ENERGY HARVESTING FROM ELLIPTICAL MACHINES:

DC-DC CONVERTER DESIGN

USING SEPIC TOPOLOGY

A Thesis

presented to

the Faculty of California Polytechnic State University, San Luis Obispo

In Partial Fulfillment

of the Requirements for the Degree

Master of Science in Electrical Engineering

By

Martin Kou

June 2012

© 2012
Martin Kou
ALL RIGHTS RESERVED

COMMITTEE MEMBERSHIP

TITLE: ENERGY HARVESTING FROM ELLIPTICAL
MACHINES: DC-DC CONVERTER DESIGN USING
SEPIC TOPOLOGY

AUTHOR: Martin Kou

DATE SUBMITTED: June 2012

COMMITTEE CHAIR: Dr. David B. Braun, Professor, Electrical Engineering

COMMITTEE MEMBER: Dr. Ahmad Nafisi, Professor, Electrical Engineering

COMMITTEE MEMBER: Dr. Taufik, Professor, Electrical Engineering

ABSTRACT

Energy Harvesting from Elliptical Machines: DC-DC Converter Design Using SEPIC

Topology

Martin Kou

Cal Poly's ongoing Energy Harvesting from Exercise Machines (EHFEM) project is a very convenient and cost-effective way for generating DC power from physical exercise and sending it back to the electrical grid as AC power, providing a renewable energy source for the future. The EHFEM project consists of numerous subprojects involving converting different types of exercise machines for power generation. This project is a continuation of one of the previous subprojects, specifically involving an elliptical machine, and focuses on improving system functionality at different machine settings without altering the elliptical user's experience by selecting a new DC-DC converter design, while keeping the other system components intact. The new proposed DC-DC converter design is based on a non-isolated, PWM-switching single-ended primary inductor converter (SEPIC) topology, as opposed to the resonant zero-current switching/zero-voltage switching (ZCS/ZVS) topology-based off-the-shelf DC-DC converter that the previous project utilized, which had poor system functionality at high physical input levels (greater than 30V input) from the elliptical trainer. This project proves that a PWM-switching SEPIC topology provides a functional DC-DC converter design for DC power generation and inverter interfacing from a dynamic input voltage generator because of its wide input voltage range, high power driving capability and inherent voltage step-up and step-down functions. The proposed DC-DC converter supplies up to 288 watts of power and outputs 36 volts, and simultaneously takes 5-65 volts from its input depending on the elliptical user's physical input level. This project details the new DC-DC converter's design and construction processes, compares its topology to other existing DC-DC converter topologies and analyzes unfeasible designs as well as the overall system's performance when converting the generated DC power to AC power, and documents any potential problems when used for this specific application.

Keywords: DC-DC converter, elliptical machine, energy harvesting, SEPIC, sustainability

ACKNOWLEDGEMENTS

First, I would like to thank Dr. David Braun, Dr. Dale Dolan, Dr. Ahmad Nafisi and Dr. Taufik for their help and guidance for my thesis project as well as for their guidance through my university career. I would also like to thank the following people who contributed to this project:

- Paul Grierson, who provided the Precor elliptical machine to Cal Poly's EE department.
- Sharon Kirk and Curtis Kahnberg from Enphase, who donated two Enphase micro-inverters.
- Dr. John S. Dunning, Cal Poly Mechanical Engineering Department Research Scholar, who provided access to the BK Precision XLN3640 power supply and BK Precision 8518 1.2kW electronic load used for low input voltage tests.

Furthermore, I would like to extend my thanks to my fellow EHFEM team members Alvin Hilario, Greg Hollister, Nicholas Lovgren, Michelle Lum and Jonathan Yuen for their help with my project. Additionally I'd like to thank Ben Braun, Jason Delgadillo and Juan Chavez for helping gather and provide data for my project. Finally, I would like to thank my family and friends for supporting me throughout my university career; I do not know where I would be today without them.

TABLE OF CONTENTS

	Page
LIST OF TABLES	ix
LIST OF FIGURES	xii
CHAPTER 1: INTRODUCTION.....	1
1.1 Design Requirements	3
1.2 Project Specifications.....	3
1.3 Primary Constraints	4
CHAPTER 2: PROJECT PLANNING AND ABET SENIOR PROJECT ANALYSIS	7
2.1 Project Timeline	7
2.2 Estimate of Parts Cost and Labor	9
2.3 ABET Senior Project Analysis.....	10
2.3.1 Economic	10
2.3.2 Environmental	11
2.3.3 Sustainability	12
2.3.4 Manufacturability	12
2.3.5 Ethical.....	13
2.3.6 Health and Safety	14
2.3.7 Social.....	15
2.3.8 Political.....	15
CHAPTER 3: SYSTEM COMPONENT CHARACTERIZATION AND DC-DC CONVERTER TOPOLOGY SELECTION.....	17
3.1 Precor EFX 546i Elliptical Trainer Characterization	17
3.2 Enphase M175-24-240 Micro-Inverter Specifications and Characterization	25
3.3 Converter Topology Selection.....	26
3.3.1 Comparison of Hard-Switching, Non-Isolated Topologies	29
CHAPTER 4: THE SEPIC	34
4.1 SEPIC Introduction	34
4.2 Basic SEPIC Operation	35
4.3 SEPIC Operation in Continuous Conduction Mode (CCM).....	36
4.3.1 SEPIC Operation in CCM with Switch Closed	38
4.3.2 SEPIC Operation in CCM with Switch Open.....	39
CHAPTER 5: FIRST DESIGN PHASE AND SIMULATION RESULTS	42
5.1 Basic SEPIC Design	42
5.2 PWM Controller Component Selection	50
5.2.1 Programming Thresholds with the LTC1871's RUN Pin.....	52
5.2.2 I_{TH} Pin.....	53
5.2.3 Output Voltage Programming with the FB Pin	56
5.2.4 Switching Frequency Programming with the FREQ Pin.....	56
5.2.5 MODE/SYNC Pin Operation	57
5.2.6 GND Pin	58
5.2.7 GATE Pin.....	58
5.2.8 $INTV_{CC}$ Regulator Bypassing and Operation ($INTV_{CC}$ Pin)	58
5.2.9 V_{IN} Pin.....	59
5.2.10 Current Sense Resistor Selection and SENSE Pin.....	59
5.3 Switching MOSFET and Diode Selection	60
5.4 Base SEPIC Simulation Implementation and Output Voltage Results	61
5.4.1 Base SEPIC Simulation Switching Waveform Analysis at 65V Input	62
5.4.2 Base SEPIC Efficiency Analysis	68

CHAPTER 6: SEPIC SNUBBER CIRCUIT.....	73
6.1 Snubber Introduction and Design	73
6.2 Snubber Operation Analysis	78
6.2.1 Snubber Stage 1 Operation.....	80
6.2.2 Snubber Stage 2 Operation.....	81
6.2.3 Snubber Stage 3 Operation.....	83
6.2.4 Snubber Stage 4 Operation.....	85
6.2.5 Snubber Stage 5 Operation.....	85
6.2.6 Snubber Stage 6 Operation.....	86
6.2.7 Snubber Stage 7 Operation.....	87
6.2.8 Snubber Stage 8 Operation.....	88
6.3 Snubber Component Selection.....	89
6.4 SEPIC Simulation Results with Snubber.....	92
6.4.1 Snubbered SEPIC Efficiency Analysis.....	95
CHAPTER 7: PRECOR EFX 546i OUTPUT VOLTAGE RIPPLE AND SEPIC INPUT PROTECTION	98
7.1 Precor EFX 546i Output Voltage Ripple	98
7.2 Input Overvoltage and Overcurrent Protection Circuit.....	99
7.2.1 FB, OUT, V_{CC} and SNS Pin Operation	102
7.2.2 GATE Pin.....	104
7.2.3 !SHDN Pin and Low Battery Threshold Detection	105
7.2.4 !FLT Pin	106
7.2.5 EN Pin	107
7.2.6 GND Pin	107
7.2.7 Fault Timer Control and TMR Pin	107
7.2.8 Pass Transistor Selection	109
7.3 Protected SEPIC Simulation Results.....	111
7.4 Finalized SEPIC Simulation Results	118
CHAPTER 8: THERMAL CONSIDERATIONS	129
8.1 Semiconductor Device Power Dissipation.....	129
8.2 Heatsink Thermal Resistance Ratings.....	131
8.3 Heatsink Selection and Component Temperature Recalculation	133
CHAPTER 9: PCB LAYOUT.....	140
9.1 SEPIC First Revision PCB Layout Design.....	140
9.2 SEPIC Second Revision PCB Layout Design.....	147
CHAPTER 10: HARDWARE TESTING.....	158
10.1 Component Soldering Process	158
10.2 No-Load Testing.....	160
10.3 Load Testing	164
10.3.1 Electronic Load in Constant Current Mode	164
10.3.2 Electronic Load in Constant Voltage Mode	170
10.3.3 First Troubleshooting Phase	172
10.3.3 Constant Voltage Load Tests Using Non-Current Limited Input Supply	185
10.3.4 Constant Voltage Load Tests Using Non-Voltage and Current Limited Input Supply.....	203
10.3.5 Second Troubleshooting Phase.....	207
10.3.6 SEPIC Efficiency Improvement	214
10.4 Precor Elliptical Trainer Full System Hardware Testing	234
CHAPTER 11: CONCLUSION, FUTURE IMPROVEMENTS AND FUTURE EHFEM PROJECTS.....	245
11.1 Conclusion	245
11.2 Future SEPIC Improvements	246

11.2.1 SEPIC Output Current Limiting	247
11.2.2 SEPIC PCB Size Reduction	251
11.2.3 Elliptical Trainer Battery Conservation	251
11.2.4 Improved Input Protection Schemes	253
11.3 Future EHFEM Projects	254
11.3.1 SEPIC Inductor Design	254
11.3.2 Inverter Control Mechanism	258
11.3.3 Inverter Design	259
APPENDIX A: ĆUK CONVERTER DESIGN	261
A1.1 PWM Controller Decision Matrix	262
A1.2 Non-Inverting Ćuk Converter Design using National LM5022 PWM Controller	268
APPENDIX B: INTERLEAVED BUCK-BOOST CONVERTER DESIGN	274
B1.1 Interleaved, Multi-Stage Buck-Boost Converter Design Using LT3845 and LTC3862 PWM Controllers	274
B1.2 Single-Phase Input and Output Stage Implementation	276
B1.3 Two-Phase Input Stage and Single-Phase Output Stage Implementation	286
B1.4 Four-Phase Input Stage and Single-Phase Output Stage Implementation	290
B1.5 Two-Phase Input and Output Stage Implementation	295
B1.6 Single-Phase Input and Output Stage Implementation at 500 kHz Switching Frequency	299
B1.7 Single-Phase Input Stage and Two-Phase Output Stage Implementation at 500 kHz Switching Frequency	302
B1.8 Two-Phase Input and Output Stage Implementation Analysis at 500 kHz and 350 kHz Switching Frequencies	305
B1.9 Four-Phase Input Stage and Two-Phase Output Stage Implementation	336
B1.10 Component Downsizing and Optimization Process	373
B1.11 Linear LTC6908-1 External Oscillator Configuration	381
B1.12 Test Circuit for LTC6908-1 External Oscillator	384
B1.13 Converter Cost Analysis	385
B1.14 Converter Downsize Proposal	389
B1.15 Downsized Converter Cost Analysis	393
APPENDIX C: FINAL SEPIC NETLIST	399
REFERENCES	401

