THE MODIFIED-MULTIPHASE BOOST CONVERTER: COMBINED
INDUCTORS AND CAPACITORS TOPOLOGY

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ABSTRACT

The Modified-Multiphase Boost Converter: Combined Inductors and Capacitors Topology

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In this work, a modified boost converter design has been implemented in a multiphase configuration with a condensed topology. The modified aspect of the design has already been proven to drastically reduce input current ripple by about 40% in a single-phase implementation. By placing two modified boost converters in parallel with interleaving main switches (multiphase), the input inductors and modified capacitors of the modified topology can be reduced to just one of each, lowering the number of components, size, and cost. Additionally, multiphase DC/DC converters lower input/output voltage and current ripples while delivering more power compared to single-phase converters. By combining the modified and multiphase benefits, this thesis creates a topology with low ripple and noise while providing high power capability. This thesis covers the analysis, simulation, hardware implementation, and testing of the Modified-Multiphase Boost Converter as well as an equivalent Standard-Multiphase Boost Converter for comparison. Simulation and hardware test results exhibited a 9% input current ripple reduction with the Modified-Multiphase design, presenting a high-power converter with considerable input noise reduction.

Keywords: dc/dc, boost, multiphase, interleave, modified, power, ripple

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1. INTRODUCTION

Across the entire world, electric power has become an essential part of modern life. The power and energy industry continues to grow every day as processes are improving and new power generation, conversion, and control methods are being developed. From generation to distribution to conversion, electric power relies on control to safely and reliably provide power. This need for control is fulfilled by the area of electrical engineering known as “Power Electronics.” The primary goal of power electronics is to efficiently and inexpensively control the flow of electric energy by using solid-state switches and other electronic devices. Power electronics surrounds us. Greater than 40% of the electric power generated worldwide utilizes power electronics systems. Some examples of these systems include consumer electronics, power supplies, renewable energy, data centers, transportation, and power converters [1].

The design of power converters utilizes power electronic design. Power converters adjust the flow of energy and vary voltage levels in Alternating Current (AC) and Direct Current (DC) systems. Rectifiers (AC/DC) and inverters (DC/AC) are the two types of converters used to transfer energy from one type of electric power (AC or DC) to the other (DC or AC) using power semiconductor switches. AC/AC converters adjust voltage levels and/or frequency of AC power through phase control and integral cycle control techniques. DC/DC converters convert fixed DC voltage to variable DC voltages. Eliminating the need of “bulky transformers and inefficient linear regulators [2],” DC/DC converters step-up and step-down DC voltages at high efficiency and frequencies leading to smaller systems and reduced costs [1].

Numerous consumer electronics and household devices run on DC power, making DC/DC converters necessary and abundant. DC/DC converters come in two types, isolated and

non-isolated. Isolated DC/DC converters incorporate transformers in their design while non-isolated DC/DC converters do not. Within the non-isolated category, there are several different topologies of converters; some of the most common are buck (step-down) and boost (step-up) converters. Buck converters are crucial since many loads typically operate at a voltage lower than their input voltage. Boost converters are also very useful in the opposite case where source voltage is lower than the load voltage it is supplying.

Due to the increasing popularity of photovoltaic (PV) systems, another type of power converter, the inverter, has also grown more popular. Solar panel installation has been increasing for some time now. Solar offers a renewable, clean energy source, promoting a greener environment. Additionally, residential solar installation continues to increase as prices drop and tax deductions and government rebates allow PV systems to pay for themselves quicker [3].

Figure 1-1 shows the price decrease and installation increase of solar panels over the last few decades [4].

Figure 1-1: Price Decrease and Installation Increase of Solar Panels

The rise of solar panels poses an interesting problem; solar panels produce DC power but the United States (as well as many other countries) run on AC power. Everything from the electrical grid to household appliances run off AC power. Therefore, inverters, which convert DC power to AC power, have become a necessity [5]. Frequently, when solar panels produce excess energy, the extra energy needs to be put back on the electric grid. However, the “grid” consists of transmission and distribution lines which often run at much higher voltages than what solar panels produce. Therefore, the solar panel voltage needs to be stepped up. One approach involves the use of power transformers after the energy has been converted to AC (after the inverter); this proves costly as transformers are bulky, heavy, and expensive. Another approach utilizes a DC/DC boost converter before the energy goes through the inverter. Boost converters have high efficiency and are simpler and smaller. Therefore, boost converters play an important role in photovoltaic systems and renewable energy in general.

Today, the power and energy industry relies increasingly more on DC power and therefore, DC/DC converters. The high efficiency and smaller size of DC/DC converters is especially appealing in consumer hardware and portable electronics as small size with long battery life is often desired. Additionally, applications such as renewable energy systems and electric vehicles are often high-powered and battery-operated [2]. For these type of applications, the multiphase method is commonly used by connecting converter units in parallel. By doing so, the converter system can achieve higher current (leading to higher power) while maintaining high efficiency. The multiphase technique has been particularly useful when the load requires relatively high power while operating at low voltage such as those found in microprocessors [6].

2. BACKGROUND

2.1 Boost Converters

As solar panels have become more popular, so have boost converters. Solar, also known as photovoltaic (PV), energy continues to grow in importance in the renewable energy field. PV cells have a greater upside compared to other forms of renewable energy (wind, hydro, etc.) due to their high efficiency and reliability, capability of delivering power directly, contribution of no land destruction, long life-span, and affordable production and installation costs [7]. Solar panels produce DC power at given ratings based on the panel size; therefore, DC/DC converters are essential to adjust power levels. Boost DC/DC converters prove especially useful as they step up voltages to higher levels, which allows for PV systems to be connected to high power electrical grids [7]. Additionally, boost converters are commonly used due to their low part count, reliability, high efficiency, and inexpensiveness [8]. Figure 2-1 illustrates a standard boost converter and observe its simplicity and low part count.

Figure 2-1: Standard DC/DC Boost Converter Schematic

2.2 Modified Boost Converters

While DC/DC converters provide simple, efficient solutions, they still have drawbacks. The main disadvantage of standard DC/DC converters is switching noise. Switching noise adds AC ripple in the DC input and output signals, negatively affecting maximum power point tracking (MPPT) in solar panel modules. Additionally, switching noise causes electromagnetic interference (EMI) in circuits, which adversely affects system control, performance, and reliability. Many modified DC/DC converters address the issues of reducing switching noise and EMI to improve reliability and efficiency of the circuit [2]. Nathan Lentz, a former Cal Poly Electrical Engineering Master's student, performed his thesis on a modified boost converter topology with the goal of achieving a reduced input current ripple. Figure 2-2 shows the modified boost topology that Lentz tested [2].

Figure 2-2: Modified Boost Converter to Reduce Input Current Ripple

In his report, Lentz shows that the DC transfer function for the modified boost is the same as that of a standard boost (where D is the duty cycle):

$$\frac{V_{OUT}}{V_{IN}} = \frac{1}{(1 - D)} \quad (2 - 1)$$

The transfer function shows that the modified and standard boosts have the same relationship between the duty cycle of the switch and the input/output quantities. That is, by adding the additional capacitor C_1 , the converter's steady state operation remains the same. Circuit analysis shows that the resonant behavior of the modified boost converter determines the minimum C_1 value, while the maximum value is constrained by the output capacitance C_o . A larger C_1 reduces voltage ripple across L_1 leading to smaller voltage changes across the inductor. The smaller voltage changes induce smaller inductor current changes; that is, smaller inductor current ripple. Therefore, since the inductor is connected to the input in a boost converter, smaller inductor current ripple means smaller input current ripple, achieving the primary goal of Lentz's thesis. The small input current ripple demonstrates the modified topology's ability to contain switching noise [2].

The differences in circuit design of Figures 2-1 and 2-2 show another key distinction between the standard and modified boost topologies, the modified boost splits up the inductor into two separate smaller inductors. Due to amp second balance (ASB), the average (dc) current through a capacitor is zero. Therefore, in a dc circuit with low ripple current, the two inductors (L_1 and L_2) are effectively in series in the modified boost [1]. Three main possibilities exist as methods of splitting up the effective series inductance ($L_{\text{series_eff}} = L_1 + L_2$), a 50/50, 25/75, or a 75/25 split. Lentz showed via Matlab simulation that a 50/50 split allows for evenly shared current ripple in each inductor as well as simplifies component selection. Furthermore, simulation and hardware test results showed that the modified boost converter achieved its goal of a 40% reduction in input current ripple as it produced less than 1% total input current ripple [2].

While, Lentz addressed input current ripple, different modified boost converters address other issues. Figure 2-3 shows a “Multistage Boost Converter” that was designed to attain high static gain [9].

Figure 2-3: Multistage Boost Converter

For a high gain in a conventional boost, the duty cycle must be large. A high duty cycle causes elevated stress on components such as the semiconductor switches. The multistage topology of Figure 2-3 cascades three boost converters. In response, the output voltage increases geometrically, leading to a high static gain:

$$\frac{V_{OUT}}{V_{IN}} = \left(\frac{1}{1-D} \right)^3 \quad (2-2)$$

Some drawbacks of cascading boost converters are high switch stress due to large input current and a high component count [9]. These additional components, especially the extra inductors, lead to larger size and higher cost [1]. Therefore, cascading isn't always practical. Figure 2-4 shows another modified boost converter design that achieves high gain but with fewer inductors. This “Hybrid Switched Inductor Boost Converter” was designed to achieve high gain and efficiency, which is beneficial in applications such as solar panel energy conversion systems, fuel cell energy conversion systems, and back-up battery systems for uninterruptable power supplies [7].

Figure 2-4: Hybrid Switched Inductor Boost Converter

The gain of this Hybrid Switched Inductor Boost Converter is higher than the gain of a standard boost converter by a factor of $(D + 1)$, where D is the duty cycle:

$$\frac{V_{OUT}}{V_{IN}} = \frac{(1 + D)}{(1 - D)} \quad (2 - 3)$$

2.3 Multiphase Boost Converters

While some modifications to boost converters can attain advantages such as higher gain, other drawbacks go unaddressed. For example, in high-power applications, boost converters experience many limitations. Some disadvantages include high voltage and current ripples, high switching losses from large switch voltage stress, and large volume and weight due to bigger components [10]. The size of the inductor in a boost converter becomes particularly large in high-power conversion. Additionally, due to the high stress on components from large ripple, problems such as heating (lower efficiency), reduced life, and reliability of components arise [8]. Lastly, in battery operated systems, a large current ripple has harmful effects on battery life [11].

Many of the discussed disadvantages in boost converters are also present in buck converters. That is, high ripples, power dissipation, and efficiency all become problems at higher power levels. However, placing two or more bucks in parallel in a multiphase design offers solution these issues. Figure 2-5 shows a two-phase multiphase buck converter design.

Figure 2-5: Multiphase Buck Topology

The key idea behind multiphase converters is interleaving. Interleaving is the process of phase shifting the converter switching signals to guarantee a uniform distribution over one switching period. For example, in Figure 2-5, S_1 and S_2 are 180 degrees out of phase of each other. For N stages, there will be N primary switches each operating $360/N$ degrees out of phase of each other. Additionally, given one switching period T (equal to $1/f_{sw}$ where f_{sw} is the switching frequency in [Hz]), each of the N switches' pulse width is equal to T/N [8]. Figure 2-6 shows the switching signals for the three primary switches in a three-phase converter [2].

Figure 2-6: Three-Phase Multiphase Interleaved Switching

Interleaved multiphase converters experience a higher effective output ripple frequency (N times the switching frequency f_{sw}) and output current ripple cancellation since the outputs of each phase are connected. Higher ripple frequency and ripple cancellation lead to reduced output filter requirements. Furthermore, a lower current ripple provides for better thermal dissipation and reduced inductor size and switching loss [12]. With a multiphase design, total current is split through each phase, which reduces conduction (I^2R) losses and increases efficiency. This leaves designers with a tradeoff of a higher efficiency with the same rated components as in single-phase, reducing component sizes to lower cost, or a combination of both [13]. Lastly, due to less energy stored in each output inductor in a multiphase buck, the load-transient performance improves.

While buck converters can be greatly improved in high-power applications through a multiphase design, a multiphase boost experiences the same advantages. Figure 2-7 shows a two-phase boost converter design.

Figure 2-7: Multiphase Boost Topology

Like its buck counterpart, a multiphase boost will also incorporate interleaved switching, yielding output current ripple cancellation, higher effective output ripple frequency, and reduced

output filter requirements. In fact, the value of the output capacitance (C_{out}) decreases by a factor of N (the number of phases) [14]. A multiphase boost also experiences the additional benefit of reduced input current ripple [15]. The input and output current ripples minimize at exactly a 50% duty cycle in a two-phase boost design. Thus, a duty cycle less than, but as close as possible to 50% is desirable for two phases [14]. Maximum ripple cancellation in N phases occurs $(N-1)$ times at duty cycle values of [8]:

$$D_{\max ripple\ cancellation} = K \frac{100\%}{N} \quad (2 - 4)$$

$$where\ K = 1, 2, \dots, N - 1$$

A multiphase boost has reduced input and output ripple for both current and voltage; this not only reduces filtering requirements but also lowers harmonics. Figures 2-8 and 2-9 show how ripple current in both the inductor (input) and output respectively decrease with additional phases [16].

Figure 2-8: Multiphase Inductor (Input) Ripple Current vs. Duty Cycle

Figure 2-9: Multiphase Output Ripple Current vs. Duty Cycle

Although using many phases may appear to be beneficial; however, this is not always the case. Each additional phase adds more components and therefore, increases complexity, size, and cost. Furthermore, as seen in Figures 2-8 and 2-9, the ripple decreases less from one phase to the next at higher total phases. That is, the benefit seen from going from one phase to two phases is much greater than the benefit from two phases to three phases [8]. Thus, a multiphase boost improves a standard boost, but the number of supplementary phases needs to be optimized for a given application.

