, "Increased performance of battery packs by active equalization," *IEEE* Vehicle Power and Propulsion Conference, pp. 323-372, 2007.
- [16] J. Schonberger, "Modeling of a Lithium Ion Cell using PLECS," www.plexim.com/files/plecs_lithium_ion_pdf.
- [17] M. Chen and G.A. Rincon-Mora, "Accurate Electrical Battery Model Capable of Predicting Runtime and *I-V* Performance," *IEEE* Transactions on Energy Conversion, vol. 21, no. 2, pp. 504-511, June 2006.
- [18] L.P. Mandal and R.W. Cox, "A Transient-Based Approach for Estimating the Electrical Parameters of a Lithium-ion Battery Model," Energy Conversion Congress and Exposition (ECCE), pp. 2635-2640, September 2011.
- [19] M. Einhorn, V. Conte, C. Kral, and J. Fleig, "Comparison of Electrical Battery Models using a Numerically Optimization Parameterization Method," Vehicle Power and Propulsion Conference (VPPC), pp. 1-7, September 2011.
- [20] L. Gao, S. Liu, and R. Dougal, "Dynamic Lithium-Ion Battery Model for System Simulation," *IEEE* Transactions on Components and Packaging Technologies, vol. 25, no. 3, pp. 495-505, September 2002.

- [21] M. Durr, A. Cruden, S. Gair, and J.R. McDonald, "Dynamic model of a lead acid battery for use in a domestic fuel system," *Journal of Power Sources*, vol. 161, pp. 1400-1411, February 2006.
- [22] G. Plett, "Extended Kalman filtering for battery management systems of LiPB-based HEV battery packs: Part 2. Modeling and Identification," *Journal of Power Sources*, vol. 134, pp. 262-276, May 2004.
- [23] H. He, R. Xiong, and J. Fan, "Evaluation of Lithium-Ion Battery Equivalent Circuit Models for State of Charge Estimation by an Experimental Approach," *energies*, vol. 4, pp. 582-598, March 2011.
- [24] Y. Hu and S. Yurkovich, "Linear parameter varying battery model identification using subspace methods," *Journal of Power Sources*, vol. 196, pp. 2913-2923, 2011.
- [25] P. Spagnol, S. Rossi, and S.M. Savaresi, "Kalman Filter SoC estimation for Li-Ion batteries," *IEEE International Conference on Control Applications (CCA)*, 2011.
- [26] X. Tang, X. Mao, J. Lin, and B. Koch, "Li-ion Battery Parameter Estimation for State of Charge," *IEEE American Control Conference*, 2011.
- [27] M. Marinelli, A. Morini, and F. Silvestro, "Modeling of DFIG Wind Turbine and Lithium Ion Energy Storage System," *Complexity in Engineering*, pp. 43-45, February 2010.

- [28] S.X. Chen, K.J. Tseng, and S.S. Choi, "Modeling of Lithium-Ion Battery for Energy Storage System Simulation," Power and Energy Engineering Conference, pp. 1-4, March 2009.
- [29] C.R. Gould, C.M. Bingham, D.A. Stone, and P. Bentley, "New Battery Model and State-of-Health Determination Through Subspace Parameter Estimation and State-Observe Techniques," *IEEE Transactions on Vehicular Technology*, vol. 58, no.8, pp. 3905-3916, October 2009.
- [30] M. Ceraolo, "New Dynamical Models of Lead-Acid Batteries," *IEEE Transactions on Power Systems*, vol. 15, no. 4, pp. 1184-1190, November 2000.
- [31] X. Hu, F. Sin, Y. Zou, and H. Peng, "Online Estimation of an Electric Vehicle Lithium-Ion Battery Using Recursive Least Squares with Forgetting," American Control Conference, pp. 935-940, July 2011.
- [32] L. Chenglin, T. Zining, and W. Lifang, "SOC Estimation of LiFeO₄ Battery Energy Storage System," Power and Energy Engineering Conference, pp. 1-4, March 2010.
- [33] M. Ceraolo and G. Pede, "Techniques for Estimating the Residual Range of an Electric Vehicle," *IEEE Transactions on Vehicular Technology*, vol. 50, no. 1, pp. 109-115, January 2001.
- [34] V.H. Johnson, A.A. Pesaran, and T. Sack, "Temperature-Dependent Battery Models for High-Power Lithium-Ion Batteries," 17th Annual Electric Vehicle Symposium, Montreal Canada, October 2000.