LIST OF TABLES

Table	Page
Table A-1: Physical Exercise Input and Electric Output Results for Participant #1	19
Table A-2: Physical Exercise Input and Electric Output Results for Participant #2	20
Table A-3: Physical Exercise Input and Electric Output Results for Participant #3	21
Table A-4: Physical Exercise Input and Electric Output Results for Participant #4	22
 Table B-1: Finalized SEPIC Component BOM	 122
 Table C-1: SEPIC Semiconductor Component Average Steady-state Power Dissipation with 65V Input and 8A Load	 130
Table C-2: SEPIC Semiconductor Component Average Steady-state Power Dissipation with 27V Input and 8A Load	130
 Table D-1: Finalized SEPIC BOM with PCB and Heatsink Costs.....	 154
 Table E-1: SEPIC Initial Load Tests with 0.2A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	 166
Table E-2: SEPIC Initial Load Tests with 0.25A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	166
Table E-3: SEPIC Initial Load Tests with 0.3A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	166
Table E-4: SEPIC Initial Load Tests with 0.35A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	166
Table E-5: SEPIC Initial Load Tests with 0.4A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	166
Table E-6: SEPIC Initial Load Tests with 0.45A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	167
Table E-7: SEPIC Initial Load Tests with 0.5A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	167
Table E-8: SEPIC Initial Load Tests with 0.6A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	167
Table E-9: SEPIC Initial Load Tests with 0.7A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	167
Table E-10: SEPIC Initial Load Tests with 0.8A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	167
Table E-11: SEPIC Initial Load Tests with 0.9A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	167
Table E-12: SEPIC Initial Load Tests with 1A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	168
Table E-13: SEPIC Initial Load Tests with 1.5A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	168
Table E-14: SEPIC Initial Load Tests with 2A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies	168
Table E-15: SEPIC Initial Load Tests with 2.2A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	168
Table E-16: SEPIC Initial Load Tests with 2.3A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies.....	168
Table E-17: SEPIC Initial Load Tests with Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies	170

Table E-18: SEPIC Initial Load Tests with 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies	171
Table E-19: Modified Feedback and Gate Nodes SEPIC Load Tests with Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies	183
Table E-20: Modified Feedback and Gate Nodes SEPIC Load Tests with 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies.....	184
Table E-21: Modified Feedback and Gate Nodes SEPIC Load Tests with Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies.....	187
Table E-22: Modified Feedback and Gate Nodes SEPIC Load Tests with 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies	188
Table E-23: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies	190
Table E-24: SEPIC Load Test with 0.02 μ F Gate Capacitor Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies	190
Table E-25: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies	194
Table E-26: SEPIC Load Test with 0.02 μ F Gate Capacitor Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies	194
Table E-27: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Additional 470 μ F Output Capacitor, Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies.....	198
Table E-28: SEPIC Load Test with 0.02 μ F Gate Capacitor and Additional 470 μ F Output Capacitor Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies	199
Table E-29: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Additional 470 μ F Output Capacitor, Constant Voltage Electronic Load and BK 9153 Power Supplies.....	205
Table E-30: SEPIC Load Test with 0.02 μ F Gate Capacitor and Additional 470 μ F Output Capacitor, Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load and BK 9153 Power Supplies	206
Table E-31: SEPIC Load Tests with LTC1871 Current Sense Pin Filter and BK 9153 Power Supplies.....	210
Table E-32: SEPIC Load Test with LTC1871 Current Sense Pin Filter, Using 160 Strides/Min. Theoretical Input Voltages from [1] and BK 9153 Power Supplies.....	211
Table E-33: SEPIC Load Tests with LTC1871 Current Sense Pin Filter and GATE Pin Damping Filter and BK 9153 Power Supplies.....	217
Table E-34: SEPIC Load Test with LTC1871 Current Sense Pin Filter and GATE Pin Damping Filter, Using 160 Strides/Min. Theoretical Input Voltages from [1] and BK 9153 Power Supplies.....	218
Table E-35: Final SEPIC BOM.....	229
Table E-36: SEPIC Test Results with Precor Elliptical Trainer Input and Constant 10 Ω Load.....	238
Table A.A-1: PWM Controller Decision Matrix – Characteristics of Each Controller IC.....	265
Table A.A-2: PWM Controller Decision Matrix – Preference Ratings for Each Controller IC	266
Table A.A-3: PWM Controller Decision Matrix – Weighted Final Preference Ratings for Each Controller IC	267
Table B.B-1: Controller IC Components Cost for Converter	385
Table B.B-2: Transistor Components Cost for Converter	386

Table B.B-3: Diode components Cost for Converter	386
Table B.B-4: Resistor Components Cost for Converter	387
Table B.B-5: Inductor Components Cost for Converter.....	387
Table B.B-6: Capacitor Components Cost for Converter.....	388
Table B.B-7: Controller IC Components Cost for Downsized Converter	394
Table B.B-8: Transistor Components Cost for Downsized Converter	394
Table B.B-9: Diode Components Cost for Downsized Converter.....	394
Table B.B-10: Resistor Components Cost for Downsized Converter	395
Table B.B-11: Inductor Components Cost for Downsized Converter	396
Table B.B-12: Capacitor Components Cost for Downsized Converter	396

LIST OF FIGURES

Figure	Page
Figure 1.1: EHFEM System Block Diagram.....	5
Figure 2.1: Project Timeline for Spring Quarter, 2010	7
Figure 2.2: Project Timeline for Winter Quarter, 2011	7
Figure 2.3: Project Timeline for Spring Quarter, 2011	8
Figure 2.4: Project Timeline for Summer 2011.....	8
Figure 2.5: Project Timeline for Winter Quarter, 2012	8
Figure 2.6: Project Timeline for Spring Quarter, 2012	8
Figure 3.1: Elliptical DC Output Power for Participants #3 and 4 vs. Resistance Level.....	22
Figure 3.2: Basic SEPIC Layout [18].....	31
Figure 3.3: Basic Zeta Converter Layout [19].....	32
Figure 4.1: Crucial SEPIC Component Waveforms [26]	37
Figure 4.2: SEPIC Operation with S_1 Conducting [18]	38
Figure 4.3: SEPIC Operation with S_1 Turned Off [18].....	39
Figure 4.4: SEPIC Component Voltage Waveforms in CCM Operation [27].....	41
Figure 5.1: LTC1871 Internal Schematic [35]	51
Figure 5.2: LTC1871 Test Fixture Pole-Zero Compensator.....	53
Figure 5.3: Error Amplifier Equivalent Circuit	53
Figure 5.4: Error Amplifier Magnitude Response.....	54
Figure 5.5: Linear Technology's Recommended LTC1871 Base SEPIC [35].....	55
Figure 5.6: LTC1871 Controller Frequency Resistor Values and Corresponding Frequencies [35].....	57
Figure 5.7: Proposed SEPIC Base Design.....	61
Figure 5.8: Base SEPIC Output Voltage	62
Figure 5.9: Base SEPIC Crucial Component Waveforms Maximum Input Voltage, Full Load and Steady-state	63
Figure 5.10: Base SEPIC Switch Drain-Source Voltage and Inductor Voltages Maximum Input Voltage, Full load and Steady-state	66
Figure 5.11: Base SEPIC Inductor Currents at Maximum Input Voltage, 25% Load and Steady- state	67
Figure 5.12: Base SEPIC Input and Output Power at Maximum Input Voltage, Full load and Steady-state.....	68
Figure 5.13: Base SEPIC Output Voltage and Switch Power Dissipation Waveforms.....	69
Figure 5.14: Base SEPIC Switching Diode Reverse Voltage and Forward Current Stress Warning	70
Figure 5.15: Base SEPIC Switching Diode Operation in Converter Steady-State Mode.....	71
Figure 6.1: Lossless Passive Snubber Cell [41]	75
Figure 6.2: Snubber Turn-On Circuit [41].....	76
Figure 6.3: Snubber Turn-Off Circuit [41].....	76
Figure 6.4: Basic SEPIC with Lossless Passive Snubber Cell [41].....	77
Figure 6.5: Key Waveforms of Boost Converter with Lossless Passive Snubber [41]	79
Figure 6.6: Stage 1 Snubber Operation [41].....	80
Figure 6.7: Stage 2 Snubber Operation [41].....	82
Figure 6.8: Stage 3 Snubber Operation [41].....	83
Figure 6.9: Stage 4 Snubber Operation [41].....	85
Figure 6.10: Stage 5 Snubber Operation [41].....	86
Figure 6.11: Stage 6 Snubber Operation [41].....	87
Figure 6.12: Stage 7 Snubber Operation [41].....	88
Figure 6.13: Stage 8 Snubber Operation [41].....	88
Figure 6.14: SEPIC Operation During $t_0 < t < t_1$ Period.....	90
Figure 6.15: SEPIC Simulation Circuit with Passive Lossless Snubber	92