The capability for high power applications is another area where a multiphase boost converter improves upon a standard boost. High power means higher current (at the same voltage level); a multiphase boost shares the high input current equally among each phase [9]. This current sharing allows for better thermal dissipation and reduced passive component sizes (inductors and capacitors), switching loss, and conduction (I^2R) losses. These improvements show that a multiphase boost converter can provide a high-power, high efficiency converter that is both reliable and fast [10].

2.4 Modified – Multiphase Boost Converters

A multiphase boost converter provides a solution that increases efficiency and decreases current and voltage ripples throughout the power stage. Lower ripple greatly improves a circuit's ability to meet Electromagnetic Interference (EMI) requirements; therefore, a multiphase boost converter provides a significant improvement over a standard boost converter with regards to EMI standards. However, the more ripple-free the input, the better the circuit will be able to meet EMI requirements. While input filters can reduce ripple, they add components and increase inductor size, increasing the total overall cost [15]. Therefore, various modified-multiphase boost converters have been investigated to inherently reduce input ripple further while minimizing added costs.

One design option for a modified-multiphase boost converter involves the use of two inductors coupled together. Figure 2-10 shows a two-phase boost converter with coupled inductors and Figure 2-11 shows the equivalent circuit of directly coupled inductors [15].

Figure 2-10: Two Phase Boost Converter with Coupled Inductors

L_{k1} and L_{k2} are leakage inductances and L_m is the mutual inductance. The coupled inductors are related by:

$$L_m = \alpha(L_1 L_2)^{1/2}$$

Figure 2-11: Equivalent Circuit of Directly Coupled Inductors

The coupled inductors technique, often called an integrated magnetic technique, reduces component count as two coupled inductors have the same effect as three inductors as evidenced by Figure 2-11. Coupled inductors also decrease the amount of magnetic core material. A coupled inductor makes use of one magnetic core with multiple windings (one winding for each inductor). Therefore, since the inductor windings all use the same core, the ripple in one winding can be transferred to the other winding. This ripple transfer allows for very small input current ripple in the boost converter application of Figure 2-10, leading to lower harmonics and a better EMI rating. Additionally, using coupled inductors leads to shorter time to stability in the input current of a boost converter, creating lower transients and a fast system [15].

Just as modifications in the single-phase boost case helped improve functionality, the same modifications can also apply to a multiphase boost. Recall the single-phase modified boost design that Lentz tested; it incorporated an additional capacitor that significantly reduced input current ripple (see Figure 2-2) [2]. A two-phase configuration of the modified boost is shown in Figure 2-12 where the additional capacitors from each phase are combined into one capacitor and the first inductor in each phase are combined.

Figure 2-12: Modified-Multiphase Boost Converter

The Modified-Multiphase Boost Converter design incorporates the lower input current ripple of Lentz's boost design, as well the benefits of a multiphase boost topology. Therefore, the objective of this thesis is to implement the boost converter design of Figure 2-12 to demonstrate low current and voltage ripples in both the input and output, high efficiency, fast transient response, and low EMI. This thesis presents an analysis and implementation of the Modified-Multiphase Boost Converter, exhibiting the aforementioned benefits over standard boost and multiphase boost topologies, while providing increased power capability.

3. DESIGN REQUIREMENTS

The Modified-Multiphase Boost Converter aims to provide a low-ripple, efficient, inexpensive, and high-power boost converter solution for applications such as energy conversion for a PV system. The input and output voltages are the first parameters that need to be determined as they affect the sizing of components and functionality of the circuit. Values of 12V and 19V were chosen for the input and output voltages respectively; many battery systems run at 12V, and larger electronics such as laptops run off 19V power supplies. Also, these voltage values were chosen to yield a duty cycle relatively close to 50% to allow for maximum ripple cancellation due to the circuit's interleaving multiphase behavior [14]. For a boost converter, the duty cycle is as follows [1]:

$$Duty\ Cycle = D = 1 - \frac{V_{in}}{V_{out}} = 1 - \frac{12V}{19V} = 0.368 = 36.8\% \quad (3 - 1)$$

Additionally, this thesis aims to develop a solution for high-power boost converters; an output power value of 95W (corresponding to 5A output current) was selected. Figure 3-1 shows the “Level 0” block diagram of the converter with the selected input and output voltages.

Figure 3-1: Level 0 Modified-Multiphase Boost Converter Diagram

In boost converter topology, the inductor is connected to the input of the circuit; therefore, the input current ripple relates to the total inductor current in a multiphase design. With standard boost converters, designers frequently target an input current ripple less than 40%. However, Lentz showed that the modified boost topology could achieve an input current ripple of around 10% in hardware tests and 1% in simulation results [2]. Consequently, an input current ripple of around 10% provides a reasonable target for the modified-multiphase boost converter.

Another important design parameter in boost converters is output voltage ripple. Many devices that run off DC power expect a clean (low ripple) DC output since a large ripple can adversely harm equipment. Due to this low ripple necessity, and since multiphase converters have lower output current ripple, a low output voltage ripple of around 2%, corresponding to 380mV (2% of 19V = 380mV), proves to be a decent goal.

Another key parameter, switching frequency, greatly affects component values. A larger switching frequency reduces the component size of inductors and capacitors, which benefits designs by having smaller size and lower cost. As seen by the “Level 1” diagram of Figure 3-2, the modified-multiphase boost converter already has a high component count compared to standard boost converters; it has additional inductor and capacitor due to it being “modified,” and it has an additional inductor, switch, and diode from its multiphase aspect. Thus, reducing size and cost would be beneficial. However, higher-frequency devices, particularly in PWM (pulse-width modulation) switching based converters, have lower efficiency, creating a tradeoff between size/cost and efficiency [17]. A target switching frequency range of 100-200kHz and an efficiency goal of at least 90% at full-load (95W) allows for flexibility in the design with respect to this tradeoff.

Figure 3-2: Level 1 Modified-Multiphase Boost Converter Diagram

Devices that run off DC power are usually rated for specific DC voltage values. Therefore, the output voltage stationarity of a boost converter proves extremely important since large changes on the output could harm the load devices. However, the source input voltage and load output power often experience fluctuations in many applications. For example, one application could contain a boost converter connected to a PV system on the input, and an off-grid DC house on the output. During a cloudy day, the input voltage supplied by the PV system would likely drop from its usual sunny-day values. Also, during night time, the DC house would demand relatively low power than during the day. Even with these fluctuations, the boost converter would still need to supply close to its rated output voltage, so it won't harm the devices or circuits in the DC house (load). The capability of a converter to maintain its output voltage while it experiences varying input voltage and output power levels is known as line regulation and load regulation respectively. For the modified-multiphase boost converter design, a percent line regulation of 10% and a percent load regulation of less than 5% were selected. Equations 3-2 and 3-3 show how line and load regulation are calculated [1]. Lastly, Table 3-1 displays all the design requirements of the modified-multiphase boost converter.

$$\%Line\ Regulation = \frac{V_{out(high-input)} - V_{out(low-input)}}{V_{out(nominal-input)}} \times 100\% \quad (3 - 2)$$

$$\%Load\ Regulation = \frac{V_{out(low-load)} - V_{out(high-load)}}{V_{out(high-load)}} \times 100\% \quad (3 - 3)$$

Table 3-1: Modified-Multiphase Boost Converter Design Requirements

Parameter	Abbreviation	Value
Input Voltage	V_{in}	12 V
Input/Inductor Current Ripple	$\%\Delta I_{in}$	~ 10%
Output Voltage	V_{out}	19 V
Output Current	I_{out}	5 A
Output Power	P_{out}	95 W
Efficiency	η	> 90% (full load)
Switching Frequency	f_{sw}	100-200 kHz
Output Voltage Ripple	$\%\Delta V_{out}$	~ 2% (380 mV)
Percent Line Regulation	$\%Reg_{line}$	~ 10% (input)
Percent Load Regulation	$\%Reg_{load}$	< 5% (output)

4. DESIGN AND SIMULATION RESULTS

Standard boost converters experience large current and voltage ripple at high power levels and therefore, have lower efficiency [10]. The Modified Boost Converter addresses these problems by drastically reducing the (input) inductor current ripple [2]. Additionally, two standard boost converters in a parallel-multiphase configuration remedy the problems of large input current and output voltage ripples at high power levels [10]. The Modified-Multiphase Boost Converter improves upon the problems in standard boost converters by combining the solutions in Lentz's Modified Boost Converter and a multiphase converter. By decreasing both input and output current and voltage ripples at high power levels, the Modified-Multiphase Boost Converter offers an efficient, low noise, and high-power solution.

4.1 State Space Analysis

Designing the Modified-Multiphase Boost Converter requires state space analysis. State space analysis completely represents a n^{th} -order (n = any real, positive number), dynamic system by its state variables in a set of first-order ($n = 1$) differential equations known as the state equation. Averaging the state space representations during the different Modified-Multiphase Boost Converter's states generates the converter's DC transfer function.

State variables describe a system's dynamic behavior. In a DC/DC converter, the energy storage components, inductors and capacitors, determine its dynamic behavior. Therefore, the use of the standard capacitor and inductor voltage/current relationship equations (4-1 and 4-2 respectively) make capacitor voltage and inductor current the most logical choices for state variables.

$$i_c = C \frac{dv_c}{dt} \quad (4 - 1)$$

$$v_L = L \frac{di_L}{dt} \quad (4 - 2)$$

Analyzing the Modified-Multiphase Boost Converter in its three different switching states yields three state equations. Averaging the three state equations produces the converter's DC transfer function. Obtaining the state equations involves the use of circuit analysis to each switch state. Kirchhoff's Voltage Law (KVL) represents inductor current in terms of capacitor voltage state variables and Kirchhoff's Current Law (KCL) represents capacitor voltage in terms of inductor current state variables.

4.1.1 Standard Boost Converter

Nathan Lentz performed the state-space analysis of a standard boost converter and of his Modified Boost Converter in his thesis. Lentz derived the standard boost transfer functions and showed how his Modified Boost achieves identical transfer functions (4-3 and 4-4, D = duty cycle) [1] [2].

$$\frac{V_{out}}{V_{in}} = \frac{1}{(1 - D)} \quad (4 - 3)$$

$$\frac{I_{out}}{I_{in}} = (1 - D) \quad (4 - 4)$$

4.1.2 Modified-Multiphase Boost Converter

Due to the additional capacitors and inductors, the state space analysis of the Modified-Multiphase Boost Converter requires more state variables than the state space analysis of a

standard boost, single-phase modified boost, or standard multiphase boost. Figure 4-1 shows the Modified-Multiphase Boost Converter with labeled capacitor and inductor voltages and currents for state space analysis.

Figure 4-1: Modified-Multiphase Boost Converter Labeled Schematic

The switching duty cycle cannot be greater than 50% to inhibit switch overlap; therefore, S_1 and S_2 will never be on at the same time. As in standard DC/DC converters, a given switch won't conduct at the same time as its corresponding diode; that is, S_1 and D_1 will never be on at the same time (the same goes for S_2 and D_2). Consequently, the Modified-Multiphase Boost Converter operates in three different states. Figure 4-2 shows the three different states and how S_1 , S_2 , D_1 , and D_2 determine each state.

Figure 4-2: Converter States Based on Switches and Diodes

As seen in Figure 4-2, the Modified-Multiphase Boost Converter operates in State 1 when S_1 and D_2 conduct (on), and S_2 and D_1 are off. Figure 4-3 depicts the converter in State 1 by modeling conducting switches and diodes as electric shorts and non-conducting switches and diodes as opens.

Figure 4-3: State 1 Modified-Multiphase Boost Converter

KVL around the loop through from V_{in} to C_{m1} and around the loop from L_2 to V_{out} provide the L_1 and L_2 current differential equations respectively (Equations 4-5 and 4-6).

$$v_{L1} = -v_{Cm1} - v_{out} + v_{in}$$

$$\frac{di_{L1}}{dt} = \frac{-1}{L_1} v_{Cm1} - \frac{1}{L_1} v_{out} + \frac{1}{L_1} v_{in} \quad (4-5)$$

$$v_{L2} = v_{Cm1} + v_{out}$$

$$\frac{di_{L2}}{dt} = \frac{1}{L_2} v_{Cm1} + \frac{1}{L_2} v_{out} \quad (4-6)$$

Observing that L_3 is in parallel with C_{m1} yields the L_3 current differential equation (4-7). Additionally, KCL at the nodes where the three inductors meet and at the output provide the capacitor voltage differential equations (4-8 and 4-9).

$$v_{L3} = v_{Cm1}$$

$$\frac{di_{L3}}{dt} = \frac{1}{L_3} v_{Cm1} \quad (4-7)$$

$$i_{Cm1} = i_{L1} - i_{L2} - i_{L3}$$

$$\frac{dv_{Cm1}}{dt} = \frac{1}{C_{m1}} i_{L1} - \frac{1}{C_{m1}} i_{L2} - \frac{1}{C_{m1}} i_{L3} \quad (4-8)$$

$$i_{Cout} = i_{L3} + i_{Cm1} - i_{out} = i_{L1} - i_{L2} - i_{out}$$

$$\frac{dv_{out}}{dt} = \frac{1}{C_{out}} i_{L1} - \frac{1}{C_{out}} i_{L2} - \frac{1}{C_{out} R_L} v_{out} \quad (4-9)$$

The state equation contains the state variable differential equations in matrix form. For State 1, the state equation can be seen in Equation 4-10.

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{Cm1}/dt \\ dv_{out}/dt \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & 1/L_2 \\ 0 & 0 & 0 & 1/L_3 & 0 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & -1/C_{out} & 0 & 0 & -1/C_{out} R_L \end{bmatrix} \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} V_{in} \quad (4-10)$$

As observed in Figure 4-2, the Modified-Multiphase Boost Converter operates in State 2 when S_2 and D_1 conduct (on), and S_1 and D_2 are off. Figure 4-4 depicts the converter in State 2 under the same assumptions as before.