- [35] V.H. Johnson, "Battery performance moles in ADVISOR," *Journal of Power Sources*, vol. 110, pp. 321-329, 2002.
- [36] J.F. Manwell and J.G. McGowan, "Lead Acid Battery Storage Model for Hybrid Energy Systems," *Solar Energy*, vol. 50, no. 5, pp. 399-405, 1993.
- [37] <http://www.a123systems.com/lithium-ion-cells-26650-cylindrical-cell.htm>
- [38] L.W. Juang, P.J. Kollmeyer, T.M. Jahns, R.D. Lorenz, "System Identification- Based Lead-Acid Battery Online Monitoring System for Electric Vehicles," *Energy Conversion Congress and Exposition (ECCE)*, pp. 3903-3910, September 2010.
- [39] <http://impedance0.tripod.com>.
- [40] P. Cassani, and S. Williamson, "Significance of battery cell equalization and monitoring for practicalcommercialization of plug-in hybrid electric vehicles," in *IEEE 2009 Applied Power Electron. Conf.*, 2009, pp.465-471.
- [41] C. Kim, M. Kim, D. Cho, and G. Moon, " A modularized charge equalizer using battery monitoring ic for series connected Li-Ion battery strings in an electric vehicle", *IEEE Energy Conversion Congress and Exposition (ECCE)*, pp. 3923-3928, Sept. 2010.
- [42] J. W. Kimball, B. T. Kuhn, P. T. Krein, "Increased performance of battery packs by active equalization", *IEEE Vehicle Power and Propulsion Conference*, 2007, pp. 323-327.

- [43] J. Hu, A.D. Sagneri, J.M. Rivas, Y. Han, S.M. Davis, and D.J. Perreault, "High-Frequency Resonant SEPIC Converter With Wide Input and Output Voltage Ranges," *IEEE Transactions on Power Electronics*, vol. 27, no. 1, pp. 189-200, January 2012.
- [44] N.O. Sokal and A.D. Sokal, "Class E – A New Class of High Efficiency Tuned Single-Ended Switching Power Amplifiers," *IEEE Journal of Solid-State Circuits*, vol. SC-10, no. 3, pp. 168-176, June 1975.
- [45] N.O. Sokal, "Class-E RF Power Amplifiers," *QEX*, pp. 9-20, February 2001.
- [46] B. Dong, and Y. Han, "A new architecture for battery charge equalization," *IEEE Energy Conversion Congress and Exposition*, pp. 928-934, Sept. 2011.
- [47] <http://www.coilcraft.com/misc/1515sqd.html>
- [48] <http://www.st.com/web/en/resource/technical/document/datasheet/CD00002924.pdf>
- [49] http://www.digikey.com/Web%20Export/Supplier%20Content/Vishay_8026PDF/VishaySiliconix_PadPatterns.pdf?redirected=1
- [50] <http://www.fairchildsemi.com/ds/FD/FDS86106.pdf>
- [51] <http://www.atceramics.com/UserFiles/700b.pdf>
- [52] <http://www.silabs.com/Support%20Documents/TechnicalDocs/Si823x.pdf>
- [53] http://www.micrometals.com/catalog_index.html
- [54] <http://www.cmi-ferrite.com/Materials/Datasheets/NiZn/N40.pdf>

- [55] Y. Han, ECE 712 Course Material, University of Wisconsin- Madison.