Figure 6.16: SEPIC Output Voltage and Q_1 Power Dissipation with Snubber at Full Load and Steady-state Operation	93
Figure 6.17: SEPIC Switching Diode Power Dissipation with Snubber at Full Load and Steady-state Operation	94
Figure 6.18: Snubber Diode D_4 Reverse Voltage and Forward Current Stress Warning	95
Figure 6.19: SEPIC with Snubber Input and Output Power at Full Load and Steady-state Operation.....	96
Figure 7.1: Typical LT4356-1 Application [55].....	100
Figure 7.2: LT4356-1 Internal Schematic [55].....	101
Figure 7.3: Linear Technology's LT4356-1 Sample Test Fixture Circuit	105
Figure 7.4: IXYS IXFK230N20T Safe Operating Area [57]	111
Figure 7.5: Complete SEPIC Schematic.....	112
Figure 7.6: Protected SEPIC Output Voltage and Q_1 Power Dissipation at Full Load and Steady-state Operation	114
Figure 7.7: Protected SEPIC Input Voltage, Clamped Input Voltage and $Q_{_OVP}$ Power Dissipation at Full Load and Steady-state Operation	115
Figure 7.8: Protected SEPIC Input Voltage, Clamped Input Voltage and $Q_{_OVP}$ Power Dissipation at Full Load and Steady-state Operation with 1000 μ F Input Capacitor.....	117
Figure 7.9: Finalized SEPIC Schematic Before PCB Layout.....	119
Figure 7.10: Finalized SEPIC Output Voltage and $Q_{_MAIN}$ Power Dissipation with Parasitic Component Values	126
Figure 7.11: Finalized SEPIC Converter Input and Output Power with Parasitic Component Values	127
Figure 8.1: Ohmite MV-102-55E TO-247 and TO-264 Heatsink [63]	134
Figure 8.2: Aavid Thermalloy 531202B02500G TO-202 and TO-220 Heatsink [64]	135
Figure 8.3: Ohmite WA-T247-101E TO-247 Heatsink [65]	136
Figure 8.4: Aavid Thermalloy 574502B00000G TO-220 Heatsink [66]	137
Figure 9.1: Linear Technology's Suggested LTC1871 Boost Converter Layout [35]	140
Figure 9.2: SEPIC First Revision PCB Layout	142
Figure 9.3: SEPIC First Revision PCB Layout Top Copper Layer	143
Figure 9.4: SEPIC First Revision PCB Layout Bottom Copper Layer.....	144
Figure 9.5: SEPIC First Revision PCB Layout Silkscreen Layer.....	145
Figure 9.6: SEPIC Second Revision PCB Layout	148
Figure 9.7: SEPIC Second Revision PCB Layout Top Copper Layer.....	150
Figure 9.8: SEPIC Second Revision PCB Layout Bottom Copper Layer	151
Figure 9.9: SEPIC Second Revision PCB Layout Silkscreen Layer	152
Figure 10.1: First Revision SEPIC, Top View	159
Figure 10.2: First Revision SEPIC, Angled View	160
Figure 10.3: SEPIC No-Load Test Block Diagram.....	161
Figure 10.4: SEPIC No-Load Test Wiring	162
Figure 10.5: SEPIC No-Load Test at 36V Input	163
Figure 10.6: SEPIC Initial Load Test Block Diagram.....	165
Figure 10.7: L_1 Current at 5V Input, 0.1A Load	173
Figure 10.8: L_1 Current at 18V Input, 1A Load	173
Figure 10.9: L_1 Current at 36V Input, 1.5A Load	174
Figure 10.10: L_1 Current at 50V Input, 2.3A Load	174
Figure 10.11: $Q_{_MAIN}$ Switching Trajectory at 5V Input and 0.1A Load	176
Figure 10.12: $Q_{_MAIN}$ Switching Trajectory at 36V Input and 1.46A Load.....	176
Figure 10.13: $Q_{_MAIN}$ Switching Trajectory at 50V Input and 2.3A Load	176
Figure 10.14: Finalized SEPIC Output Voltage, $Q_{_MAIN}$ Switching Trajectory, L_1 Current and Feedback and Sense Node Voltages from LTSpice Simulation.....	178

Figure 10.15: Finalized SEPIC with 0.01 μ F Feedback Capacitor Output Voltage, Q_MAIN Switching Trajectory, L ₁ Current and Feedback and Sense Node Voltages from LTSpice Simulation	180
Figure 10.16: Finalized SEPIC with 0.01 μ F Feedback Capacitor and 0.01 μ F Q_MAIN Gate Capacitor Output Voltage, Q_MAIN Switching Trajectory, L ₁ Current and Feedback and Sense Node Voltages from LTSpice Simulation.....	181
Figure 10.17: Finalized SEPIC with 0.01 μ F Feedback Capacitor and 0.01 μ F Q_MAIN Gate Capacitor Test Setup Block Diagram.....	186
Figure 10.18: Finalized SEPIC with 0.01 μ F Feedback Capacitor and 0.02 μ F Q_MAIN Gate Capacitor Output Voltage, Q_MAIN Switching Trajectory, L ₁ Current and Feedback and Sense Node Voltages from LTSpice Simulation.....	189
Figure 10.19: Q_MAIN Switching Trajectory with 0.02 μ F Gate Capacitor at 5V Input and 0.03A Load	192
Figure 10.20: Q_MAIN Switching Trajectory with 0.02 μ F at 36V Input and 2.484A Load	192
Figure 10.21: SEPIC Output Voltage Ripple at 36V Input and 2.164A Load.....	197
Figure 10.22: SEPIC Output Voltage Ripple at 36V Input and 2.231A Load With Additional 470 μ F Output Capacitor	200
Figure 10.23: Q_MAIN Switching Trajectory with 0.02 μ F Discrete Gate Capacitor and Additional 470 μ F Capacitor at 5V Input and 0.085A Load.....	201
Figure 10.24: Q_MAIN Switching Trajectory with 0.02 μ F Discrete Gate Capacitor and Additional 470 μ F Capacitor at 36V Input and 2.231A Load.....	202
Figure 10.25: Finalized SEPIC with 0.01 μ F Feedback Capacitor, Additional 470 μ F Output Capacitor and 0.02 μ F Q_MAIN Gate Capacitor Test Setup Block Diagram (Non-Voltage and Current Limited)	203
Figure 10.26: LTC1871 Current Sense Pin Switching Waveform at 36V Input and 2.28A Load	207
Figure 10.27: LTC1871 Current Sense Pin RC Low Pass Filter	208
Figure 10.28: LTC1871 Current Sense Pin Waveform With Low Pass RC Filter In Steady-State Operation.....	209
Figure 10.29: SEPIC Maximum Load Current Driving Ability vs. Input Voltage with LTC1871 Current Sense Pin Filtering	212
Figure 10.30: SEPIC Efficiency vs. Input Voltage with LTC1871 Current Sense Pin Filtering.....	213
Figure 10.31: Q_MAIN Performance Enhancement Circuit [55]	214
Figure 10.32: LTC1871 GATE Pin Damping Filter.....	215
Figure 10.33: SEPIC Maximum Load Current Driving Ability vs. Input Voltage with LTC1871 Current Sense Pin Filtering and GATE Pin Damping Filter	219
Figure 10.34: SEPIC Efficiency vs. Input Voltage with LTC1871 Current Sense Pin Filtering and GATE Pin Damping Filter	220
Figure 10.35: LTC1871 Current Sense Pin Switching Waveform with All SEPIC Performance Enhancement Modifications at 36V Input and 2.748A Load.....	222
Figure 10.36: Q_MAIN Gate Switching Trajectory with All SEPIC Performance Enhancement Modifications at 36V Input and 2.748A Load	223
Figure 10.37: SEPIC Output Voltage Ripple at 30V Input and 1.838A Load (Boost Mode)	224
Figure 10.38: SEPIC Output Voltage Ripple at 36V Input and 2.748A Load (Buck-Boost Mode).....	225
Figure 10.39: SEPIC Output Voltage Ripple at 60V Input and 7.498A Load (Buck Mode)	226
Figure 10.40: Final SEPIC Schematic.....	228
Figure 10.41: SEPIC Testing with Precor Elliptical Trainer, Angled View.....	235
Figure 10.42: SEPIC Testing with Precor Elliptical Trainer, Instrument and Grid Panel View	236
Figure 10.43: SEPIC Load Current Driving Ability vs. Elliptical Training Resistance Level with Precor Elliptical Trainer Input	240
Figure 10.44: SEPIC Efficiency vs. Elliptical Training Resistance Level with Precor Elliptical Trainer Input	240
Figure 11.1: Active Overvoltage Crowbar [72].....	248

Figure 11.2: Overcurrent Crowbar [73].....	249
Figure 11.3: DC-DC Converter Output Current Limiter Using Multiplier [53].....	250
Figure 11.4: Elliptical Trainer Battery Minimum Voltage Threshold Detection Circuit [10].....	252
Figure 11.5: L_1 Measured Inductance at 1 kHz	255
Figure 11.6: L_1 Measured ESR at 1 kHz	256
Figure 11.7: L_1 Measured Inductance at 100 kHz	256
Figure 11.8: L_1 Measured ESR at 100 kHz	257
Figure 11.9: Possible Control Scheme for Enphase M175-24-240 Micro-Inverter [10]	259
Figure A-1: Non-Inverting Cuk Converter Using National LM5022 PWM Controller	269
Figure A-2: Non-Inverting Cuk Converter Performance Characteristics at 60V Input, No Load, Using National LM5022 PWM Controller.....	270
Figure A-3: Non-Inverting Cuk Converter Performance Characteristics at 60V Input, Full Load, Using National LM5022 PWM Controller.....	272
Figure B-1: Single-Phase, Multi-Stage Buck-Boost Converter Design.....	277
Figure B-2: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 1A Load	282
Figure B-3: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 2A Load	283
Figure B-4: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 4A Load	284
Figure B-5: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 5A Load	284
Figure B-6: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load	285
Figure B-7: Two-Phase Buck, Single-Phase Boost, Multi-Stage Buck-Boost Converter Design	287
Figure B-8: Two-Phase Input, Single-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load	288
Figure B-9: Two-Phase Input, Single-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 6A Load	289
Figure B-10: Two-Phase Input, Single-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 4.75A Load	290
Figure B-11: Four-Phase Buck, Single-Phase Boost, Multi-Stage Buck-Boost Converter Design	291
Figure B-12: Four-Phase input, Single-Phase Output Buck-Boost converter Individual Stage Output Voltages at 60V Input, 8A Load	292
Figure B-13: Four-Phase Buck, Two-Phase Boost, Multi-stage Buck-Boost Converter Design	293
Figure B-14: Four-Phase Input, Two-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load	294
Figure B-15: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	296
Figure B-16: Two-Phase Input, Two-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load	297
Figure B-17: Two-Phase Input, Two-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load, Using 500 kHz Switching Frequencies for Each Stage	298
Figure B-18: Single-Phase, Multi-Stage Buck-Boost Converter Design with 500 kHz Switching Frequency for Each Stage	300
Figure B-19: Single-Phase, Multi-Stage Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load, Using 500 kHz Switching Frequencies for Each Stage.....	301
Figure B-20: Single-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	303
Figure B-21: Single-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Output Voltage with 60V Input and 8A Load.....	304

Figure B-22: Inductor Current of L_1 (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	305
Figure B-23: Inductor Current of L_4 (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	306
Figure B-24: Inductor Current of L_2 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	307
Figure B-25: Inductor Current of L_3 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	308
Figure B-26: Inductor Current of L_3 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	309
Figure B-27: Inductor Current of L_8 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	310
Figure B-28: Q_1 Drain Current (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	311
Figure B-29: Q_2 Drain Current (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	312
Figure B-30: Q_4 Drain Current (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	313
Figure B-31: Q_5 Drain Current (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	314
Figure B-32: Q_1 and Q_2 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	316
Figure B-33: Q_4 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	317
Figure B-34: Q_5 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	318
Figure B-35: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost converter Design With Improved Diodes and MOSFETs.....	320
Figure B-36: Output Voltage and Q_1 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs.....	321
Figure B-37: Zoomed-in View of Q_1 Power Dissipation Transients of Dual-Phase Buck-Boost Converter Under Maximum Input and Load.....	322
Figure B-38: Width of Largest Q_1 Power Dissipation Transient Under Maximum Input and Load	323
Figure B-39: Q_1 Drain-Source Voltage and Drain Current Under Maximum Input and Load.....	324
Figure B-40: Q_1 Drain-Source Voltage and Drain Current at Turn-On Transition Under Maximum Input and Load.....	325
Figure B-41: Q_1 Drain-Source Voltage, Gate Voltage, Gate-Source Voltage and Drain Current at Turn-On Transition Under Maximum Input and Load.....	326
Figure B-42: Q_2 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs	327
Figure B-43: Q_4 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs	328
Figure B-44: Q_5 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs	329
Figure B-45: Q_1 Gate Voltage of Dual-Phase Buck-Boost Converter Under Maximum Input and Load	330
Figure B-46: Q_4 Gate Voltage of Dual-Phase Buck-Boost Converter Under Maximum Input and Load	331
Figure B-47: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 350 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs	335
Figure B-48: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 350 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs.....	336