Figure 4-4: State 2 Modified-Multiphase Boost Converter

KVL around the loop through V_{in} to C_{m1} determines the L_1 current differential equation (4-11). Observing that L_2 is in parallel with C_{m1} provides the L_2 current differential equation (4-12). Additionally, KVL around the loop from L_3 to V_{out} yields the L_3 current differential equation (4-13).

$$v_{L1} = -v_{Cm1} - v_{out} + v_{in}$$

$$\frac{di_{L1}}{dt} = \frac{-1}{L_1} v_{Cm1} - \frac{1}{L_1} v_{out} + \frac{1}{L_1} v_{in} \quad (4-11)$$

$$v_{L2} = v_{Cm1}$$

$$\frac{di_{L2}}{dt} = \frac{1}{L_2} v_{Cm1} \quad (4-12)$$

$$v_{L3} = v_{Cm1} + v_{out}$$

$$\frac{di_{L3}}{dt} = \frac{1}{L_3} v_{Cm1} + \frac{1}{L_3} v_{out} \quad (4-13)$$

KCL at the nodes where the three inductors meet and at the output determine the capacitor voltage differential equations (4-14 and 4-15).

$$i_{Cm1} = i_{L1} - i_{L2} - i_{L3}$$

$$\frac{dv_{Cm1}}{dt} = \frac{1}{C_{m1}} i_{L1} - \frac{1}{C_{m1}} i_{L2} - \frac{1}{C_{m1}} i_{L3} \quad (4-14)$$

$$i_{Cout} = i_{L2} + i_{Cm1} - i_{out} = i_{L1} - i_{L3} - i_{out}$$

$$\frac{dv_{out}}{dt} = \frac{1}{C_{out}} i_{L1} - \frac{1}{C_{out}} i_{L3} - \frac{1}{C_{out} R_L} v_{out} \quad (4-15)$$

For State 2, the state equation can be seen in Equation 4-16.

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{Cm1}/dt \\ dv_{out}/dt \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & 0 \\ 0 & 0 & 0 & 1/L_3 & 1/L_3 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & 0 & -1/C_{out} & 0 & -1/C_{out}R_L \end{bmatrix} \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} V_{in} \quad (4-16)$$

As seen in Figure 4-2, the Modified-Multiphase Boost Converter operates in State 3 when D_1 and D_2 conduct (on), and S_1 and S_2 are off. Figure 4-5 depicts the converter in State 3 under the same assumptions as before.

Figure 4-5: State 3 Modified-Multiphase Boost Converter

KVL around the loop through V_{in} to C_{m1} provides the L_1 current differential equation (4-17). Observing that L_2 and L_3 are in parallel with C_{m1} determine the L_2 and L_3 current differential equations (4-18 and 4-19).

$$v_{L1} = -v_{Cm1} - v_{out} + v_{in}$$

$$\frac{di_{L1}}{dt} = \frac{-1}{L_1} v_{Cm1} - \frac{1}{L_1} v_{out} + \frac{1}{L_1} v_{in} \quad (4-17)$$

$$v_{L2} = v_{Cm1}$$

$$\frac{di_{L2}}{dt} = \frac{1}{L_2} v_{Cm1} \quad (4-18)$$

$$v_{L3} = v_{Cm1}$$

$$\frac{di_{L3}}{dt} = \frac{1}{L_3} v_{Cm1} \quad (4-19)$$

KCL at the nodes where the three inductors meet and at the output provide the capacitor voltage differential equations (4-20 and 4-21).

$$i_{Cm1} = i_{L1} - i_{L2} - i_{L3}$$

$$\frac{dv_{Cm1}}{dt} = \frac{1}{C_{m1}} i_{L1} - \frac{1}{C_{m1}} i_{L2} - \frac{1}{C_{m1}} i_{L3} \quad (4-20)$$

$$i_{Cout} = i_{L2} + i_{L3} + i_{Cm1} - i_{out} = i_{L1} - i_{out}$$

$$\frac{dv_{out}}{dt} = \frac{1}{C_{out}} i_{L1} - \frac{1}{C_{out} R_L} v_{out} \quad (4-21)$$

For State 3, the state equation can be seen in Equation 4-22.

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{Cm1}/dt \\ dv_{out}/dt \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & 0 \\ 0 & 0 & 0 & 1/L_3 & 0 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & 0 & 0 & 0 & -1/C_{out} R_L \end{bmatrix} \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} V_{in} \quad (4-22)$$

Summing the three state equations weighted by their respective time durations in (terms of duty cycle) provides the average state equation (4-23).

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{Cm1}/dt \\ dv_{out}/dt \end{bmatrix} = D A_1 \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + D A_2 \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + (1 - 2D) A_3 \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + (D + D + 1 - 2D) \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} V_{in}$$

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{Cm1}/dt \\ dv_{out}/dt \end{bmatrix} = (DA_1 + DA_2 + (1 - 2D) A_3) \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{Cm1} \\ v_{out} \end{bmatrix} + \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} V_{in}$$

$$A_1 = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & 1/L_2 \\ 0 & 0 & 0 & 1/L_3 & 0 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & -1/C_{out} & 0 & 0 & -1/C_{out}R_L \end{bmatrix}$$

$$A_2 = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & 0 \\ 0 & 0 & 0 & 1/L_3 & 1/L_3 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & 0 & -1/C_{out} & 0 & -1/C_{out}R_L \end{bmatrix}$$

$$A_3 = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & 0 \\ 0 & 0 & 0 & 1/L_3 & 0 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & 0 & 0 & 0 & -1/C_{out}R_L \end{bmatrix}$$

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{cm1}/dt \\ dv_{out}/dt \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & -1/L_1 & -1/L_1 \\ 0 & 0 & 0 & 1/L_2 & D/L_2 \\ 0 & 0 & 0 & 1/L_3 & D/L_3 \\ 1/C_{m1} & -1/C_{m1} & -1/C_{m1} & 0 & 0 \\ 1/C_{out} & -D/C_{out} & -D/C_{out} & 0 & -1/C_{out}R_L \end{bmatrix} \begin{bmatrix} i_{L1} \\ i_{L2} \\ i_{L3} \\ v_{cm1} \\ v_{out} \end{bmatrix} + \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} V_{in} \quad (4-23)$$

The DC transfer functions of a converter describe its relationship between the input and output DC values. Under these DC, steady-state conditions, the differential equations approach zero (Equation 4-24).

$$\begin{bmatrix} di_{L1}/dt \\ di_{L2}/dt \\ di_{L3}/dt \\ dv_{cm1}/dt \\ dv_{out}/dt \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} \quad (4-24)$$

Combining Equations 4-23 and 4-24 determines the DC transfer functions. The two differential equations pertaining to the inductor currents in L_1 and L_2 provide important voltage relations (4-25 and 4-26) that lead to the derivation of the voltage transfer function (4-27).

$$\frac{di_{L1}}{dt} = 0 = \frac{-1}{L_1}V_{cm1} - \frac{1}{L_1}V_{out} + \frac{1}{L_1}V_{in}$$

$$V_{in} = V_{cm1} + V_{out} \quad (4-25)$$

$$\frac{di_{L2}}{dt} = 0 = \frac{1}{L_2}V_{cm1} + \frac{D}{L_2}V_{out}$$

$$V_{cm1} = -D(V_{out}) \quad (4-26)$$

$$\frac{V_{out}}{V_{in}} = \frac{1}{(1-D)} \quad (4-27) \text{ same as 4-3}$$

The two bottom differential equations that relate to capacitances C_{m1} and C_{out} provide two important current relations (4-28 and 4-29) that lead to the derivation of the current transfer function (4-30). Note that we already know from the circuit schematics in Figures 4-1, 4-3, 4-4, and 4-5 that $I_{IN} = I_{L1}$.

$$\frac{dv_{cm1}}{dt} = 0 = \frac{1}{C_{m1}} I_{L1} - \frac{1}{C_{m1}} I_{L2} - \frac{1}{C_{m1}} I_{L3}$$

$$I_{L1} = I_{L2} + I_{L3} = I_{in} \quad (4-28)$$

$$\frac{dv_{out}}{dt} = 0 = \frac{1}{C_{out}} I_{L1} - \frac{D}{C_{out}} I_{L2} - \frac{D}{C_{out}} I_{L3} - \frac{1}{C_{out} R_L} V_{out}$$

$$I_{L1} - D(I_{L2} + I_{L3}) = \frac{V_{out}}{R_L} = I_{out} \quad (4-29)$$

$$I_{in} - D(I_{in}) = I_{out}$$

$$\frac{I_{out}}{I_{in}} = (1 - D) \quad (4-30) \text{ same as 4-4}$$

Observe that state space analysis and averaging of differential equations derive the same DC transfer functions (4-27 and 4-30) presented in Section 4.1.1 (4-3 and 4-3 respectively). That is, the Modified-Multiphase Boost Converter operates under the exact same DC transfer functions as that of Lentz's modified boost and a standard boost converter. All three converters experience the same relationships between inputs, outputs, and duty cycle.

4.2 Modified-Multiphase Boost Design Process

The Modified-Multiphase Boost Converter's more complex topology leads to a more complicated design. The sizing of components depends heavily on the design requirements discussed in Chapter 3. Components such as inductors and capacitors directly affect converter

performance such as ripple and efficiency. Furthermore, with additional inductors and capacitors, issues such as resonant frequencies arise in the Modified-Multiphase design.

4.2.1 Resonant Frequencies

The configuration of the Modified-Multiphase Boost Converter complicates the analysis of its functionality. As discussed and seen in Lentz's thesis, the modifications of the converter create resonant frequencies. The Modified-Multiphase Converter in this thesis shares the same resonant frequency behavior as the modified converter in Lentz's thesis (detailed analysis and graphs can be seen in Lentz's thesis). Therefore, selecting a switching frequency (f_{sw}) of 200kHz increases comparability between the two designs. Analysis of this resonant behavior shows that the value of the additional capacitor (C_{m1}) should be greater than 1 μ F to inhibit the converter's maximum resonant frequency from exceeding the 200kHz switching frequency [2].

4.2.2 Controller Selection

The choice of a switching controller stands as one of the first steps in a DC/DC converter design. The Modified-Multiphase Boost Converter required a single controller that could handle a modified design due to the importance of the interleaving functionality between the two main switches. Additionally, the controller needed to meet the design requirements (Chapter 3) of this converter design so it could handle the voltage, current, and power levels.

Linear Technology (LT) was selected as a top choice for a company for controller selection due to the ease of simulation of their components in LTSpice. Research yielded two potential LT controllers that minimized controller pin count and size, and that could meet all the necessary design requirements. The LTC3787 showed potential as a new LT product for

multiphase boost designs. However, the LTC3787 design featured the current sense resistors in series with the inductors. Due to the modifications in this thesis's design with splitting of the inductors and additional capacitor, sense resistors in series with the inductors could lead to potential behavioral problems. Therefore, the selection of the similar LT3782A multiphase boost controller proved advantageous as its design called for current sense resistors in series with the switches, separated from the modifications of this thesis's design. Table 4-1 describes each LT3782A pin's function [18].

Table 4-1: LT3782A Pin Descriptions

Pin Name	Pin Number	Description
SGATE2	1	2 nd Phase Synchronous Drive Signal. An external driver buffer is needed to drive the top synchronous power FET.
SGATE1	2	1 st Phase Synchronous Drive Signal. An external driver buffer is needed to drive the top synchronous power FET.
GND	4, Exposed Pad	Ground. Solder exposed pad to PCB ground plane for rated thermal performance. The exposed pad should be connected to the GND pin as close to the IC as possible.
SYNC	5	Synchronization Input. The pulse width can range from 10% to 70%. Note that the operating frequency is half of the sync frequency.
DELAY	6	When synchronous drivers are used, the programmable delay that delays BGATE turns on after SGATE turns off.
DCL	7	Programs the limit of the maximum duty cycle. When connected to R _{SET} , it operates at natural maximum duty cycle, approximately 90%.
SENSE1 ⁺	8	1 st Phase Current Sense Amplifier Positive Input. An RC filter is required across the current sense resistor. Current limit threshold is set at 63mV.
SENSE1 ⁻	9	1 st Phase Current Sense Amplifier Negative Input.
SLOPE	10	A resistor from SLOPE to GND increases the internal current mode PWM slope compensation.
R _{SET}	11	A resistor from R _{SET} to GND sets the oscillator charging current and the operating frequency.
SENSE2 ⁻	12	2 nd Phase Current Sense Amplifier Negative Input.

SENSE2 ⁺	13	2 nd Phase Current Sense Amplifier Positive Input. An RC filter is required across the current sense resistor. Current limit threshold is set at 63mV.
SS	14	Soft-Start. A capacitor on this pin sets the output ramp up rate. The typical time for SS to reach the programmed level is $(C \cdot 2.44V) / 10\mu A$.
V _C	15	The output of the g _m error amplifier and the control signal of the current loop of the current-mode PWM. Switching starts at 0.7V, and higher V _C voltages corresponds to higher inductor current.
FB	16	Error Amplifier Inverting Input. A resistor divider to this pin sets the output voltage.
RUN	17	LT3782A goes into shutdown mode when this pin is below 2.3V and goes to low bias current shutdown mode when it's below 0.3V.
V _{EE2}	19	Gate Driver BGATE2 Ground. This pin should be connected to ground as close to the IC as possible.
BGATE2	20	2 nd Phase MOSFET Driver.
GBIAS2	21	Bias for Gate Driver BGATE2. Should be connected to GBIAS or an external power supply between 12V to 14V.
GBIAS1	22	Bias for Gate Driver BGATE1. Should be connected to GBIAS2.
BGATE1	23	1 st Phase MOSFET Driver.
V _{EE1}	24	Gate Driver BGATE1 Ground. This pin should be connected to ground as close to the IC as possible.
V _{CC}	27	Chip Power Supply. Good supply bypassing is required.
GBIAS	28	Internal 11V regulator output for biasing internal circuitry. Should be connected to GBIAS1 and GBIAS2. A bypass low ESR capacitor of 2μF or larger is needed and should be connected directly to the pin to minimize parasitic impedance.
NC	3, 18, 25, 26	Not Connected. Can be connected to GND.