Figure B-49: Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 300 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs	338
Figure B-50: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 300 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs.....	339
Figure B-51: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 300 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs	340
Figure B-52: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs	341
Figure B-53: Q_1 Drain-Source Voltage, Gate Voltage, Power Dissipation and Drain Current at Turn-On Transition Under Maximum Input and Load.....	342
Figure B-54: Basic Turn-Off Snubber Circuit.....	343
Figure B-55: Basic Turn-On Snubber Circuit	344
Figure B-56: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	348
Figure B-57: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz switching frequency and Turn-On and Turn-Off Snubbers	349
Figure B-58: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	350
Figure B-59: Output Voltage, Input Current and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers.....	351
Figure B-60: Q_1 Power Dissipation, Drain-Source Voltage and Drain Current of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers.....	352
Figure B-61: Simplified Two-Phase Output Boost Converter Stage.....	353
Figure B-62: Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	354
Figure B-63: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	355
Figure B-64: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	356
Figure B-65: Q_1 Power Dissipation, Drain-Source Voltage and Drain Current of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers.....	357
Figure B-66: Q_1 Switching Trajectory at Maximum Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-On Snubbers.....	358

Figure B-67: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	359
Figure B-68: Turn-On and Turn-Off Snubber Resistor Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	360
Figure B-69: Input Current of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	361
Figure B-70: Output Voltage, Boost Stage Output Voltage and M_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	362
Figure B-71: Output Voltage, Boost Stage Output Voltage and M_2 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	363
Figure B-72: Output Voltage, Boost Stage Output Voltage, Inductor Currents, Diode Currents and M_1 and M_2 Drain Currents Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers	364
Figure B-73: Output Voltage, Boost Stage Output Voltage and M_1 and M_2 Drain-Source Voltages for Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers.....	365
Figure B-74: Output Stage for Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers.....	369
Figure B-75: Output Voltage, Buck Stage Output Voltage and M_1 Power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	370
Figure B-76: Steady-State Output Voltage, Buck Stage Output Voltage and M_1 power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	371
Figure B-77: Steady-State Input Current, Output Voltage, Buck Stage Output Voltage and M_1 Power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	372
Figure B-78: Output Voltage, Buck Stage Output Voltage, M_1 Drain Current and M_1 Power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	373
Figure B-79: Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	375
Figure B-80: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	376
Figure B-81: Output Voltage, Buck Stage Output Voltage, and M_1 Power Dissipation of Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design	377
Figure B-82: Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with ES1D Snubber Diodes	379
Figure B-83: Steady-State Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Optimized 100 kHz Two-Phase boost, Multi-Stage Buck-Boost Converter Design with ES1D Snubber Diodes	380
Figure B-84: Steady-State Output Voltage, Buck Stage Output Voltage, and M_1 Power Dissipation of Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with ES1D Snubber Diodes	381
Figure B-85: LTC6908-1 Configuration for use with the LT3845A Controllers	382
Figure B-86: Complete 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with LTC6908-1 External Oscillators	383
Figure B-87: Test Circuit for LTC6908-1 External Oscillator	384

Figure B-88: Downsized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with LTC6908-1 External Oscillators	390
Figure B-89: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Downsized 100 kHz Two-Phase Boost, Multi-stage Buck-Boost Converter Design.....	391
Figure B-90: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Downsized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design.....	392
Figure B-91: Output Voltage, Buck Stage Output Voltage and Q_1 and Q_3 Power Dissipation of Downsized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design.....	393

NOTE TO THE READER:

Thank you for your interest in the Energy Harvesting from Elliptical Machines project. When reading this report, please note that this project continues the research and work completed in the first phases of the project, Energy Harvesting from Exercise Machines Self-Generating Elliptical Machines [1] and Energy Harvesting from Elliptical Machines: DC Converter Troubleshooting [1].

CHAPTER 1: INTRODUCTION

This project provides the existing Precor elliptical trainer machine from Dr. Braun's ongoing energy harvesting from exercise machines (EHFEM) project with a functional, self-designed DC-DC converter necessary for the energy harvesting system to provide AC power for the electrical grid. Harvesting energy from physical exercise provides a sustainable energy source for the future and reduces dependence on already-scarce natural resources. In addition, it also provides monetary benefits because self-generated electricity from physical exercise reduces net energy consumption and results in utility cost savings. However, energy harvesting from physical exercise also requires additional modifications as well electronic devices for the harvesting device (in this project, an elliptical trainer). Those required additional components include a DC-DC converter as well as a grid-tie inverter. This project focuses on the DC-DC converter portion of the EHFEM energy harvesting system. DC-DC converters are widely used today in electronic applications for providing a stable DC voltage from another (usually rectified) DC voltage level. This stable DC voltage is in turn used for powering the main circuit in an electronic device, such as a computer, MP3 player or cellular phone. The main power source for the EHFEM power generation system is Precor's EFX 546i elliptical trainer, which generates a rectified DC voltage from its onboard 6-phase generator [2]. This generated DC voltage varies in magnitude in proportion to the user's physical exercise level (which includes adjustable physical resistance and incline levels), thus the generated DC voltage magnitude is unstable if the user does not maintain a constant physical exercise level. This project also selects the commercially-available

Enphase M175-24-240 micro-inverter for the EHFEM system's inverter component. The Enphase M175-24-240 micro-inverter used for taking user-generated DC voltages from the elliptical trainer and sending AC power back to the power grid requires a stable DC input voltage range for proper operation [3]. Therefore, this project designs a DC-DC converter for providing a stable DC voltage (from a user-generated DC voltage from physical exercise on the elliptical machine) as the input to a power inverter that converts its DC input voltage into a stable $240V_{\text{RMS}}$ AC voltage to feed back to the electrical grid. There are many different types of DC-DC converter circuit topologies available, such as buck (step-down), boost (step-up) and buck-boost (step-up or down). Furthermore, these topologies are broken down into other sub-categories, such as non-isolated and isolated (the latter makes use of a transformer), as well as PWM-switching converters (otherwise known as hard-switching converters) and soft-switching converters. The non-isolated, PWM-switching single-ended primary inductor converter (SEPIC) topology used in this project outputs a stable DC voltage either greater than or less than its input voltage, which is what the elliptical energy harvesting system requires for generating power from its users' variable stride rate and inputted physical training resistance level. The previous off-the-shelf resonant zero-current switching/zero-voltage switching (ZCS/ZVS) based soft-switching Vicor Maxi 28V DC-DC converter used in one of the previous elliptical projects did not function with the energy harvesting system at all at input voltage levels above 30V, because of its limited input voltage range (9-36V) [4]. This project proves that a self-designed PWM-switching SEPIC topology provides a functional DC-DC converter design for DC power generation and inverter interfacing from a dynamic

voltage generation system. Based on user exercise data in this report as well as [1], this thesis project defines in sections 1.1 and 1.2 the electrical and mechanical design constraints for its designed DC-DC converter, based on Precor elliptical user exercise data from [1].

1.1 Design Requirements

This section outlines this project's DC-DC converter's design requirements listed below:

- Conformity to UL 1741 specifications [5]
- Conformity to IEEE 1547 specifications [6]
- Conformity to PG&E safety requirements [7]
- Conformity to the National Electric Code (NEC) [8]
- No long-term costs to ASI or Cal Poly
- No change in the Precor elliptical trainer's user experience with the new system
- The system's long-term operational benefits must be greater than its total implementation costs.
- Must be functionally compatible with existing elliptical power generation system (sans current Vicor Maxi 28V DC-DC converter) at all input resistance levels
- Must meet all specifications outlined in the next section

These requirements are safety considerations as well as economic considerations for maintaining a safe exercise experience for the elliptical trainer's user as well as providing minimal operation costs to Cal Poly should it implement this power generation system.

The next section describes quantitative specifications for this project's converter.

1.2 Project Specifications

- Output voltage step-down (buck) and step-up (boost) functions
- Non-inverting, PWM-switching SEPIC circuit topology
- 8-60V nominal DC input from the elliptical trainer
- 5V minimum DC input from the elliptical trainer
- 65V absolute maximum DC input from the elliptical trainer
- 6.5A absolute maximum input current (DC) at 65V input
- 7.5A nominal maximum output current (DC) at 60V input
- 8A absolute maximum output current (DC) at 65V input
- 36V DC nominal output (feeds into the Enphase micro-inverter)

- 288W maximum power output
- Efficiency at 60V input must be $\geq 75\%$ (individual DC-DC converter)
- Line regulation when input changes from 8-60V must be $\leq 5\%$
- Peak-to-peak output voltage ripple must be $\leq 10\%$
- Both DC-DC converter inductors must run in continuous conduction mode (CCM) down to 25% maximum output load current (2A, which occurs around 31V input according to table E-34)
- Overall input impedance to the converter must be 10Ω , in order to keep the elliptical trainer's eddy current braking system intact
- Must physically fit within the Precor EFX 546i elliptical trainer's mechanical confines

In addition to the converter's requirements and specifications outlined in this section and section 1.1, section 1.3 outlines further electrical design constraints that this project's converter must satisfy.

1.3 Primary Constraints

The following constraints control and influence the converter design approach for this project:

- The DC-DC converter needs to be able to tolerate the wide input voltage range that users generate from the elliptical trainer.
- The DC-DC converter's input impedance needs to match that of the elliptical machine's output resistor coils (10Ω), in order to keep the elliptical trainer's eddy current braking system intact so that it does not alter the end-user's physical exercise experience. In Spring Quarter 2010, this project's author measured the EFX 546i elliptical trainer's resistor coils having 9.7Ω series and parallel nominal resistance, no capacitance, 0.0151 mH series inductance at 1kHz with no parallel inductance, and 0.017 mH series inductance at 120 Hz with no parallel inductance. Thus it is still possible for this project's DC-DC converter to have some additional input inductance and not affect the elliptical machine's braking system.
- The DC-DC converter must be compatible with all other existing system components, including the Enphase micro-inverter.
- The system and DC-DC converter must overall ensure safe electrical operation for the end-user, components and electrical grid, per UL, IEEE, NEC and PG&E safety requirements.

- Voltage, current and power levels in the system should never exceed individual components' absolute maximum ratings (including individual components within the DC-DC converter).