The use of external components (resistors and capacitors) determine the vital functions in the LT3782A such as output voltage, switching frequency, peak inductor current, and loop compensation.

4.2.3 Voltage and Frequency Programming

The controller uses a feedback amplifier to control the output voltage of the converter.

The LT3782A has a 2.44V feedback reference voltage (V_{REF}) that determines the feedback resistor ratio by:

$$V_{OUT} = 2.44V \left(1 - \frac{R_{F1}}{R_{F2}} \right) \quad (4 - 31)$$

$$\frac{R_{F1}}{R_{F2}} = 1 - \frac{V_{OUT}}{2.44V} = 1 - \frac{19V}{2.44V} = 6.7869$$

$$\text{Choose } R_{F2} = 24.9k\Omega$$

$$R_{F1} = 6.7869(R_{F2}) = 6.7869(24.9k\Omega) = 169k\Omega$$

The internal functionality of the controller also determines the switching frequency.

Design requirements specify a switching frequency of 200kHz. An external resistor (R_{FREQ}) sets this frequency and with a value of 100k Ω determined from Figure 4-6 [18].

Figure 4-6: Switching Frequency versus R_{FREQ}

Additionally, the value of R_{FREQ} also relates to slope compensation. The LT3782A allows for higher power applications (such as this thesis), and therefore, includes slope compensation. Increasing slope compensation can help reduce the resulting noise and switching jitter from high power applications. The controller sets the slope compensation by an external resistor (R_{SLOPE}) with minimum size based on:

$$R_{SLOPE(min)} = \frac{R_{FREQ}}{2} \quad (4 - 32)$$

$$R_{SLOPE(min)} = \frac{100k\Omega}{2} = 50k\Omega$$

A standard value of $R_{SLOPE} = 59k\Omega$ was chosen.

4.2.4 Operating Conditions

Two important parameters, output resistance (R_o) and duty cycle (D), determine circuit behavior and component sizing. Based on the design requirements and the DC voltage transfer function of the Modified-Multiphase Boost Converter (4-27), the duty cycle can be determined from the input and output voltages (4-33). Additionally, output voltage and current ratio (via Ohm's Law) determines the output resistance (4-34).

$$D = 1 - \frac{V_{in}}{V_{out}} \quad (4 - 33)$$

$$D = 1 - \frac{12V}{19V} = 0.3684 = 36.84\%$$

$$R_o = \frac{V_{out}}{I_{out}} \quad (4 - 34)$$

$$R_o = \frac{19V}{5A} = 3.8\Omega$$

4.2.5 Inductor Calculations

The selection of inductor size in a DC/DC boost converter proves especially important as it affects input ripple current and therefore, efficiency throughout a converter. First, the desired inductor (input) ripple current needs to be determined. According to design requirements, the input current ripple ($\% \Delta I_{in}$) was chosen to be 10% for the Modified-Multiphase Boost Converter. However, standard boost converters typically experience a much higher input current ripple of around 40%. The purpose of this thesis is to show how the modified and multiphase aspects of this converter reduce ripple, not the sizing of the inductors. Therefore, to show improvement upon a standard boost based on the modified and multiphase aspects, the inductors should be sized according to a 40% input current ripple (4-35). The current ripple determines the inductor's critical inductance (L_c), the smallest possible value of inductance (4-36) [18].

$$\Delta I_{in_{40\%}} = 40\% \times \frac{I_{out} \times V_{out}}{V_{in}} \quad (4 - 35)$$

$$\Delta I_{in_{40\%}} = 40\% \times \frac{5A \times 19V}{12V} = 3.167A$$

$$L_c = \frac{V_{in} \times D}{f_{sw} \times \Delta I_{in_{40\%}}} \quad (4 - 36)$$

$$L_c = \frac{12V \times 0.3684}{200kHz \times 3.167A} = 6.98\mu H$$

A standard inductance value of 10 μ H was selected.

Note, this inductance value of 10 μ H corresponds to the value of each (single) inductor in a standard boost multiphase topology. In the Modified Boost configuration, Lentz recommended a 50/50 split in inductor sizing to allow for evenly shared current ripple in each inductor (for each phase in multiphase) and to simplify component selection. Observe that in the Modified-

Multiphase Boost Converter topology there are three inductors, and L_1 connects to both phases. Therefore, choosing a standard inductance value of $4.7\mu\text{H}$ for inductors L_1 , L_2 , and L_3 allows for a 50/50 split between inductors in each phase ($4.7\mu\text{H} + 4.7\mu\text{H} = 9.4\mu\text{H} \approx 10\mu\text{H}$). Notice that each pair per phase, L_1/L_2 and L_1/L_3 , represent the 50/50 split in inductance for each phase.

4.2.6 Current Sensing Circuitry

The LT3782A controller monitors and controls the current in the Modified-Multiphase Boost Converter. The LT3782A utilizes sense resistors (R_S) in series with the switches (MOSFETS). When the switch is on, the controller limits the voltage drop across the sense resistor to 63mV (V_{SENSE}), thereby limiting the peak inductor current to $63\text{mV}/R_S$. Thus, Equation 4-37 yields the maximum value of R_S [18].

$$R_S \leq V_{\text{SENSE}} \times \frac{1 - D}{1.2 \times \frac{I_{\text{out}}}{2}} \quad (4 - 37)$$

$$R_S \leq 63\text{mV} \times \frac{1 - 0.3684}{1.2 \times \frac{5\text{A}}{2}} = 13.26\text{m}\Omega$$

A standard value of $R_S = 10.5\text{m}\Omega$ was selected.

Additionally, the current sense pins on the LT3782A controller require an RC filter (R_1 and C_1 for phase 1) to ensure a clean signal so the controller can easily interpret the current levels. Refer to Figure 4-7 for the switch, sense resistor, and RC circuit. Selecting values for the RC filter(s) based on values in typical applications for the controller yields values of:

$$R_1 = R_2 = 10\Omega$$

$$C_1 = C_2 = 10\text{nF}$$

Figure 4-7: Current Sense Circuit

4.2.7 Output and Modified Capacitor Sizing

In a DC/DC converter, the output capacitor helps alleviate output voltage ripple and improve stability of the converter. The standard boost output capacitance equation (4-38) provides a rough estimate at a minimum capacitance value. Note, the equation has “ $2f_{sw}$ ” to account for the double switching frequency the output of the converter experiences due to the two-phase configuration and interleaving [12].

$$C_{out} = \frac{D}{\frac{\Delta V_{out}}{V_{out}} \times 2f_{sw} \times R_o} \quad (4 - 38)$$

$$C_{out} = \frac{0.3684}{2\% \times 2(200kHz) \times 3.8\Omega} = 12.1\mu F$$

Actual design requires a larger capacitance value to ensure a stable converter and low output voltage ripple, using a larger capacitance of $42\mu F$ achieves these requirements.

Combining a larger 22 μ F electrolytic capacitor (C_{out1}) and two 10 μ F ceramic capacitors (C_{out2} and C_{out3}) in parallel attains a value of 42 μ F.

As previously discussed, the value of the additional modified capacitor (C_{m1}) should be larger than 1 μ F to ensure resonant frequencies don't interfere with the 200kHz switching frequency. Additionally, the value of C_{m1} cannot be larger than the value of the output capacitance(42 μ F). Therefore, choosing $C_{m1} = 30\mu F$ meets these requirements.

4.2.8 Input and Additional Capacitor Sizing

Input capacitance helps reduce input ripple in a DC/DC converter. The standard boost input capacitance equation (4-39) provides a rough estimate at a minimum capacitance value. Note, the inductance value L in Equation 4-39 should be the inductance value per phase. Also, while not specified in the design requirements, the percent input voltage ripple, $\Delta V_{in}/V_{in}$, should be very small such as 2%.

$$C_{in} = \frac{D}{\frac{\Delta V_{in}}{V_{in}} 8 f_{sw}^2 L} \quad (4 - 39)$$

$$C_{in} = \frac{0.3684}{\frac{2\%}{100} 8 (200kHz)^2 (10\mu H)} = 5.76\mu F$$

As with output capacitance, to ensure low voltage ripple, the actual converter design requires larger input capacitance such as 20 μ F. Combining two 10 μ F ceramic capacitors in parallel (C_{in1} and C_{in2}) achieves this 20 μ F total capacitance.

As with most controllers, the LT3782A has complexity and requires a delay, or start up time, upon powering up to fully initialize before operation. An external capacitor connected to the soft-start pin (SS) on the controller determines the soft-start time by Equation 4-40. A soft-

start time of greater than 1msec proves reasonable to ensure proper functionality of the controller.

$$t_{ss} = \frac{C_{ss} \times V_{ss}}{i_{ss}} = \frac{C_{ss} \times 2.44V}{10\mu A} > 1msec \quad (4 - 40)$$

$$C_{ss} > \frac{10\mu A}{2.44V} \times 1msec = 4.1nF$$

A standard value of $C_{ss} = 6.8nF = 6800pF$ was selected.

The LT3782A requires additional capacitors on its V_{CC} and $GBias$ pins. These pins are not affected by different converter topologies and the datasheet recommends a value of C_{bias} larger than $2\mu F$. Choosing values from typical applications of the LT3782A suffices:

$$C_{VCC} = 1\mu F$$

$$C_{bias} = 2.2\mu F$$

4.2.9 Control Loop Compensation

Implementing control loop compensation ensures a stable system. A Type 2-B compensation creates a pole and a zero in the control-to-output transfer function using an RC circuit connected to the compensation feedback pin (V_C) of the controller. A right-hand plane (RHP) zero arises from the disconnection of the inductor(s) and output (load) during the switch on states. Equation 4-41 shows the calculations of the RHP zero frequency (f_{z-RHP}).

Recommended control loop compensation places the crossover frequency (f_c) (Equation 4-42) about a decade below the RHP zero [19].

$$f_{z-RHP} = \frac{R_o \left(\frac{V_{in}}{V_{out}} \right)^2}{2\pi \times L} \quad (4 - 41)$$

$$f_{z-RHP} = \frac{3.8\Omega \left(\frac{12V}{19V}\right)^2}{2\pi \times 10\mu H} = 24.12kHz$$

$$f_c \approx \frac{f_{z-RHP}}{10} \quad (4-42)$$

$$f_c \approx \frac{24.12kHz}{10} = 2.412kHz$$

The Type 2-B compensation creates a pole and zero approximately at frequencies two decades below the crossover frequency and at the crossover frequency respectively. Many error amplifiers have an output resistance (R_{out}) of around $50k\Omega$. The pole and zero frequencies (f_{p-comp} and f_{z-comp}) determine the necessary compensation capacitor and resistor (C_{comp} and R_{comp} respectively) sizes (4-43 and 4-44) [19].

$$f_{p-comp} = \frac{1}{2\pi \times C_{comp} R_{out}} \approx \frac{f_c}{100} \quad (4-43)$$

$$f_{z-comp} = \frac{1}{2\pi \times C_{comp} R_{comp}} \approx f_c \quad (4-44)$$

$$C_{comp} \approx \frac{1}{2\pi \times \frac{f_c}{100} \times R_{out}} = \frac{1}{2\pi \times \frac{2.412kHz}{100} \times 50k\Omega} = 131.97nF$$

A standard value of $C_{comp} = 133nF$ was selected.

$$R_{comp} \approx \frac{1}{2\pi \times C_{comp} \times f_c} = \frac{1}{2\pi \times 133nF \times 2.412kHz} = 496.13\Omega$$

A standard value of $R_{comp} = 500\Omega$ was selected.

4.2.10 Design Summary

Table 4-2 shows a summary of the components, their descriptions, and their values.

Table 4-2: Modified-Multiphase Boost Component Description and Values

Symbol(s)	Description	Value
R_{F1}	Voltage feedback resistor #1	169 k Ω
R_{F2}	Voltage feedback resistor #2	24.9 k Ω
R_{FREQ}	Frequency set resistor	100 k Ω
R_{SLOPE}	Slope compensation resistor	59 k Ω
L_1, L_2, L_3	Power stage inductors	4.7 μ H
R_{S1}, R_{S2}	Current sense resistors	10.5 m Ω
R_1, R_2	Current sense RC filter resistors	10 Ω
C_1, C_2	Current sense RC filter capacitors	10 nF
C_{out}	Output capacitance	42 μ F
C_{m1}	Modified boost capacitance	30 μ F
C_{in}	Input capacitance	20 μ F
C_{ss}	Soft-start capacitance	6800 pF
C_{vcc}	Controller power supply bypass capacitor	1 μ F
C_{bias}	Bias gate driver bypass capacitor	2.2 μ F
C_{comp}	Control loop compensation capacitor	133 nF
R_{comp}	Control loop compensation resistor	500 Ω

4.3 Design Simulation

The Modified-Multiphase Boost Converter design needs verification via computer simulation to validate component calculations, to determine if it meets design requirements, and to compare against a standard multiphase boost converter. Using LTSpice software allows the use of the actual LT3782A controller model in simulation.

4.3.1 Standard-Multiphase Boost Simulation

To determine the Modified-Multiphase Boost Converter's improvements, it needs comparison against a standard-multiphase boost. Figure 4-8 shows the LTSpice schematic of a standard-multiphase boost converter design and the transient simulation setup.

Figure 4-8: LTSpice Simulation for Standard-Multiphase Boost

Figure 4-9 shows the standard-multiphase boost's input and output voltage waveforms. Observe how steady-state (DC) values yield: $V_{in} = 12V$ and $V_{out} = 18.999V$, very close to the design requirement values of 12V and 19V respectively. Additionally, Figure 4-10 shows a zoomed-in view of the output voltage to determine the output voltage ripple (4-45).