Figure 1.1 shows the EHFEM power generation system block diagram.

Figure 1.1: EHFEM System Block Diagram

From figure 1.1, at the EHFEM system's input stage, an elliptical trainer user exercises on the Precor EFX 546i and generates an unstable rectified DC voltage. That rectified DC voltage is unstable because of the user's volatile and unpredictable exercise habits and levels. That unstable DC voltage feeds into a DC-DC converter. The DC-DC converter's purpose is converting that unstable user-generated DC-voltage to a stable DC voltage level that the Enphase Micro-Inverter requires for properly converting to the 240V RMS AC voltage (at the system's output stage) that the electrical grid uses. Sections 3.1 and 3.2 describe each stage's quantitative requirements in greater detail. Like the previous elliptical energy harvesting project in [1], all system components except for the DC-DC converter are readily available. This project focuses on the second part of the block diagram in figure 1.1, however, unlike what occurred in previous projects [1], this DC-DC converter uses discrete components and is more functional with the elliptical energy harvesting system at input voltages above 30V, while still following required specifications and physical space constraints. Designing this DC-DC converter and meeting all project requirements, specifications and constraints requires project

planning, topology selection, EHFEM system characterization, design, simulation, hardware testing and troubleshooting phases.

The remainder of this report documents from this project's planning, converter topology selection, EHFEM system characterization, design, design simulation and hardware testing and troubleshooting phases, from project inception to completion. These project phases span several academic quarters. This report also provides detailed appendices documenting designs that this project deemed unfeasible based on its requirements, constraints and specifications. Those unfeasible designs contributed to this project's long timeframe and report's length, but helped narrow down a feasible DC-DC converter design for this project. The next chapter describes this project's planning phase, analyzes this project's lifecycle costs and explains how this project fits in with ABET's senior project guidelines.

CHAPTER 2: PROJECT PLANNING AND ABET SENIOR PROJECT ANALYSIS

2.1 Project Timeline

Figures 2.1 to 2.6 show this project's design timeline during each quarter that the author contributed work towards the project.

Figure 2.1: Project Timeline for Spring Quarter, 2010

Figure 2.2: Project Timeline for Winter Quarter, 2011

Figure 2.3: Project Timeline for Spring Quarter, 2011

Figure 2.4: Project Timeline for Summer 2011

Figure 2.5: Project Timeline for Winter Quarter, 2012

Figure 2.6: Project Timeline for Spring Quarter, 2012

The Gantt charts in figures 2.1 to 2.6 measure duration in days. This project required 6 quarters for completion because of the author's priority for focusing on completing non-thesis coursework. Furthermore, the project also underwent a few converter design stages that resulted in unfeasible designs for the EHFEM power generation system, adding delay to this project's completion. Refer to appendices A and B for more details on these unfeasible designs. During this project's final two quarters the author completed the printed circuit board (PCB) layout as well as soldered all required components onto the PCB for this project's DC-DC converter. Section 2.2 details this project's estimated component costs for its implementation into the EHFEM power generation system.

2.2 Estimate of Parts Cost and Labor

Monetary costs (which include lifecycle costs) as well as payback period affect this project's DC-DC converter's implementation feasibility into the EHFEM power generation system. This section provides an initial estimate of component costs for the converter.

- PWM controller for controlling transistor switching in converter - \$8-10 (for final design; samples for prototype design are free)
- DC-DC converter circuit components (Includes resistors, capacitors, Schottky power diodes and power MOSFETs) - \$60-90
- PCB (for final converter design): \$65-300, depending on number of layers, board size and trace sizes.
- Labor: ~9 hours per week (minimum); has ranged between 10 to 12.5 hours per week throughout this project's timeline. Furthermore, this labor figure also includes time lost (roughly 4-5 hours per week) due to projects and assignments for other courses outside of thesis.

In the worst case scenario, such a converter would cost \$400 (excluding the chassis and any other shielding material), which is almost twice that of Vicor Maxi 28V DC-DC

converter previously used in an earlier project [1]. Therefore, if this project's converter does end up costing that much, then its lifecycle cost must remain low and its payback period must be short. Section 10.3.6 describes the final converter component cost in greater detail. Section 2.3.1 discusses the converter's lifecycle costs as well as the EHFEM system's payback period.

2.3 ABET Senior Project Analysis

2.3.1 Economic

Economic feasibility of this project's converter requires the EHFEM system paying itself off over its lifespan and reducing the Cal Poly Recreation Center's energy costs, as using electricity costs money. Thus in the long run the project's implementation and maintenance costs must equal that of its payback value. Eventually, the project will also pay back to the school, thus the school generates profit off this project, and the total amount of money saved over such a period of time will be greater than the project's implementation and maintenance costs. Implementing such a system in a frequently-used gym such as Cal Poly's Recreation Center can also raise electricity conservation awareness and reduce overall energy consumption in the city of San Luis Obispo. Furthermore, it can also draw attention from potential donors and sponsors that may frequent Cal Poly's Recreation Center, leading to reduced overall system lifecycle costs.

According to Dr. David Braun's EHFEM project proposal in [9], for an entire power generation system with 80% DC-DC converter efficiency and 90% inverter efficiency, the added cost for modifying a single exercise machine for energy harvesting such as the Precor elliptical trainer must lie under \$360, in order to achieve a zero system

lifecycle cost after ten years operation. A zero lifecycle cost after ten years means that the EHFEM system completely offsets its initial system costs (which include implementation) after ten years. This \$360 maximum added cost value derives from a typical 100W generation figure from exercise, along with 0.5% system efficiency degradation, 12 hours per day machine usage for 41 weeks out of a year and \$0.12 per kWh electricity cost for the first year while increasing 3% annually [9]. This \$360 figure also assumes that the DC-DC converter and inverter do not require replacement over time. This project's DC-DC converter and the Enphase micro-inverter contribute to the added cost for the EHFEM system. Using a DC-DC converter with 75% efficiency further reduces this added cost allowance. According to [1], an individual Enphase M175-24-240 micro-inverter costs \$209. Therefore, this project should keep its total DC-DC converter costs at \$120 or below for meeting the \$360 maximum added system cost for achieving zero EHFEM system lifecycle cost after ten years operation. A DC-DC converter that costs \$400 in the worst case as mentioned in section 2.2 would not suffice for achieving this maximum added system cost value. However, if this project does not offset the initial system costs from usage in Cal Poly's Recreational Center after ten years, reduced campus electricity usage from electricity conservation awareness by students and faculty as well as monetary donations to Cal Poly's Recreational Center contribute as implicit benefits to Cal Poly as a result of implementing this project.

2.3.2 Environmental

Making this converter environmentally friendly requires this project designing it using the fewest materials possible (while still maintaining safety and functionality

standards) so that it does not leave a large footprint on the environment when system retirement is necessary. Additionally, the components used in this project require safe disposal when necessary. Furthermore, this project benefits the environment in the long run by using a renewable energy source (in this case, physical exercise) that does not release any harmful by-products and because that energy source is renewable, humans do not need to further tap for already scarce resources in the environment for meeting future electricity needs.

2.3.3 Sustainability

Sustainability is the main purpose of this project – this project allows for sustainability by creating electricity from physical exercise, which would otherwise be wasted. Physical exercise is always renewable and sustainable from those who use Cal Poly's Recreation Center. Powering the Recreation Center using physical exercise, allows for a reusable energy source in the future without depleting or harming any other resources.

2.3.4 Manufacturability

This project needs must be affordable to the end user, and its manufacturing cost as well as materials cost must be as low as possible, while maintaining functionality and performance requirements listed in sections 1.1-1.3. Furthermore, the converter design must be simple, easy to fabricate and be easily modified in order to allow for future modifications that may enhance efficiency or performance of the converter.

2.3.5 Ethical

As with any engineering project, this project requires its converter's design, manufacture, implementation and disposal processes conducted in an ethical manner. From the author's standpoint, this project is ethical only if everyone benefits from it in the long run. Ethical issues that arise include constructing this project's DC-DC converter using RoHS components alongside leaded components for cost reduction, as well as monetary benefits for the EHFEM system's end user and provider.

Using cheaper, leaded components in this project's DC-DC converter for reducing added system costs provides increased monetary benefits for individuals and organizations (such as gyms and fitness centers) but at the same time harms the environment when an EHFEM system component or the system itself requires retirement. Using purely RoHS components for this project's DC-DC converter reduces the available component choice and may drive up the converter's cost, increasing the system's overall payback period. Therefore, this project attempts using as many environmentally-friendly components as possible for its DC-DC converter while simultaneously maximizing operation reliability such that the converter does not require frequent replacement and disposal.

The other ethical issue that arises is that people who use the EHFEM system in organizations such as gyms and fitness centers would receive no monetary reimbursement from generating electricity from their own exercise. However, organizations that provide the exercise equipment already pay for the equipment's implementation and operation costs receive monetary reimbursement from user exercise

for covering those costs, in addition to fees such as gym membership. The users enjoy added health benefits. Thus, in that situation both the provider and user benefit from system implementation. Applying this case to Cal Poly's Recreation Center yields a similar scenario. Because this project eventually will lower the Recreation Center's operating costs, students will enjoy lower tuition costs. Also because this project helps save the school power, areas and communities with great power needs can use that saved power. Finally, the school gains environmentally-friendly power generation equipment while at the same time supporting this project for future senior project students pursuing a sustainable senior project. Therefore, everyone who this project affects benefits in the long run.

2.3.6 Health and Safety

Health and safety are crucial aspects of this project, as the various components of this project transfer and output electrical energy at lethal voltage, current and power levels. This project designs its converter in a manner such that it does not shock, burn or physically injure in any other way its end-user. Also, because this project's initial DC-DC converter prototype has components hand-soldered onto a PCB, this project must take all safety precautions into account during the soldering process. Food and liquids are also commonplace on or near gym equipment in standard gym settings, thus for future extensions of this project, the chassis for this converter must properly shield the electronic components from such dangerous disturbances.

Implementing an energy harvesting system using exercise machines such as the Precor elliptical trainer also provides health benefits for its users. Raising awareness

about generating electricity from exercise encourages people to exercise more frequently because of potential monetary benefits such as cutting electricity costs. The EHFEM system helps its users burn calories from bodily exercise. In turn, exercising more frequently leads to healthy weight maintenance and potentially reduces the risk of harmful diseases such as heart attacks, obesity and strokes.

2.3.7 Social

Gaining wide acceptance of this project's converter in the business world requires using visually pleasing and user-friendly converter components. In other words, this converter must not get in the way of the elliptical machine's user such that it alters their exercise experience. Thus this project designs its converter in such a manner. However the EHFEM system's most crucial aspect is its energy-saving ability while simultaneously providing a renewable energy source – currently “green” machines such as hybrid vehicles are among the largest social trends, particularly among the younger generation of the population. By having more institutions and gyms adopt this type of project, it attracts more people to exercise because of its “green” aspect and in turn help save more power in the long run, emphasizing the sustainability aspect of this project.

2.3.8 Political

Finding ways for the United States of America to use less energy and conserve it has been a large part of various political debates for a long time in the past, and such debates are still ongoing. Because this project uses purely human energy and not any natural resources (except for those used in producing the components for this project), this project helps institutions and businesses save money while at the same time conserve

more electrical energy, which is already a costly and scarce resource. Eventually this project would reach outside of Cal Poly's Recreation Center and expand to other institutions and businesses worldwide, helping cut down the world's dependence for scarce natural resources for energy production.

CHAPTER 3: SYSTEM COMPONENT CHARACTERIZATION AND DC-DC CONVERTER TOPOLOGY SELECTION

3.1 Precor EFX 546i Elliptical Trainer Characterization

For this project's DC-DC converter, the Precor elliptical machine's power characteristics and physical specifications, the Enphase micro-inverter's specifications, and exercise tests based on the elliptical machine's user input by typical Cal Poly students govern the converter's design requirements and specifications. This project's DC-DC converter also must not modify the end user's experience at all, with the only difference from a standard elliptical trainer machine being that it harvests and sends electrical energy back to the electrical grid. After the user-generated power flows through the DC-DC converter, the Enphase micro-inverter converts the user-generated DC voltage from the DC-DC converter and outputs a $240V_{\text{RMS}}$ AC voltage, with a maximum power of 175 watts [3], which suffices for residential or commercial applications if many of these modified machines simultaneously work in parallel.

Designing the DC-DC converter using specifications outlined in sections 1.1-1.3 first requires gathering data on how hard typical Cal Poly students exercise on an elliptical trainer. The Precor EFX 546i elliptical trainer has 20 preset physical resistance settings (corresponding to resistance levels 1-20 on the machine), as well as an incline modifier. The elliptical trainer also dumps its output power generated from user exercise into an onboard 10Ω resistor load. That power becomes heat when dumped. Any resistance change to the elliptical trainer's 10Ω load results in an altered physical resistance that the user feels when exercising, hence causing an altered user experience

[10]. Therefore this project's DC-DC converter must maintain a 10Ω input impedance during user exercise for maintaining the elliptical trainer user's exercise experience. The elliptical trainer also has a digital readout that measures the user's exercise speed in strides per minute. For the preliminary exercise tests, this project keeps the machine's incline constant while modifying the test subject's resistance levels in steps of one level. Tables A-1 to A-4 and figure 3.1 show the results of these exercise tests. For the first two exercise participants, the author could not perform current (and thus power) measurements, until he devised a method in which he could perform such measurements for the last two participants.

Table A-3-1: Physical Exercise Input and Electric Output Results for Participant #1

Test Participant #1		
Resistance Level	Exercise Rate (Strides/min.)	Output Voltage (V)
1	154	0.04
2	150	8.85
3	154	11.84
4	148	13.51
5	148	16.23
6	154	19.01
7	150	21.59
8	152	23.79
9	148	27.33
10	150	29.56
11	138	31.6
12	134	34.07
13	112	33.81
14	112	34.13
15	104	34.69
16	86	30.37
17	82	28.7
18	72	27.86
19	68	28.82
20	72	30.72

Table A-3-2: Physical Exercise Input and Electric Output Results for Participant #2

Test Participant #2		
Resistance Level	Exercise Rate (Strides/min.)	Output Voltage (V)
1	188	0.04
2	166	9.87
3	158	12.76
4	170	15.29
5	148	16.29
6	162	20.45
7	150	22.54
8	174	26.5
9	164	29.26
10	158	30.75
11	172	35.63
12	162	37.3
13	160	40.01
14	115	37.48
15	104	37.16
16	98	33.98
17	100	36.9
18	80	29.22
19	94	34.9
20	100	42.84

Table A-3-3: Physical Exercise Input and Electric Output Results for Participant #3

Test Participant #3				
Resistance Level	Exercise Rate (Strides/min.)	Output Voltage (V)	Output Current (A)	Power (W)
1	124	0.03	0	0
2	130	8.08	0.78	6.3024
3	130	10.56	1.02	10.7712
4	126	12.97	1.32	17.1204
5	130	15.66	1.53	23.9598
6	130	16.75	1.7	28.475
7	129	20.23	1.97	39.8531
8	128	21.86	2.15	46.999
9	129	25.72	2.45	63.014
10	130	27.15	2.77	75.2055
11	128	31.11	2.9	90.219
12	124	31.33	3.17	99.3161
13	132	34.81	3.56	123.9236
14	126	37.45	3.66	137.067
15	128	38.77	3.84	148.8768
16	125	40.37	4.09	165.1133
17	115	34.32	3.85	132.132
18	98	37.2	3.52	130.944
19	90	39.52	3.53	139.5056
20	80	29.81	2.81	83.7661

Table A-3-4: Physical Exercise Input and Electric Output Results for Participant #4

Test Participant #4				
Resistance Level	Exercise Rate (Strides/min.)	Output Voltage (V)	Output Current (A)	Power (W)
1	130	0.03	0	0
2	130	9.27	0.75	6.9525
3	130	11.42	1.08	12.3336
4	130	13.01	1.33	17.3033
5	130	15.43	1.56	24.0708
6	130	17.41	1.77	30.8157
7	130	20.08	2.01	40.3608
8	130	22.39	2.23	49.9297
9	130	25.09	2.42	60.7178
10	130	29.16	2.66	77.5656
11	130	30.15	2.91	87.7365
12	130	33.16	3.2	106.112
13	130	33.24	3.05	101.382
14	130	37.07	3.55	131.5985
15	130	37	3.17	117.29
16	120	33.89	2.49	84.3861
17	90	27	1.85	49.95
18	90	32	3.49	111.68
19	90	29	2.11	61.19
20	80	20	2.63	52.6

Figure 3.1: Elliptical DC Output Power for Participants #3 and 4 vs. Resistance Level

As expected, the user's physical exercise rate decreases as the elliptical machine's physical resistance level increases. Output power falloffs occur at higher training resistance levels because of the participants' inability to maintain a 120 strides per minute exercise pace. This physical resistance relies on an eddy current braking system within the elliptical machine for making a workout more difficult for the elliptical trainer's user at higher physical resistance settings, hence the lower exercise rates at higher resistance settings. However, increasing the physical resistance level helps generate more output power as long as the user exercises at a constant rate relative to lower resistance levels. In the case of participants #3 and 4, the author attempted having both participants maintain to the best of their abilities a 130 strides per minute exercise rate. Out of the four participants, the maximum output voltage obtained from the elliptical trainer was 42.84V and the maximum output current obtained was 4.09A. This corresponds to a maximum output (input to the DC-DC converter) power of 175.2W, which is well under the 288W converter output load power specification that this project set. The reason this project set a higher specification was for headroom, after analyzing exercise data from [1]'s project report (in that report, the group members reported a 350W output from the elliptical trainer on a participant performing 160 strides per minute at resistance level 20). Furthermore, that same group obtained a 60V, 6A output from the elliptical trainer when their participant exercised at 160 strides per minute at resistance level 20. Therefore, this project selects 65V as its peak input voltage – with the additional 5V as headroom. The project also selects 6.5A as its nominal maximum output current level at 60V input from the Precor elliptical trainer. The next section selects an absolute maximum output current

level. Even with these set maximum specifications, the converter requires input and output protection from voltage and current levels that exceed those specifications. For basic protection from such conditions this converter requires input and output fuses rated for their set maximum current levels. The results in tables A-1 to A-4 also show that the elliptical trainer does not begin outputting current until it reaches at least 8V output. Therefore, this project selects 5V as its minimum required input voltage for converter output regulation – the 3V from the minimum 8V input is for headroom, likewise with the maximum input voltage selection process. This project selects 8-60V as the nominal input voltage range based on exercise data from tables A-1 to A-4 as well as [1]. Also, according to a survey that same group performed, however, most Cal Poly students using the ASI Recreation Center typically exercise between resistance levels 5 to 10 at speeds between 140 to 180 strides per minute – translating to a 60-100W power generation range [1]. This range is far below the physical exercise level and consequently, the maximum electrical levels that the elliptical machine could theoretically output. Later in the DC-DC converter testing process, this project also discovered that the Precor elliptical trainer does not output a ripple-free DC voltage [10]. Section 7.1 describes this problem in greater detail and it results in minor changes to this project's DC-DC converter later on. Section 3.2 explains the reasoning for the 288W maximum converter output load power level, as well as specifications from the Enphase micro-inverter that govern this project's DC-DC converter's design.

3.2 Enphase M175-24-240 Micro-Inverter Specifications and Characterization

The current Precor EFX 546i-based EHFEM system uses an Enphase M175-24-240 micro-inverter for power inversion and sending generated power to the electrical grid. This inverter complies with FCC Part 15 Class B, UL 1741 and IEEE 1547 standards, making it safe for this project's EHFEM system [3]. This inverter also features maximum power tracking. The inverter activates its maximum power tracking feature if its input voltage is between 25 and 40V. The inverter can also tolerate up to 54V maximum input voltage, along with a recommended 8A maximum input current (though it can tolerate up to 10A maximum input short circuit current) [3]. This project uses Enphase's 8A maximum input current specification as the output current for its DC-DC converter. Therefore, this project selects 8A as its DC-DC converter's maximum output (load) current. Enphase also recommends 210W as the inverter's input power, though its datasheet does not specify a maximum input power level [3]. On its output side, the inverter also outputs 175W of power (maximum), 750mA nominal RMS current and 240V nominal RMS AC voltage (varying between 211-264V_{RMS}) at 60 Hz nominal frequency (varying between 59.3-60.5 Hz) with at least 0.95 power factor [3]. Enphase also states that the M175-24-240's peak efficiency is 95% [3]. EHFEM project member Alvin Hilario also concluded that the inverter's optimal input voltage is 36V for obtaining peak efficiency from it [10]. Therefore, this project selects 36V as its required DC-DC converter output voltage. However, the DC-DC converter still requires a maximum average input current specification. Obtaining that value requires estimating the DC-DC converter's efficiency. This project uses a conservative, 75% efficiency

estimate at full load (65V input, 8A output load current with 36V output) for its DC-DC converter. Equations (3.1) and (3.2) obtain the converter's maximum average input current.

$$\eta = \frac{P_{OUT}}{P_{IN}} = \frac{V_{OUT}I_{OUT}}{V_{IN}I_{IN}} \quad (3.1)$$

$$I_{IN} = \frac{V_{OUT}I_{OUT}}{\eta V_{IN}} = \frac{(36V)(8A)}{(0.75)(65V)} = 5.91A \quad (3.2)$$

V_{OUT} denotes the converter's average output voltage, I_{OUT} denotes its average output current, V_{IN} denotes its average input voltage, and I_{IN} denotes its average input current.