Figure 4-9: Standard-Multiphase Boost Input and Output Voltage

Figure 4-10: Standard-Multiphase Boost Output Voltage Ripple

$$\% \Delta V_{out} = \frac{\Delta V_{out}}{V_{out(avg)}} \times 100 \quad (4 - 45)$$

$$\% \Delta V_{out} = \frac{42.308mV}{18.999V} \times 100 = 0.223\%$$

Note that the standard-multiphase boost even meets the output voltage ripple requirements (~2%). Figure 4-10 also shows that the output voltage has a frequency around 389kHz, almost exactly twice the switching frequency (200kHz ideally), as expected, due to the two-phase configuration. Due to a multiphase topology, this higher frequency at the output leads to ripple cancellation, providing an explanation of the low output voltage ripple [12].

Figure 4-11 shows the standard-multiphase boost's input and output current waveforms. Notice how the steady-state (DC) values yield: $I_{in} = 8.256A$ and $I_{out} = 5.0016A$, very close to the design requirement values of 8.8A (from $P_{out} = 95W$ and $\eta = 90\%$) and 5A respectively. Additionally, Figure 4-12 shows a zoomed-in view of the input current to determine the input current ripple (4-46).

Figure 4-11: Standard-Multiphase Boost Input and Output Current

Figure 4-12: Standard-Multiphase Boost Input Current Ripple

$$\% \Delta I_{in} = \frac{\Delta I_{in}}{I_{in(avg)}} \times 100 \quad (4 - 46)$$

$$\% \Delta I_{in} = \frac{0.848A}{8.256A} \times 100 = 10.27\%$$

Observe how the standard-multiphase boost nearly meets the input current ripple design requirements of ~10%. As discussed in Chapter 3, standard single-phase boost converters typically target an input current ripple of around 40%. However, with a standard *multiphase* boost converter, with close to 50% duty cycle, an input current ripple of around 20% or less proves a reasonable target due to ripple cancellation from interleaving [14]. Therefore, the standard-multiphase converter producing an input current ripple of 10.27% is reasonable.

Based on the voltage and current simulation results in Figures 4-9 and 4-11, the input and output power can be calculated (4-47 and 4-48). From these power calculations, the efficiency of the standard-multiphase boost converter can be determined as calculated using equation (4-49).

$$P_{in} = V_{in} \times I_{in} \quad (4 - 47)$$

$$P_{in} = 12V \times 8.256A = 99.072W$$

$$P_{out} = V_{out} \times I_{out} \quad (4 - 48)$$

$$P_{out} = 18.999V \times 5.0016A = 95.025W$$

$$Efficiency = \eta = \frac{P_{out}}{P_{in}} \times 100\% \quad (4 - 49)$$

$$\eta = \frac{95.025W}{99.072W} \times 100\% = 95.92\%$$

Note that the standard-multiphase boost converter still achieves a high efficiency and meets the design requirements (> 90%).

Figure 4-13 shows the gate switching frequency for each MOSFET.

Figure 4-13: Standard-Multiphase Boost Switching Frequency

First, observe the switching frequency; $f_{sw} = 193.6\text{kHz}$, very close to the 200kHz designed for. Also, notice the simulated duty cycle, based on each switch on time and period (4-50), yields a value of 40.08% , very close to the expected value of 36.84% .

$$D = \frac{t_{on}}{T_{sw}} \times 100\% \quad (4-50)$$

$$D = \frac{2.07\mu\text{sec}}{5.165\mu\text{sec}} \times 100\% = 40.08\%$$

Lastly, notice the interleaving behavior of the two switches; neither switch conducts (high voltage) at the same time as they operate 180 degrees out-of-phase.

Figure 4-14 shows the inductor current for both L_1 and L_2 . Equation 4-51 determines the inductor current ripple.

Figure 4-14: Standard-Multiphase Boost Inductor Currents

$$\% \Delta I_L = \frac{\Delta I_L}{I_{L(avg)}} \times 100 \quad (4 - 51)$$

$$\% \Delta I_L = \frac{2.366A}{4.108A} \times 100 = 57.59\%$$

Large inductor current ripple is typical, especially with standard boost designs. Also, observe the multiphase interleaving behavior once more as I_{L1} and I_{L2} are perfectly 180 degrees out-of-phase of each other.

4.3.2 Modified-Multiphase Boost Simulation

For comparison against the standard-multiphase boost and for verification of design requirements, the Modified-Multiphase Boost requires LTSpice simulation. Figure 4-15 shows the LTSpice schematic of the Modified-Multiphase Boost and the transient simulation setup.

Figure 4-15: LTSpice Simulation for Modified-Multiphase Boost

Figure 4-16 depicts the Modified-Multiphase Boost's input and output voltage waveforms. Notice how steady-state (DC) values produce $V_{in} = 12V$ and $V_{out} = 18.789V$, both close to the design requirement values of 12V and 19V respectively. Furthermore, Figure 4-17 shows a zoomed-in view of the output voltage to determine the output voltage ripple (4-52).

Figure 4-16: Modified-Multiphase Boost Input and Output Voltage

Figure 4-17: Modified-Multiphase Boost Output Voltage Ripple

$$\% \Delta V_{out} = \frac{\Delta V_{out}}{V_{out(avg)}} \times 100 \quad (4-52)$$

$$\% \Delta V_{out} = \frac{62.34mV}{18.789V} \times 100 = 0.332\%$$

Observe that the Modified-Multiphase Boost meets the output voltage ripple requirements (~2%). Figure 4-17 also depicts that the output voltage has a frequency around 388.9kHz, nearly exactly twice the switching frequency (200kHz ideally), as expected, due to the two-phase configuration. The multiphase configuration leads to this higher frequency output and ripple cancellation, leading to the low output voltage ripple in the LTSpice simulation.

Figure 4-18 shows the Modified-Multiphase Boost's input and output current waveforms. Observe how the steady-state (DC) values produce: $I_{in} = 8.085A$ and $I_{out} = 4.944A$, very close to the design requirement values of 8.8A (from $P_{out} = 95W$ and $\eta = 90\%$) and 5A respectively. Furthermore, Figure 4-19 shows a zoomed-in view of the input current to determine the input current ripple (4-53).

Figure 4-18: Modified-Multiphase Boost Input and Output Current

Figure 4-19: Modified-Multiphase Boost Input Current Ripple

$$\% \Delta I_{in} = \frac{\Delta I_{in}}{I_{in(avg)}} \times 100 \quad (4 - 53)$$

$$\% \Delta I_{in} = \frac{101.07mA}{8.085A} \times 100 = 1.25\%$$

Observe how the Modified-Multiphase Boost easily meets the maximum input ripple design requirement of 10%. This shows how the modified design reduces the input ripple current to even lower values than a standard multiphase design.

Based on the voltage and current simulation results in Figures 4-16 and 4-18, the input and output power can be calculated (4-54 and 4-55). From these power calculations, the efficiency of the Modified-Multiphase Boost Converter can be determined (4-56).

$$P_{in} = V_{in} \times I_{in} \quad (4 - 54)$$

$$P_{in} = 12V \times 8.085A = 97.02W$$

$$P_{out} = V_{out} \times I_{out} \quad (4 - 55)$$

$$P_{out} = 18.789V \times 4.944A = 92.893W$$

$$Efficiency = \eta = \frac{P_{out}}{P_{in}} \times 100\% \quad (4 - 56)$$

$$\eta = \frac{92.893W}{97.02W} \times 100\% = 95.75\%$$

Note, the Modified-Multiphase Boost Converter achieves high efficiency and meets the larger than 90% design requirement.

Figure 4-20 shows the gate switching frequency for each MOSTFET.

Figure 4-20: Modified-Multiphase Boost Switching Frequency

Notice the simulated switching frequency obtained, $f_{sw} = 196.46\text{kHz}$, very close to the 200kHz value designed for. Additionally, observe the duty cycle, based on switch on and off time (4-57) yields a value of 40.59%, close to the expected value of 36.84%.

$$D = \frac{t_{on}}{T_{sw}} \times 100\% \quad (4 - 57)$$

$$D = \frac{2.066\mu sec}{5.09\mu sec} \times 100\% = 40.59\%$$

Finally, observe the interleaving behavior of the two switches as neither conducts at the same time as they operate 180 degrees out-of-phase. Note also the similarities of Figure 4-20 and 4-2, showing how the switches and diodes determine the different states of the converter. Switch 1 and Switch 2 conduct at different times (States 1 and 2 respectively), and there exists a time in-between each switch going on or off where neither conduct (State 3).

Figure 4-21 shows the inductor current for L₁ and Figure 4-22 shows the inductor current for both L₂ and L₃. Equation 4-58 determines the inductor current ripples.

Figure 4-21: Modified-Multiphase Boost L₁ Inductor Current

Figure 4-22: Modified-Multiphase Boost L_2 and L_3 Inductor Currents

$$\% \Delta I_L = \frac{\Delta I_L}{I_{L(avg)}} \times 100 \quad (4 - 58)$$

$$\% \Delta I_{L1} = \frac{85.82mA}{8.059A} \times 100 = 1.065\%$$

$$\% \Delta I_{L2/L3} = \frac{5.006A}{4.026A} \times 100 = 124.34\%$$

Due to the modifications, L_1 has significantly lower current ripple, leading to the desirable low input current ripple. However, the lower input/ L_1 current ripple corresponds to larger L_2/L_3 current ripple. When it comes to physical, hardware implementation, L_2 and L_3 need to have appropriate sizing to ensure excessive overheating does not occur from the large ripple currents. That way, the Modified-Multiphase Boost Converter can be efficient with clean input signals. Additionally, notice how the current waveforms for L_2 and L_3 are perfectly 180 degrees out-of-phase of each other, depicting the multiphase interleaving behavior.

Figures 4-23 and 4-24 show the zoomed-in waveforms of the modified capacitor's (C_{m1}) steady-state voltage and current respectively. Equation 4-59 determines the voltage ripple of the modified capacitor.

Figure 4-23: Modified-Multiphase Boost Modified Capacitor Voltage

Figure 4-24: Modified-Multiphase Boost Modified Capacitor Current

$$\% \Delta V_{C_{m1}} = \frac{\Delta V_{C_{m1}}}{|V_{C_{m1}(avg)}|} \times 100 \quad (4 - 59)$$

$$\% \Delta V_{C_{m1}} = \frac{42.397mV}{|-6.797V|} \times 100 = 0.624\%$$

Notice how the voltage across C_{m1} fluctuates around -6.8V, nearly the same as the difference between the input and output voltage levels ($12V - 19V = -7V$). Due to Volt-Second Balance (VSB), an inductor sees zero average voltage across it [1]. Therefore, in DC steady-state, the modified capacitor (C_{m1}) essentially connects the input and output terminals of the converter, showing that -7V across C_{m1} is expected. Additionally, the percent ripple voltage of C_{m1} yields a value around 0.624%, very small as expected. The RMS value of the current through C_{m1} , 0.556A, proves useful in sizing C_{m1} for hardware implementation.

4.3.3 Simulation Comparison

Comparing the simulation results of a standard multiphase boost converter and that of the Modified-Multiphase Boost Converter provides preliminary insight into the potential benefits the Modified-Multiphase Boost could have in actual, physical implementation. Table 4-3 shows the LTSpice simulation measurements of both the standard-multiphase boost and the Modified-Multiphase Boost converters.

As shown in Table 4-3, both multiphase boost converters have similar values for most parameters. Both have incredibly low output voltage ripples (both less than 0.4%), indicative of a smooth and efficient output. The two multiphase converters also both have extremely high efficiency (>95%). One main difference between the two multiphase converters resides in input current ripple. While the standard-multiphase boost converter offers considerably lower input

current ripple than many standard, single-phase boost converters (10% versus 40%), the Modified-Multiphase Boost Converter lowers the input current ripple even more. The lower current ripple arises from a large current ripple in L_2 and L_3 . However, this large ripple occurs within the power stage and with the help of the modified capacitor (C_{m1}), and it doesn't create adverse effects in the input and output signals. With an input current ripple of only about 1%, the Modified-Multiphase Boost offers a high-power solution with a clean input and output.

Table 4-3: Multiphase Boosts Simulation Comparison

Parameter	Standard-Multiphase Boost	Modified-Multiphase Boost
V_{in}	12 V	12 V
V_{out}	18.999 V	18.789 V
$\% \Delta V_{out}$	0.223%	0.332%
I_{in}	8.256 A	8.085 A
$\% \Delta I_{in}$	10.24%	1.25%
I_{out}	5.0016 A	4.944 A
η	95.92%	95.75%
f_{sw}	193.6 kHz	196.46 kHz
D	40.08%	40.59%
$I_{L1(avg)}$ (modified)	N/A	8.059 A
$I_{L(avg)}$ (standard)	4.108 A	4.026 A
$I_{L2/L3(avg)}$ (modified)		
$\% \Delta I_{L1}$ (modified)	N/A	1.065%
$\% \Delta I_L$ (standard)	57.59%	124.34%
$\% \Delta I_{L2/L3}$ (modified)		
V_{AVG_Cm1}	N/A	-6.797 V
$\% \Delta V_{Cm1}$	N/A	0.624%
I_{RMS_Cm1}	N/A	0.556 A

5. HARDWARE, TEST, AND RESULTS

Testing of any DC/DC converter involves selection of components, layout and design of printed circuit board (PCB), construction of the PCB, and test and measurement of the converter's performance. Performing the previously listed steps for both the Modified-Multiphase and the Standard-Multiphase boost converters allows for comparison between the two circuits to observe any differences or improvements.