(3.2) determines this project's DC-DC converter's required maximum input current as 5.91A, but this project selects 6.5A for headroom. The 6.5A limit also helps the converter maintain a 10Ω input resistance at 65V input, helping maintain the elliptical user's exercise experience. After selecting the converter's electrical specifications, this project selects a topology for its DC-DC converter. Sections 3.3 and 3.3.1 documents this project's entire DC-DC converter topology selection process.

3.3 Converter Topology Selection

Numerous topologies exist for wide-input DC-DC conversion. However, the topology selected for this project's DC-DC converter must tolerate the requirements and specifications outlined in sections 1.1-1.3. Furthermore, the selected topology must minimize circuit complexity and component count while maintaining functionality as well as safety to both the EHFEM system and the elliptical trainer user. The previous off-the-shelf converter used by the EHFEM group in [1] shut down at input voltages above 30V and in turn caused physical resistance loss to the elliptical trainer's user, creating a

safety hazard for physical injuries. Furthermore, that same EHFEM group also reported that “at higher resistance levels, if the user decreases pace, he or she feels a large spike of physical resistance against their feet” [1]. This sudden “spike of physical resistance” not only alters the elliptical user’s exercise experience, but like the physical resistance loss phenomenon mentioned earlier, it also creates a safety hazard for physical injuries to the elliptical trainer’s user. Therefore it is crucial that this project’s DC-DC converter maintains a steady 10Ω input impedance, which prevents such abrupt physical resistance changes during user exercise. Furthermore, this project’s DC-DC converter must also function and output voltage and current throughout the specified 5-65V input range, which helps prevent sudden physical resistance loss to the elliptical trainer’s user. Also, the previous-off-the-shelf converter used by the EHFEM group in [1] peaked at 80% efficiency, therefore this project’s DC-DC converter’s efficiency must be as high as possible for minimizing its parts cost, which in term minimizes its payback period for achieving zero life-cycle cost, as outlined in section 2.3.1 [1].

Selecting an appropriate DC-DC converter topology also requires consideration for typical operation from typical user exercise levels on the elliptical trainer. As mentioned in section 3.1, typical elliptical users in the Cal Poly Recreational Center exercise between training resistance levels 5 and 10, which translates to between 15-30V typical input for the DC-DC converter. With the DC-DC converter outputting 36V, this means that the converter will typically operate in voltage step-up (boost) mode. Although this EHFEM system requires a topology with both boost and voltage step-down (buck) modes because of its 5-65V input range specification, this project puts more emphasis on

a converter topology with very efficient boost mode operation, because high efficiency in typical operation settings reduce the EHFEM system's payback period.

Firstly, DC-DC converters with both boost and buck functions are available in non-isolated and isolated topologies. Isolated topologies typically use a transformer for providing isolation between the converter's input and output stages. Such topologies include flyback, forward, push-pull, half-bridge and full-bridge DC-DC converters. Selecting an appropriate, commercially-available transformer for these topologies that fits this project's converter's electrical requirements can be difficult because of turns-ratio considerations as well as current-carrying capacity of each winding. Not having an appropriate, commercially-available transformer also resorts the project to using a custom transformer, which can significantly increase the converter's cost and control complexity. A push-pull converter, for instance, requires two switches and a transformer with four windings [11]. Half-bridge converters require a 3-winding transformer and two switches, and full bridge converters require four switches in addition to a 3-winding transformer, resulting in very complex converter circuitry and increased component count. Flyback and forward converters are the simplest of the isolated converter topologies, but have very low power density, making it unsuitable for high current output loads that this project requires [11]. Therefore, such converter topologies are unfeasible for this project. Non-isolated topology-based converters typically have a lower parts count and their operation isn't as complex as isolated converters. UL standards also dictate that operating voltages above 42.5V require isolation between input and output for preventing user access [10]. However, the Precor elliptical trainer outputs voltages above 42.5V and uses

a hefty enclosure, preventing access to internal components for typical end-users. Furthermore, the Enphase micro-inverter already contains isolation transformers between its DC input and AC output branches. Therefore, a non-isolated based converter is appropriate for this project's EHFEM system.

The second consideration in topology selection for this project's DC-DC converter is whether to use a hard-switching or resonant soft-switching topology. The Vicor Maxi 28V DC-DC converter that the previous EHFEM group in [1] used is a resonant ZCS/ZVS (soft-switching) architecture [4]. However, this converter did not function with the EHFEM system at higher input voltages and also altered the elliptical user's experience, thus making its topology unsuitable for this project's EHFEM system. Resonant soft-switching converters typically yield high efficiency [12], but typically have more complex circuitry than hard-switching converters, driving up converter cost [13]. Furthermore, a resonant soft-switching converter is not necessary for achieving high efficiencies with the typical 15-30V exercise output range [10]. Therefore, a hard-switching, non-isolated topology is the best choice for this project's EHFEM system. Section 3.3.1 compares different hard-switching, non-isolated topologies and selects the appropriate topology for this project.

3.3.1 Comparison of Hard-Switching, Non-Isolated Topologies

Perhaps the most well-known hard-switching, non-isolated DC-DC converter topologies with boost and buck functions are the buck-boost and Ćuk topologies. Their basic implementations, however, are unsuitable for this project because they output a voltage with an inverted polarity with respect to their inputs [14]. Therefore, this project

requires variations of such topologies if implemented into the EHFEM system. Obtaining a non-inverted output voltage polarity from either a buck-boost or Ćuk converter requires additional components such as switches or transformers, increasing circuit complexity and component count [15, 16]. Although Enphase's M175-24-240 datasheet does not explicitly state that the M175-24-240 micro-inverter requires a positive polarity DC voltage input, this project assumes that the datasheet's specified input voltage notation implies that the micro-inverter requires positive input DC voltage polarity [3].

Furthermore, adding switches to a DC-DC converter requires additional control circuitry, usually additional or more complex control ICs. This problem can significantly increase converter cost as well as decrease operation reliability.

This project at one point considered a four-switch buck-boost topology, as it yields a non-inverted output voltage and high efficiency [17], but its overall complexity and potential cost deemed it unsuitable for this project. Also, early on in the selection process, this project selected a PWM-switching, non-inverting Ćuk DC-DC converter topology, but simulation results uncovered severe load regulation problems during a full load condition (60-65V input, 8A load), thus making it unsuitable for this project. Furthermore, this project was also unable to find a suitable control IC that could properly drive the two switches that the non-inverting Ćuk converter required while still tolerating the 5-65V input specification. Later on, this project's author discovered that the Precor Elliptical trainer's onboard 12V battery [1] could power and bias a control IC, bypassing the 5-65V input range requirement for the control IC. However, by that time, the author proposed the DC-DC converter design that this project uses and analyzes. This project

also experimented with a non-inverting, multi-phase PWM switching cascaded buck-boost converter in simulations. Such a converter consists of a buck converter input stage and boost converter output stage connected in cascade. However, this project scrapped that design because of its circuit complexity, high component count, high cost and low efficiency (87.1% maximum) relative to its monetary cost. Refer to appendices A and B for more information on the failed non-inverting Ćuk and cascaded buck-boost designs as well as their simulation results.

After recognizing the non-inverting Ćuk and cascaded buck-boost design's unfeasibility, this project examined two underutilized hard-switching, non-isolated topologies: the single-ended primary inductor converter (SEPIC) and Zeta topologies. SEPICs and Zeta converters are essentially modified Ćuk converters, but their basic implementations output a non-inverted polarity voltage with respect to their input voltages. Figures 3.2 and 3.3 show the SEPIC and Zeta converters in their basic implementations [18, 19].

Figure 3.2: Basic SEPIC Layout [18]

Figure 3.3: Basic Zeta Converter Layout [19]

The basic SEPIC and Zeta converter implementations both use two inductors for energy exchange and outputting a non-inverted voltage, while also performing both boost and buck functions. Most implementations of both converters magnetically couple the two inductors onto a single core, saving physical PCB space as well as cost. However, section 5.1 indicates that using a magnetically coupled inductor for such converter implementations is unfeasible for this project, because of high current requirements resulting from the 8A maximum output load current requirement described in section 1.2. Using two discrete inductors on a SEPIC or Zeta converter, however, while adding cost and requiring additional physical PCB space, is easier than using an isolated topology because high-current inductors are widely available in the commercial market and have a less complex selection process than transformers. Furthermore, both converters' basic implementations maintain a low component count when compared to isolated converters or buck-boost or Ćuk converter variations. From figures 3.2 and 3.3, SEPICs and Zeta converters have identical layouts with respect to each other, except that the inductor,

switch and diode positions are different. The SEPIC's layout is more akin to that of a boost converter [20, 21], while the Zeta converter's layout more resembles that of a buck converter [22]. Section 3.3 indicated that based on elliptical user exercise data, this project's DC-DC converter will typically operate in boost mode. Therefore, this project prefers the SEPIC over the Zeta converter. Both converter topologies yield roughly the same theoretical efficiencies (~95%), with the Zeta converter having slightly higher efficiency [23, 24], but because the SEPIC functions more similarly to a boost converter and because this project's DC-DC converter will typically operate in boost mode, the SEPIC is this project's choice. Neither converter is also as efficient as a standard Ćuk converter, but the SEPIC and Zeta converters' main advantage is that they output a non-inverted voltage while still maintaining a low component count [23]. Furthermore, both converters provide electrical isolation between their inputs and outputs via a coupling capacitor, but they are not typically considered "isolated" topologies because their typical applications do not require transformers. This electrical pseudo-isolation makes it suitable for high power applications such as this project. This project selects the SEPIC as its DC-DC converter topology because of its low component count and its functionality similar to that of a boost converter. Furthermore, it only requires one switch for proper operation (and hence one control IC), thus reducing this project's DC-DC converter's overall circuit complexity. The next chapter provides a brief introduction to the SEPIC topology and its functionality.

CHAPTER 4: THE SEPIC

4.1 SEPIC Introduction

As mentioned in section 3.3.1, a SEPIC is a type of non-isolated DC-DC converter that can convert a DC voltage to another higher or lower stable DC voltage level at its output. “Single-ended” means that only one switch in the converter controls energy exchange between its components [25].

The main advantage that the SEPIC has over boost and buck-based converter topologies is that it outputs a voltage (V_O in figure 3.2) that is either higher or lower than that of its input voltage (V_{IN} in figure 3.2) [26], whereas a boost converter can only output a voltage higher than that of its input voltage, and a buck converter can only output a voltage lower than that of its input voltage. The overall layout is similar to that of a boost converter [20, 21], with overall good input current characteristics because inductor L_1 in figure 3.2 helps keep the input current waveform close to that of a DC waveform, but bad output current characteristics because diode D_1 always sends a pulsating current to the converter’s output. Essentially, a SEPIC is a cascaded boost/buck-boost converter, with its input stage similar to that of a basic boost converter, and its output stage is similar to that of a basic buck-boost converter. Overall, a SEPIC functions akin to a buck-boost converter, but has the additional advantages of having its output voltage polarity non-inverted with respect to its input voltage [26], having a true shutdown mode – i.e. when switch S_1 turns off, the converter’s output voltage reduces to 0V, and having isolation between the input and output (from coupling capacitor C_1) [25]. The capacitive isolation prevents unwanted current from flowing from the input to output

(and potentially cause premature switch failure) [27]. S_1 's duty cycle controls the SEPIC's output voltage. Typically S_1 is an electrically controlled switch, such as a power metal oxide semiconductor field effect transistor (MOSFET), power bipolar junction transistor (BJT) or insulated-gate bipolar transistor (IGBT). Its switching actions are controlled by a pulse-width modulation (PWM) or pulse-frequency modulation (PFM) controller. A PWM controller varies S_1 's duty cycle while keeping its switching frequency constant, while a PFM controller varies S_1 's switching frequency while keeping its duty cycle constant [28]. The next section describes the SEPIC's rudimentary operation.