5.1 Component Selection

Every DC/DC converter components must have voltage, current, and/or power ratings that exceed the expected maximum voltage, current, and/or power levels they would experience in the circuit respectively. Since both the Modified-Multiphase and the Standard-Multiphase boost converters have the same design requirements, they will both experience similar voltage, current, and power levels. Therefore, the components in each boost converter will undergo similar stresses, allowing for the nearly all of the same components to be used in each.

5.1.1 Inductor Selection

With a maximum input/inductor current of around 8A in both converters, the chosen inductors must have larger current ratings. The selection of the Würth Elektronik WE-HCC SMD High Current Cube line of inductors was chosen due to their high current and temperature handling capabilities. Additionally, the Würth High Current Cube inductors all come in the same size package and have a large enough range of values with high enough current ratings to allow for the same PCB design to be used for the 4.7 μ H and 10 μ H sizes for the modified and standard

designs respectively. Table 5-1 summarizes the essential ratings of each inductor. Note that the saturation current ratings (I_{sat}) and the root mean squared (RMS) current ratings (I_{rms}) easily meet current requirements [20].

Table 5-1: MSS1278T Series Inductor Ratings

Circuit	Inductance	Tolerance	I_{sat}	I_{rms}	$R_{\text{DC(MAX)}}$	Footprint
Modified-Multiphase Boost	4.7 μH	20%	17.0 A	15.5 A	7.9 m Ω	138 mm ²
Standard-Multiphase Boost	10 μH	20%	10.0 A	9 A	16.7 m Ω	138 mm ²

5.1.2 Switch and Diode Selections

The switch and diode components undergo similar voltage and current stresses since they alternately conducting and therefore see the same peak current. Each boost converter uses an IRF60B217 n-channel MOSFET for the switch and a MBR10100G Schottky diode [21] [22]. Table 5-2 summarizes each component's ratings. Observe that the voltage and current ratings of the switch and diode greatly surpass the levels expected in both boost converters.

Table 5-2: Switch and Diode Ratings

IRF60B217 MOSFET Switch		MBR10100G Diode	
Peak Drain-Source Voltage (V_{DS})	60V	Peak Repetitive Reverse Voltage (V_{rrm})	100V
Continuous Drain Current (I_{D})	60A	Average Forward Current (I_{F})	10A
Max Drain-Source Resistance ($R_{\text{DS(ON)}}$)	9m Ω	Maximum Instantaneous Forward Voltage (V_{F})	0.95V

5.1.3 Capacitor Selection

Both the Modified-Multiphase and the Standard-Multiphase boost converters utilize an aluminum electrolytic capacitor for one of the output capacitors. This type of capacitor allows for a large capacitance to be obtained in a small physical package. For both converters, one 22 μ F aluminum electrolytic capacitor is used in each output capacitor bank. Both boost converters also use multilayer ceramic capacitors (MLCCs). MLCCs can have small size while minimizing the equivalent series resistance (ESR). Minimizing ESR proves key in reducing losses and maximizing efficiency. Therefore, MLCCs are used for all input, signal, and filter capacitors. Configuring multiple capacitors in parallel sums each capacitance to a total, larger capacitance. Creating a larger capacitance on the output helps to reduce output ripple current. Additionally, placing capacitors in parallel effectively places their ESRs in parallel; resistors in parallel reduces overall resistance. Therefore, by placing two 10 μ F MLCCs and the 22 μ F aluminum electrolytic capacitors in parallel at the converter's output, the large overall capacitance of 42 μ F and small overall ESR can be obtained to diminish output ripple and maximize efficiency.

5.1.4 Efficiency Calculation

Attaining a high efficiency in a DC/DC converter remains a primary goal; however, losses in power components, such as the switch and diode, cause converters to have lower than ideal (100%) efficiencies. While other power losses, such as in C_{in} and C_{out} ESR dissipation, do occur, they typically account for less than 2% of total losses and can be minimized via techniques such as multiple input/output capacitors in parallel. The total efficiency can be calculated by adding all the power losses and dividing the output power by the sum of the output power and total losses (5-1).

$$\% \text{Efficiency} = \eta = 100 \times \frac{P_{out}}{P_{in}} = 100 \times \frac{P_{out}}{P_{out} + P_{Loss}} \quad (5 - 1)$$

Where P_{Loss} is the sum of all power losses.

Recall from design requirements in Table 3-1, and output voltage (V_{out}) of 19V and output current (I_{out}) of 5A yield an output power (P_{out}) of 95W.

The switching controller, LT3782A, presents itself as the first source of power loss to consider. According to the datasheet, the power loss in the controller (P_{IC}) depends on the DC supply current to the controller (I_Q), specified in the electrical characteristics, and the gate charge (Q_G) of the MOSFET switch (specified in the MOSFET's datasheet) (5-2) [18] [21].

$$I_Q = 11mA \text{ and } Q_G = 44nC$$

$$P_{IC} = V_{in}(I_Q + f \times Q_G) = 12V(11mA + 200kHz \times 44nC) = 237.6mW \quad (5 - 2)$$

The switch also incurs a reasonable power loss. Like with all switches, the losses arise from switching and conduction losses [21]. Equation 5-3 shows the MOSFET losses [18].

$$P_{FET} = R_{DS(ON)}D\rho_T \left(\frac{I_{out(max)}}{2(1-D)} \right)^2 + kV_{out}^2 C_{RSS} f_{sw} \left(\frac{I_{out(max)}}{2(1-D)} \right) \quad (5 - 3)$$

$$P_{FET} = (7.3m\Omega)(0.3684)(0.4\%/^{\circ}C) \left(\frac{9.24A}{2(1-0.3684)} \right)^2$$

$$+ (1.7)(19V)^2(140pF)(200kHz) \left(\frac{9.24A}{2(1-0.3684)} \right) = 183.25mW$$

Losses in the current sense resistors (5-4), windings/resistance of the inductors (5-5), and in the diodes (5-6) also contribute to the total overall losses. Simple calculations based on the current, duty cycle, and resistance can be performed to account for these power losses [18].

Additionally, for the inductor loss of Equation 5-5 (P_{L_DCR}), notice how the equation needs to be doubled to account for the split inductor configuration of the Modified-Multiphase Boost Converter. That is, for any given interleaving switching cycle, current flows through two inductors at a time.

$$P_{R(sense)} = R_S \times D \times \left(\frac{I_{out(max)}}{2(1-D)} \right)^2 \quad (5-4)$$

$$P_{R(sense)} = (10.5m\Omega)(0.3684) \times \left(\frac{9.24A}{2(1-0.3684)} \right)^2 = 206.97mW$$

$$P_{L(DCR)} = 2 \times R_{DCR} \times \left(\frac{I_{out(max)}}{2(1-D)} \right)^2 = 2(15.5m\Omega) \left(\frac{9.24A}{2(1-0.3684)} \right)^2 = 1.659W \quad (5-5)$$

$$P_{Diode} = V_D \frac{I_{out(max)}}{2} = (0.5V) \frac{9.24A}{2} = 2.31W \quad (5-6)$$

As stated before in Equation 5-1, the summation of each loss yields the total losses (5-7), which divided by the output power provides an approximate efficiency estimate.

$$P_{Loss} = P_{IC} + P_{FET} + P_{R(sense)} + P_{L(DCR)} + P_{Diode} \quad (5-7)$$

$$P_{Loss} = 237.6mW + 183.25mW + 206.97mW + 1.659W + 2.31W = 4.597$$

$$\%Efficiency = \eta = 100 \times \frac{P_{out}}{P_{in}} = 100 \times \frac{P_{out}}{P_{out} + P_{Loss}} = 100 \times \frac{95W}{95W + 4.597W} = 95.38\%$$

Observe how the calculated estimated efficiency of 95.38% is very close to the simulated value of 95.75% presented in Table 4-3.

5.2 Implementation

Once rating values have been calculated and components selected, the next step in hardware implementation involves the design of printed circuit board (PCB) layout. Altium Designer provides the necessary tools to develop a PCB and layout the components from a simple circuit schematic.

5.2.1 PCB Schematic

As previously seen in Figures 4-8 and 4-15, the Modified-Multiphase and Standard-Multiphase boost converter schematics have many similarities. The only differences arise with the splitting of the inductor and addition of a capacitor in the Modified-Multiphase Boost topology. To design and manufacture the same PCB for both converter designs, the Modified-Multiphase Boost Converter schematic provides the base for the PCB circuit schematic, requiring only a couple of additions. First, observe that by simply placing a 0Ω resistor in parallel with L_1 in the Modified-Multiphase topology would short out the inductor location and yield the same inductor configuration as the Standard-Multiphase topology. Also, observe that by shorting out this inductor and not soldering any component in the modified capacitor location (C_{m1}) in the modified topology yields the exact same power stage configuration in the standard topology. Lastly, adding connections in series with each inductor location allows for wire loops to be created in series with the inductors to allow for placement of a current probe clamp to test and view the inductor currents.

5.2.2 PCB Layout

As described before, with a few additions, both the Modified-Multiphase and Standard-Multiphase boost converter designs can utilize the same PCB layout design. The basics steps to PCB design follow Altium Designer's recommended tutorial on designing and manufacturing a PCB. The tutorial includes steps on sizing the board, placing and connecting components, adjusting the layers of the board, setting board rules, and error checking [23]. Knowing the manufacturer of the boards also helps in the PCB design process. Selecting Oshpark as the board manufacturer allows for inexpensive boards to be ordered in small batches and arrive in 2-3 weeks, ideal for prototyping and proof-of-concept designs such as the one conducted in this thesis. Selecting the PCB manufacturer before the layout design allows for the designer to know the manufacturer's sizing specifications, including minimum hole size and component spacing.

The LT3782A controller chip specifies PCB layout considerations that should be closely followed. These requirements include locating all power components in a tight area to minimize losses and to place the current sense resistors (R_{S1} and R_{S2}) as close as possible to each respective power switch with short traces to limit noise. The LT3782A datasheet also states that a solid ground plane under the circuit should be used with local vias to ground plane connections for all component pads that connect to ground.

Using all design and layout considerations above, the PCB design uses a two-layer rectangular board with copper traces and connections. Copper polygon pours connected to ground on both layers meet the ground plane requirements by providing short paths from components to ground to limit conduction losses and provide a large area to dissipate any excess heat. Figures 5-1 and 5-2 depict the Altium Designer 2-dimensional annotated top and bottom layers respectively.

Figure 5-1: PCB Top Layer

Figure 5-2: PCB Bottom Layer

Construction of the PCB follows standard through-hole and surface mount soldering techniques. Figures 5-3 and 5-4 show the completely assembled Standard-Multiphase and Modified-Multiphase boost converters respectively.

Figure 5-3: Assembled Standard-Multiphase Boost Converter (Top View)

Figure 5-4: Assembled Modified-Multiphase Boost Converter (Top View)

5.3 Testing

Physical hardware testing of both the Standard-Multiphase and Modified-Multiphase boost converters requires the use of electronic test equipment to gather data and obtain waveform images. The data can then be analyzed and plotted with computer applications such as Microsoft Excel or Matlab.

5.3.1 Test Configuration

The hardware testing utilizes two Rigol DP832 Programmable DC Power Supplies, two Rigol DM3058E Multimeters, a BK Precision 8540 DC Electronic Load, and a Teledyne LeCroy HDO4104 Oscilloscope. Testing requires two Rigol power supplies to source enough input current to the boost converters. With a maximum input current around 9A, paralleling four channels (two channels per power supply), each rated at 3A, allows for a 12A maximum source current, meeting the boost input current requirement. Summing each Rigol power supply channel's source current yields the total DC input current while the electronic load shows the DC output current. The two multimeters measure the DC input and output voltages with high precision at each boost converter's input/output terminals. Lastly, the oscilloscope, setup to trigger on the converter's top-phase MOSFET gate drive signal, allows for measurement, data gathering, and waveform capture of numerous converter variables. These variables include MOSFET gate drive signals, output voltage ripple, and input and inductor current ripples. Figure 5-5 illustrates the block diagram of hardware test configuration used for both boost converters and Figure 5-6 shows a picture of the test setup in a laboratory. Note for Figure 5-5, red arrows indicate the power flow through the circuit while black arrows depict placement of voltage/current probes.

Figure 5-5: Hardware Test Configuration Block Diagram

Figure 5-6: Hardware Test Configuration Laboratory Setup

5.3.2 Initial Results

The initial test of the Standard-Multiphase Boost Converter yielded expected results for nearly all values and state variables. However, both boost converters experienced limited output voltages during initial testing, especially the Modified-Multiphase Boost Converter. Due to the high inductor current ripple on the inner inductors (L_2 and L_3), the current sense resistor triggered the current limit, thereby restraining the pulse-width modulation (PWM) duty cycle and consequently, lowering the output voltage. The modified boost converter configuration minimizes input/ L_1 current ripple by transferring the ripple to the other inductors and the modified capacitor. However, the higher inductor ripple currents must be accommodated for by adjusting the current-sense resistance. Lowering the current-sense resistance increases the controller's current limit which should allow for the full PWM duty cycle and output voltage to be achieved. Table 5-3 shows the differences in duty cycle and output voltage from the initial current-sense resistance of $10.5\text{m}\Omega$ to the altered value of $5\text{m}\Omega$ for both multiphase boost converters.

Table 5-3: Sense Resistor Adjustment, Duty Cycle, and Output Voltage

Sense Resistance	Converter	V _{OUT} (V)	Duty Cycle (%)
10.5m Ω	Standard-Multiphase	18.663	39.84%
	Modified-Multiphase	16.016	31.75%
5m Ω	Standard-Multiphase	18.906	40.61%
	Modified-Multiphase	18.94	41.51%

In DC/DC converters, the steady-state DC values at full-load provide ample information regarding the functionality and behavior of a circuit. Table 5-4 displays each boost converter's steady-state DC values at full load (5A or ~95W). Observe that both boost converters achieve the

design requirements very close to the ideal values, including around a 19V output voltage, 95W output power, 200kHz switching frequency, and >90% efficiency.

Table 5-4: Full-Load Steady-State Converter Characteristics

Converter	V _{IN} (V)	I _{IN} (A)	P _{IN} (W)	V _{OUT} (V)	I _{OUT} (A)	P _{OUT} (W)	η (%)	f _{sw} (kHz)	D (%)
Standard-Multiphase	11.558	8.74	101.013	18.906	5.0	94.528	93.58	209.9	40.61
Modified-Multiphase	11.538	8.97	103.496	18.94	5.0	94.70	91.50	209.1	41.51

5.3.3 Standard-Multiphase Boost Results

Measuring efficiency and load regulation proves important in determining a boost converter's load-handling capabilities. Varying the electronic load current from no-load (0A) to full-load (5A) allows these values to be obtained. Figure 5-7 depicts the efficiency versus load current plot and Equation 5-8 shows the load regulation calculation. Note, at no-load 0A condition, the boost converter has 0% efficiency.

Figure 5-7: Standard-Multiphase Boost Efficiency vs. Load Current

$$\%Load\ Regulation = \frac{V_{out(low-load)} - V_{out(high-load)}}{V_{out(high-load)}} \times 100\% \quad (5-8)$$

$$\%Load\ Regulation = \frac{18.935V - 18.906V}{18.906V} \times 100\% = 0.153\%$$

Observe that this percent load regulation value of 0.153% easily meets the design requirement of less than 5%.

While load regulation measures the converter's ability to maintain output voltage at varying load currents, another important quantity, line regulation, demonstrates the converter's ability to maintain output voltage at varying input voltages. Equation 5-9 shows how to calculate percent line regulation and Table 5-5 displays the line regulation results by varying the input

voltage by $\pm 5\%$ and $\pm 10\%$. Notice how both cases easily meet the design requirement of 10% or less line regulation.

$$\% \text{Line Regulation} = \frac{V_{out(\text{high-input})} - V_{out(\text{low-input})}}{V_{out(\text{nominal-input})}} \times 100\% \quad (5-9)$$

Table 5-5: Standard-Multiphase Boost Converter Line Regulation

% V_{IN}	$V_{IN(\text{nominal})}$ (V)	$V_{IN(\text{low})}$ (V)	$V_{IN(\text{high})}$ (V)	Line Regulation
$\pm 5\%$	12	11.4	12.6	0.005%
$\pm 10\%$	12	10.8	13.2	0.026%

Output voltage and its ripple present another important metric in the behavior of boost converters. Figure 5-8 illustrates the Standard-Multiphase Boost's output voltage ripple and the switch gate driving signals. Ignoring the parasitic ringing, the top (blue) waveform follows the expected output voltage behavior, a negative linear slope when each switch conducts, resulting in an effective output voltage ripple frequency twice that of the switching frequency. Equation 5-10 calculates the percent output voltage ripple.

Figure 5-8: Standard-Multiphase Boost Output Voltage Ripple

$$\% \Delta V_{out} = \frac{\Delta V_{out}}{V_{out(avg)}} = \frac{590mV}{18.906V} = 3.121\% \quad (5 - 10)$$

While output voltage characteristics prove vital with boost converters, so does the input and inductor current behavior. Figure 5-9 displays the inductor current waveform for each phase along with the switch gate drive signals. Observe the interleaving behavior of the inductor current waveforms as they are 180 degrees out-of-phase of each other (as are the switching signals). Additionally, Equation 5-11 calculates the percent current ripple for the inductors.

Figure 5-9: Standard-Multiphase Boost Inductor Current Waveforms

$$\% \Delta I_L = \frac{\Delta I_L}{I_{L(avg)}} \times 100 = \frac{2.36A}{4.35A} \times 100 = 54.25\% \quad (5 - 11)$$

Due to the use of multiple power supplies to source DC input power, accurate input current ripple measurements could not be obtained with a current probe for the Standard-Multiphase Boost Converter due to the numerous wires. However, the Teledyne LeCroy

Oscilloscope has the functionality to take data points of a waveform. Therefore, by obtaining data points for each inductor current waveform as seen in Figure 5-9, and summing the two waveforms in Matlab, by KCL the addition of both inductor currents yields the input current. Figure 5-10 shows the Matlab plot of input current ripple from the summation of the inductor currents. Note that Matlab was also used to obtain average, minimum, maximum, and consequently, ripple for the input current. APPENDIX – A contains the *input_current_addition.m* Matlab file that performs the current addition and waveform calculations.

Figure 5-10: Standard-Multiphase Boost Input Current Ripple

Table 5-6 displays the Standard-Multiphase Boost Converter's hardware test results along with the design requirements and simulation results for comparison.

Table 5-6: Standard-Multiphase Boost Hardware Test Results

Parameter	Design Requirement	Simulation	Hardware
$\% \Delta I_{in}$	$\sim 10\%$	10.24%	8.97%
P_{out}	95 W	95.025 W	94.528 W
η	$> 90\%$	95.92%	93.58%
f_{sw}	200 kHz	193.6 kHz	209.9 kHz
$\% \Delta V_{out}$	$\sim 2\%$	0.223%	3.121%
$\% Reg_{line}$	$\sim 10\%$	N/A	0.026%
$\% Reg_{load}$	$< 5\%$	N/A	0.153%

5.3.4 Modified-Multiphase Boost Results

As before, a boost converter's load-handling capabilities can be determined through its efficiency and load regulation. Changing the electronic load from no-load current (0A) to full-load current (5A) allows for the calculation of these values. Figure 5-11 displays the efficiency versus load current plot and Equation 5-12 calculates load regulation. Notice how percent load regulation easily meets the design requirement of less than 5%.

Figure 5-11: Modified-Multiphase Boost Efficiency vs. Load Current

$$\%Load\ Regulation = \frac{V_{out(low-load)} - V_{out(high-load)}}{V_{out(high-load)}} \times 100\% \quad (5 - 12)$$

$$\%Load\ Regulation = \frac{18.96V - 18.94V}{18.94V} \times 100\% = 0.106\%$$

Line regulation measures a converter's ability to hold an output voltage steady while changing input voltages. Refer to Equation 5-9 for how to calculate line regulation. Table 5-7 shows the line regulation results by varying input voltage by $\pm 5\%$ and $\pm 10\%$. In both instances, the Modified-Multiphase Boost easily meets the design requirement of 10% or less line regulation.

Table 5-7: Modified-Multiphase Boost Converter Line Regulation

% V _{IN}	V _{IN(nominal)} (V)	V _{IN(low)} (V)	V _{IN(high)} (V)	Line Regulation
±5%	12	11.4	12.6	-0.011%
±10%.	12	10.8	13.2	0.048%

The output voltage ripple of the modified boost converter can be seen in Figure 5-12 along with the switch drive signals. At switching transitions, the parasitic oscillation, largely due to probe/measurement interference, can be neglected. Observe that the output voltage (top, blue waveform) follows a negative linear slope during the on-time of each switch, causing the effective output voltage ripple frequency to be double the switching frequency. Equation 5-13 displays the percent output voltage ripple calculation.

Figure 5-12: Modified-Multiphase Boost Output Voltage Ripple

$$\% \Delta V_{out} = \frac{\Delta V_{out}}{V_{out(avg)}} = \frac{870mV}{18.94V} = 4.593\% \quad (5 - 13)$$

Figure 5-13 depicts the inner inductor (L_2 and L_3) current waveforms for each phase along with the switch gate drive signals. Notice again the interleaving, 180 degrees out-of-phase, behavior of the current waveforms (as well as the switching signals). Additionally, observe the slight mismatch in height of each inductor current waveform. This imbalance is likely due to imperfect matching of component sizes as the inductors only have 20% tolerance in their inductance values. Therefore, one phase has a slightly larger portion of the overall system current than the other. Equation 5-14 calculates the percent current ripple for the inner inductors.

Figure 5-13: Modified-Multiphase Boost Inner Inductor Current Waveforms

$$\% \Delta I_L = \frac{\Delta I_L}{I_{L(avg)}} \times 100 \quad (5 - 14)$$

$$\% \Delta I_{L2} = \frac{4.53A}{4.803A} \times 100 = 94.32\%$$

$$\% \Delta I_{L3} = \frac{4.56A}{4.211A} \times 100 = 108.29\%$$

As expected and seen before in simulation, the inner inductors (L_2 and L_3) have very large ripple currents do to the modifications. However, even with large ripple currents, the inner inductors don't hit boundary conduction mode (BCM), the point when the minimum inductor current is 0A for only an instant, until around half of full-load output current at 2.4A. Thus, the Modified-Multiphase Boost Converter still operates in its typical PWM continuous conduction mode with a duty cycle of around 40% from half to full-load. Figure 5-14 shows the inner inductor current waveforms at BCM with a 2.4A load current. Notice how the minimum points of the current waveforms barely hit 0A while maintaining the sharp triangular shape.

Figure 5-14: Modified-Multiphase Boost Inner Inductor Current at BCM

Discontinuous conduction mode (DCM) occurs when the load current decreases to the point where the bottom part of the triangular inductor current waveform gets chopped and remains at 0A for some time during the switching period. Figure 5-15 shows the inner inductor current waveforms at one part of DCM with a 1.4A load current. Observe how the current

waveforms hit 0A and stay there until they begin to recharge with the start of the next switching cycle.

Figure 5-15: Modified-Multiphase Boost Inner Inductor Current at DCM

With the Modified-Multiphase Boost Converter topology, the input current ripple and first inductor, L_1 , have basically the same current waveforms. The only additional current would be purely AC (no DC offset) and go through the input capacitors; this should essentially be zero as the input source utilizes an ideal DC supply. Figure 5-16 displays the input/ L_1 current waveform and, using Equation 5-14, the current ripple calculation can be seen below.

Figure 5-16: Modified-Multiphase Boost Input/L₁ Current Waveform

$$\% \Delta I_{IN} = \% \Delta I_{L1} = \frac{75.5 \text{ mA}}{8.937 \text{ A}} \times 100 = 0.845\%$$

As anticipated, the Modified-Multiphase Boost Converter has extremely low input/L₁ current ripple.

With the new modifications, the modified capacitor (C_{m1}) voltage also needs measuring and verification. Since the modified capacitor goes from the node (X₁) between the first inductor (L₁) and inner inductors (L₂ and L₃) to the output node, it can't be probed directly with the oscilloscope without shorting the output. However, the modified capacitor voltage waveform can be obtained by taking separate measurements of the X₁ and output nodes and subtracting the two waveforms using the “Math” functions on the oscilloscope. Figure 5-17 shows the X₁ and output voltage waveforms (top) and the resulting modified capacitor voltage (yellow, bottom). Observe that, at the switching transitions, large parasitic oscillation occurs. These oscillations are largely

due to probe/measurement interference and can be neglected. Additionally, Equation 5-15 calculates the modified capacitor voltage ripple.

Figure 5-17: Modified-Multiphase Boost Modified Capacitor Voltage

$$\% \Delta V_{Cm1} = \frac{\Delta V_{Cm1}}{|V_{Cm1(avg)}|} \times 100 = \frac{580mV}{|-7.571V|} \times 100 = 7.66\% \quad (5-15)$$

Notice the modified capacitor voltage fluctuates around -7V , approximately the difference between input and output voltages ($12\text{V} - 19\text{V} = -7\text{V}$). Because of Volt-Second Balance, the inductors should see zero average voltage. Therefore, observing -7V across the modified capacitor at DC steady-state should be expected.

Table 5-8 displays the Modified-Multiphase Boost Converter’s hardware test results, design requirements, and simulation results for comparison.

Table 5-8: Modified-Multiphase Boost Hardware Test Results

Parameter	Design Requirement	Simulation	Hardware
$\% \Delta I_{in}$	$\sim 10\%$	1.25%	0.845%
P_{out}	95 W	92.893 W	94.7 W
η	$> 90\%$	95.75%	91.50%
f_{sw}	200 kHz	196.46 kHz	209.1 kHz
$\% \Delta V_{out}$	$\sim 2\%$	0.332%	4.593%
$\% Reg_{line}$	$\sim 10\%$	N/A	0.048%
$\% Reg_{load}$	$< 5\%$	N/A	0.106%

5.4 Hardware Comparison

The potential improvements that the Modified-Multiphase Boost Converter offers can be determined and observed by comparing the test results between the Standard-Multiphase and Modified-Multiphase boost converters. Table 5-9 compares the two converter's pertinent hardware results such as voltage and current ripples, efficiency, and switching frequency.

Table 5-9: Multiphase Boost Hardware Test Comparison

Parameter	Standard-Multiphase Boost	Modified-Multiphase Boost
$\% \Delta I_{IN} = \% \Delta I_{L1}$ (modified)	8.97%	0.845%
η	93.58%	91.50%
f_{sw}	209.9 kHz	209.1 kHz
$\% \Delta V_{out}$	3.121%	4.593%
$\% Reg_{line}$	0.026%	0.048%
$\% Reg_{load}$	0.153%	0.106%
$I_{IN(avg)} = I_{L1(avg)}$ (modified)	8.74 A	8.937 A
$I_{L(avg)}$ (standard)	4.35 A	4.803 A
$I_{L2/L3(avg)}$ (modified)		4.211 A
$\% \Delta I_L$ (standard)	54.25%	94.32%
$\% \Delta I_{L2/L3}$ (modified)		108.29%
V_{AVG_Cm1}	N/A	-7.571 V
$\% \Delta V_{Cm1}$	N/A	7.66%

By utilizing similar designs and identical components and layout, both the Standard-Multiphase and Modified-Multiphase boosts share many similarities in hardware results. Observe that both converters have around 209kHz switching frequency, both close to the design requirement of 200kHz. Additionally, both the Standard-Multiphase and Modified-Multiphase boost converters have efficiencies of 93.58% and 91.5% respectively, meeting the 90% efficiency design requirement. The Standard-Multiphase Boost likely has a slightly larger efficiency due to its lower part count, and thus has less conduction losses. Also, notice the extremely low line and load regulations for each converter, both easily meet the design requirements, indicating two robust converters that can handle varying supply voltages and load current demands. Lastly, both converters have slightly larger output voltage ripples than the 2% design requirement. Due to the large parasitic oscillations obtained using scope probes, the output voltage ripple measurement was difficult to obtain precisely. Furthermore, both ripples only slightly surpass the 2% requirement at around 3% and 4.5% for the standard and modified respectively. With an output capacitance of only 42 μ F, both the Standard-Multiphase and Modified-Multiphase boosts already produce a very small output voltage ripple with a considerably lower output capacitance than many other boost or DC/DC converter designs. Adding more capacitance via additional parallel capacitors on the output would help to lower the output voltage ripple. However, output capacitance was kept low for the sake of comparison and measuring improvements between the two multiphase converters and Lentz's previous single-phase modified design.

As expected, the Standard-Multiphase and Modified-Multiphase converters hardware results experienced differences in input and inductor currents. Notice how the input current and each phase's inductor currents maintained similar averages between standard and modified

designs. However, as expected, phase inductor current increased from 54% in the Standard-Multiphase design to 94% and 108% in the Modified-Multiphase design. This increase in inner/phase inductor current offers another explanation as to why the Modified-Multiphase Boost has marginally lower efficiency. Additionally, the phase/inner inductor ripple increase, due to the Modified-Multiphase Converter's modifications, allows for the additional input inductor (L_1), and therefore input to the converter, to have significantly lower current ripple. Input current ripple decreased from 8.97% to just 0.845% from the standard to modified designs. Thus, the Modified-Multiphase Boost Converter achieved its primary goal of reducing input current ripple. Figure 5-18 displays a bar graph showing the input current ripple percentage and the difference for both the Standard-Multiphase and Modified-Multiphase boost converters from LTSpice simulation and hardware results.

Figure 5-18: Input Current Ripple Percentage Comparison

6. CONCLUSION

This thesis utilizes a Modified-Multiphase Boost Converter design to investigate the advantages of both the modified boost converter presented in Nathan Lentz's thesis and multiphase DC/DC converters. State-space analysis at steady state showed that, even with the modified and multiphase changes, the Modified-Multiphase Boost Converter exhibited the same transfer function as a standard, single-phase boost. LTSpice simulation verified the functionality of the design and selection of components and yielded preliminary results to compare against hardware implementation.

Hardware test results produced slightly lower efficiencies and higher currents, compared to simulation, due to power dissipation in non-ideal components. These higher currents tripped the controller's current limit and lowered output voltage. To produce the desired 19V output voltage, the current sense resistance was lowered to raise the current limit and improve output voltage regulation. The hardware measurements also showed marginally larger output voltage ripples in both converters compared to simulation results. This is largely due to transients and additional ripple/losses in physical components instead of ideal simulated ones. The output voltage ripple could be lowered by increasing the output capacitance.

Comparing the Modified-Multiphase Boost Converter with a Standard-Multiphase Boost Converter allows observation of the potential improvements from the modified aspect of the design. This thesis showed a reduction in input current ripple of approximately 9% from Modified-Multiphase to Standard-Multiphase topologies. Due to the multiphase aspect, input current ripple should decrease with any multiphase DC/DC converter (standard or modified) [15]. Therefore, achieving a less drastic input current ripple reduction as the 40% decrease

experienced between single-phase converters in Lentz's thesis was expected [2]. However, the reduced input ripple comes at a price; the Modified-Multiphase Boost Converter requires an additional inductor and capacitor (at minimum), leading to larger size and higher costs.

Contrasting the Modified-Multiphase Boost Converter against Lentz's modified single-phase design allows the possible improvements of a multiphase design to be observed. While relatively low output voltage ripple was achieved with the multiphase design, Lentz's single-phase had lower percent output voltage ripple. However, the Modified-Multiphase Boost Converter still attained a low output voltage ripple while producing more than three times the output power of the modified, single-phase boost (95W to 30W respectively). Additionally, the Modified-Multiphase Boost had an efficiency around 92% while the modified, single-phase version only achieved an efficiency of 87% at a much lower power level [2].

Many applications could benefit from a lowered input current ripple in a highly efficient and high-power boost converter. The input current ripple reduction presents a tradeoff: it either reduces the input current ripple with the same filtering as before in a non-modified/multiphase design, or it reduces the input filtering required for a specific input ripple.

Complete adoption of the Multiphase-Modified Boost Converter for applications requires additional future work. Testing of this thesis required two channels each in two separate Rigol DC power supplies. This caused excess wiring as the four channels were connected in parallel to deliver the 12V input at higher current. The abundance of wires not only added to power loss and reduced efficiency, but also made it difficult to obtain an accurate input current waveform with a current probe. Therefore, using a single, higher-power DC power supply would be recommended. Additionally, Equation 5-6 showed that the diodes would experience a large power loss of 2.3W (over 2% efficiency loss). This loss was confirmed during testing as the

diodes dissipated significant heat after conducting for some time. A possible improvement could entail replacing the diodes with ones with lower forward conducting voltage (V_D), or utilize a synchronous design with MOSFETS instead of diodes. Using MOSFETS with low on resistance ($R_{DS(ON)}$ for N-channel), such as the ones used for the main switches in this thesis, would reduce power loss and improve converter efficiency. However, a synchronous design would require a different controller capable of handling a multiphase synchronous boost with four total switches. Another improvement would involve the use of lower tolerance inductors. The Modified-Multiphase Boost Converter utilized inductors with a $\pm 20\%$ tolerance. This likely led to the slight current imbalance experienced during hardware testing. Better matched inductors would have current peaks and troughs at the same levels and would help reduce current ripple even further and increase efficiency.

One possible change in the design aspect of this thesis would be to lower target efficiency for maximum current calculations. This would cause designs to plan for larger currents, obtain lower sense resistor values, and not have the issue of hitting controller current limits which can cause voltage regulation problems. Additionally, this thesis only focused on CCM operation of the Modified-Multiphase Boost Converter. Possible future work could also investigate the DCM operation of the converter as boost converters can benefit from DCM operation due to the Right Half Plane (RHP) zero in the system's transfer function. Lastly, evaluating the PCB layout and design could lead to future improvements. More robust input and output connections could be used to handle the high-power currents and provide for easier access to supply and load connections. Also, placement and routing of components and traces respectively can greatly affect noise and EMI. Proper PCB layout has increased importance with higher power and faster switching converters.

While the Modified-Multiphase Boost Converter presents a low-ripple, high efficiency, and high-power solution, future improvements could produce a more robust system for applications.

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APPENDICES

A. input_current_addition.m

```
% Input Time and Current Data from L3 and L4
% Add currents, => Input Current Plot and Ripple

% --- L3 Data Read In ---
length_test=csvread('C4L300000.csv',5,0);
temp_dsd=size(length_test)-[1,0];
row_num=temp_dsd(1);
length_test=1;
end_row=row_num+5;
L3=csvread('C4L300000.csv',5,0,[5 0 end_row 1]);

[row_num column_num]=size(L3);

time_L3=L3(1:row_num,1);
current_L3=L3(1:row_num,2);

% --- L4 Data Read In ---
length_test=csvread('C4L400000.csv',5,0);
temp_dsd=size(length_test)-[1,0];
row_num=temp_dsd(1);
length_test=1;
end_row=row_num+5;
L4=csvread('C4L400000.csv',5,0,[5 0 end_row 1]);

[row_num column_num]=size(L4);

time_L4=L4(1:row_num,1);
current_L4=L4(1:row_num,2);

% --- Plot Each Inductor Current Waveform for Verification ---
figure(1)
plot(time_L3,current_L3, time_L4,current_L4)
title('Inductor Current Waveforms')
xlabel('Time [s]')
ylabel('Current [A]')
legend('L3','L4')
xlim([-1E-5 1E-5])

% --- Add Currents => Input Current ---
current_IN = current_L3 + current_L4;

% --- Plot All Three Current Waveforms on Same Graph ---
figure(2)
plot(time_L3,current_L3, time_L4,current_L4, time_L4,current_IN)
title('ALL Current Waveforms')
xlabel('Time [s]')
ylabel('Current [A]')
legend('L3','L4','IN')
xlim([-1E-5 1E-5])
```

```

% --- Math on Input Current Waveform ---
Iavg  = mean(current_IN)
Irms  = rms(current_IN)

% --- Plot Input Current Ripple ---
figure(3)
plot(time_L4,current_IN)
title('Input Current Waveform')
xlabel('Time [s]')
ylabel('Current [A]')
xlim([-1E-5 1E-5])
ylim([8.2 9.2])
% Iavg_line = reline([0 Iavg]);
% Iavg_line.Color = 'g';

% --- Find and Label: ---
% ----- Minimum, Maximum, Average, Peak-Peak, and Percent Ripple -----
indexmin = find(min(current_IN) == current_IN);
t_Imin = time_L4(indexmin(1));
Imin = current_IN(indexmin(1));
txt_Imin = ['Min = ', num2str(Imin)];
text(t_Imin,Imin,txt_Imin,'HorizontalAlignment','right');

txt_Iavg = ['Avg = ', num2str(Iavg)];
text(-0.2E-5,Imin,txt_Iavg,'HorizontalAlignment','left');

indexmax = find(max(current_IN) == current_IN);
t_Imax = time_L4(indexmax(1));
Imax = current_IN(indexmax(1));
txt_Imax = ['Max = ', num2str(Imax)];
text(t_Imax,Imax,txt_Imax,'HorizontalAlignment','right');

Ipk_pk = Imax-Imin
txt_Ipk_pk = ['Pk-Pk = ', num2str(Ipk_pk)];
text(-0.2E-5,Imin-0.03,txt_Ipk_pk,'HorizontalAlignment','left');

perc_ripple = 100*Ipk_pk/Iavg
txt_perc_ripple = ['Percent Ripple = ', num2str(perc_ripple)];
text(-0.2E-5,Imin-0.06,txt_perc_ripple,'HorizontalAlignment','left');

```

B. Altium Designer PCB Schematic

Figure B-1: Altium Designer PCB Schematic

C. Altium Designer PCB Layout

Figure B-2: Altium Designer PCB Layout – Top Layer

Figure B-3: Altium Designer PCB Layout – Bottom Layer

D. Bill of Materials

Table B-1: Bill of Materials – Totals for Standard and Modified

Altium Designator	Description	Part Number	Quantity	Manufacturer/Link
C1, C2	CAP CER 10000PF 100V X7R 1206	490-10753-1-ND	4	Murata Electronics
Cb	CAP CER 2.2UF 16V X7R 1206	490-1799-1-ND	2	Murata Electronics
Cin1, Cin2, Co2, Cml	CAP CER 10UF 50V X7R 1206	1276-6767-1-ND	9	Samsung Electro-Mechanics
Css	CAP CER 6800PF 100V X7R 1206	490-11619-1-ND	2	Murata Electronics
D1, D2	Diode Schottky 100V 10A Through Hole TO-220-2	MBR10100GOS-ND	4	ON Semiconductor
GND, M1, M2, Vin+, Vin-, X1, X2	TERM TURRET SINGLE L=5.56MM TIN	36-1502-2-ND	24	Keystone Electronics
L1, L2	10μH Shielded Wirewound Inductor 9A 16.7 mOhm Max Nonstandard	732-2145-1-ND	2	Würth Electronics
L1, L2, L3, L4	4.7μH Shielded Wirewound Inductor 15.5A 7.9 mOhm Max Nonstandard	732-2142-1-ND	4	Würth Electronics
Q1, Q2	N-Channel 60V 60A (Tc) 83W (Tc) Through Hole TO-220AB	IRF60B217-ND	4	Infineon Technologies
R1, R2	RES SMD 10 OHM 1% 1/4W 1206	1276-3513-1-ND	4	Samsung Electro-Mechanics
Rf1	RES SMD 169K OHM 1% 1/4W 1206	541-169KFCT-ND	2	Vishay Dale
Rf2	RES SMD 24.9K OHM 1% 1/4W 1206	541-24.9KFCT-ND	2	Vishay Dale
Rfs	RES SMD 100K OHM 1% 1/4W 1206	541-100KFCT-ND	2	Vishay Dale
RL1, RL2, Rshort	RES SMD 0 OHM JUMPER 1206	HCJ1206ZT0R00CT-ND	1	Stackpole Electronics
Rs1a, Rs1b, Rs2a, Rs2b	RES 0.01 OHM 1% 1/2W 1206	541-2745-1-ND	8	Vishay Dale
Rslp	RES SMD 59K OHM 1% 1/4W 1206	541-59.0KFCT-ND	2	Vishay Dale
Rvc	RES SMD 536 OHM 1% 1/4W 1206	541-536FCT-ND	2	Vishay Dale
U1	Boost Regulator Positive Output Step-Up DC-DC Controller IC 28-TSSOP-EP	LT3782AEFE#PBF-ND	2	Linear Technology/Analog Devices
Vc1	CAP CER 0.012UF 100V X7R 1206	490-16825-1-ND	2	Murata Electronics
Vc2	CAP CER 0.12UF 50V X7R 1206	478-1557-1-ND	2	AVX Corporation
Vcc	CAP CER 1UF 50V X7R 1206	490-3908-1-ND	2	Murata Electronics
Ru1	RES SMD 825K OHM 1% 1/4W 1206	P825KFCT-ND	2	Panasonic Electronic Components
Ru2	RES SMD 274K OHM 1% 1/4W 1206	P274KFCT-ND	2	Panasonic Electronic Components
n/a	Standoffs & Spacers HEX SPACER 4-40 1/2"	534-1902C	8	Keystone Electronics
n/a	Screws & Fasteners 1/4 4-40 NYLON PAN	534-9327	8	Keystone Electronics