4.2 Basic SEPIC Operation

The basic SEPIC performs DC-DC voltage conversion through energy exchange between its coupling capacitor and switching inductors (C_1 , L_1 and L_2). S_1 controls the energy exchange amount between the capacitor and inductors.

Maximizing energy exchange efficiency (and overall converter efficiency) requires this project's SEPIC design operating in continuous conduction mode (CCM) [29]. In CCM, the currents through L_1 and L_2 never reach 0A for any significant time period relative to S_1 's switching period. Discontinuous conduction mode (DCM) occurs if the currents through L_1 and L_2 remain at 0A for any significant time period relative to S_1 's switching period [25]. Operating a SEPIC in DCM yields higher efficiency at lighter current loads [30, 31], but this project's SEPIC requires operating at high current loads (> 1A). A survey performed by the previous EHFEM group in [1] in May 2009 shows that most elliptical trainer users in the Cal Poly ASI Recreational Center exercise at levels that

do not drive very light loads [1]. Therefore, operating in CCM should yield better overall converter efficiency for this project's SEPIC. A third operation mode known as boundary conduction mode (BCM), occurs when the currents through L_1 and L_2 drop to 0A but immediately rise back to their maximum values. BCM is the transition point between the CCM and DCM operation modes [29].

4.3 SEPIC Operation in Continuous Conduction Mode (CCM)

Operating the SEPIC in CCM means never letting the currents through L_1 and L_2 reduce to 0A – i.e. never letting L_1 and L_2 completely discharge. When the SEPIC reaches steady-state operation, the average voltage across C_1 will be equal to that of V_{IN} . Additionally, the average current through C_1 (I_{C1}) is 0A in steady-state, because capacitors block DC [32]. When this steady-state phenomenon occurs, L_2 is the only source of current to the output load. Thus, L_2 's average current equals that of the output load's, and is independent of V_{IN} .

In CCM, the sum of the average voltages across the SEPIC's energy storage elements (excluding input and output filter capacitors C_{IN} and C_2) equal that of the SEPIC's input voltage, as described in (4.1) [18]:

$$V_{IN} = V_{L1} + V_{C1} + V_{L2} \quad (4.1)$$

Since the average voltage across C_1 equals that of V_{IN} , V_{C1} equals V_{IN} , leading to (4.2):

$$V_{L1} = -V_{L2} \quad (4.2)$$

Under CCM in steady-state, the SEPIC's operation further splits into two operation modes: when S_1 conducts (turned on or closed) and when it does not conduct (turned off

or open). Figure 4.1 shows in greater detail the SEPIC's crucial CCM voltage and current waveforms during S_1 's conduction and non-conduction periods [26].

Figure 4.1: Crucial SEPIC Component Waveforms [26]

S_1 in figure 4.1 refers to the SEPIC's switch S_1 's gate voltage, I_{S1} refers to S_1 's drain current, I_{D1} refers to D_1 's forward current, I_{C1} refers to the current flowing through C_1 , I_{L1} refers to the current flowing through L_1 and I_{L2} refers to the current flowing through L_2 . Analyzing the SEPIC's entire operation in CCM requires analyzing it in S_1 's conduction and non-conduction modes, which the next section covers in greater detail.

4.3.1 SEPIC Operation in CCM with Switch Closed

Figure 4.2 shows the SEPIC's operation when S_1 conducts. Figure 4.1 shows the SEPIC's component waveforms in S_1 's "ON" state [18]. The green arrow denotes L_1 and switch S_1 current flow while the red arrow denotes coupling capacitor C_1 and L_2 current flow. Filter capacitors C_{IN} and C_2 are assumed to be in steady-state, thus no current flows through these two components until they discharge. Furthermore, C_{IN} and C_2 are also assumed to be large enough in capacitance such that the SEPIC's input and output ripple voltages are nearly 0V.

Figure 4.2: SEPIC Operation with S_1 Conducting [18]

When S_1 conducts during the first half-switching cycle, the current through L_1 (I_{L1}) increases in the positive direction while the current through L_2 (I_{L2}) increases in the negative direction (or decreases in the mathematical sense). Hence L_1 charges via V_{IN} , while L_2 discharges (acting as a source) through C_1 . S_1 remains closed for a short time period (assuming a fast switching frequency – 50 kHz or greater) and during this time period the instantaneous voltage across C_1 (V_{C1}) equals V_{IN} . Thus, V_{L1} and V_{L2} both equal approximately V_{IN} in magnitude. The only difference between the two voltages is that V_{L2} 's polarity is reversed (i.e. negative) because L_2 is discharging. C_1 in turn

discharges and supplies current to L_2 in order to store energy in it, so that L_2 can supply current to the output load during the next half-switching cycle when S_1 no longer conducts. During this entire half-switching cycle, diode D_1 does not conduct – in other words, it is reverse-biased, as in figure 4.1. Thus C_2 discharges and hence is the only component that helps maintain the output load current when S_1 is conducting. During the second half-switching cycle, S_1 turns off.

4.3.2 SEPIC Operation in CCM with Switch Open

Figure 4.3 shows the SEPIC's operation when S_1 does not conduct [18]. The red arrow denotes L_2 current flow while the green arrow denotes current flow through all other components (except C_{IN} and C_2).

Figure 4.3: SEPIC Operation with S_1 Turned Off [18]

At the end of one half-switching cycle, S_1 turns off. The new path for the input current is through L_1 and C_1 . Because current cannot change instantaneously through an inductor [33], I_{L1} and I_{L2} do not immediately change. Thus, I_{C1} equals I_{L1} . L_2 continues to discharge, but during this half-switching cycle it discharges into C_2 , thus turning on D_1 and supplying current (I_{L2}) to the output load. However, the direction of I_{L2} causes it to

add to the input current ($I_{L1} = I_{C1}$) that already flows to the output load. By Kirchhoff's Current Law, the current through D_1 (I_{D1}) equals:

$$I_{D1} = I_{C1} - I_{L2} \quad (4.3)$$

However, because $I_{L1} = I_{C1}$:

$$I_{D1} = I_{L1} - I_{L2} \quad (4.4)$$

Thus, when S_1 does not conduct, both L_1 and L_2 supply current to the output load. V_{IN} and L_1 charge C_1 (which discharged during the half-switching cycle when S_1 conducted), and L_2 continues discharging to the output load until S_1 conducts again at the beginning of the next half-switching cycle (when C_1 supplies current to charge L_2). By Kirchhoff's Voltage Law, the voltages across L_1 and L_2 (V_{L1} and V_{L2}) during this half-switching cycle equal:

$$V_{L1} = -V_{L2} = -V_O \quad (4.5)$$

Hence, L_1 and L_2 both discharge and supply current to the output load when S_1 does not conduct. In both S_1 conduction cases, the voltages across L_1 and L_2 are always both equal in magnitude. Their phases are both equal if the SEPIC's input voltage is less than its output voltage. If the SEPIC's input voltage is greater than its output voltage, then V_{L1} and V_{L2} are 180° out of phase with respect to one another [27]. Figure 4.4 shows this phenomenon [27].

Figure 4.4: SEPIC Component Voltage Waveforms in CCM Operation [27]

V_{Q1} in figure 4.4 refers to S_1 's gate voltage, V_{L1a} refers to the voltage across L_1 and V_{L1b} refers to the voltage across L_2 . Understanding the SEPIC's operation is crucial for designing a converter that properly operates under any required specifications. The next section describes this project's SEPIC's design and component selection process.

CHAPTER 5: FIRST DESIGN PHASE AND SIMULATION RESULTS

5.1 Basic SEPIC Design

Designing the SEPIC for proper operation under the proposed specifications requires selecting the proper component values. The crucial components in the SEPIC are inductors L_1 and L_2 , switching transistor S_1 , diode D_1 and coupling capacitor C_1 .

The input and output filter capacitors, C_{IN} and C_2 , also need to be large enough in capacitance to minimize input and output voltage ripple. One major disadvantage with the SEPIC topology is that its output voltage ripple is inherently large because a pulsating diode (D_1) connects to its output. As mentioned in sections 4.3.1 and 4.3.2, D_1 conducts when S_1 turns off and does not conduct when S_1 turns on, resulting in a pulsating current fed to the converter's output. Hence C_2 needs to be large in capacitance to effectively quell any output voltage ripple that results from D_1 's pulsating current.

For S_1 , a MOSFET is preferable over a BJT because MOSFETs offer higher input impedance and lower voltage drop across its main current path compared to BJTs [34]. Furthermore, a BJT needs to be biased with additional resistors as current differences control BJT switching, unlike MOSFETs, in which voltage differences control switching [34].

From equation (5.20), L_1 and L_2 can be wound on the same core, resulting in a single magnetically-coupled inductor. Thus, it is reasonable to use a single coupled inductor to house L_1 and L_2 , in order to allow for any potential PCB real estate savings. However, using a single coupled inductor later proves unfeasible, from the required

calculated specifications in (5.13)-(5.19) as well as physical availability of such a component meeting the required calculated specifications.

The first step in this project's design process was selecting an appropriate switching controller for S_1 . The SEPIC in this project uses PWM for switching control rather than PFM, for keeping the switching frequency constant. Varying the switching frequency in PFM leads to higher power dissipation (losses) across crucial components if the switching frequency is too high, which typically occurs if the input voltage is very low. Low input voltages (below 36V) and high switching frequencies require the elliptical trainer to supply more input current to the SEPIC and require S_1 to conduct more frequently than at high input voltages (above 36V) and low switching frequencies. This results in greater power dissipation stress across crucial components. Therefore PWM was selected as the switching control method for the SEPIC in this project. Linear Technology's LTC1871 PWM controller was selected as the switching controller for the SEPIC in this project, because of its high duty cycle capability (up to 92%) and programmable switching frequency (via a frequency set resistor) [35]. Furthermore, as with previous designs using Linear Technology's ICs (refer to Appendices A and B), LTSpice contains an LTC1871 simulation model. However, the LTC1871's main caveat is that its maximum input voltage is 36V [35]. Using the Precor elliptical trainer's onboard 12V battery for supplying the bias input voltage for the controller (which only controls S_1) while using the elliptical trainer's user-generated exercise as the input source for the SEPIC in general remedies that problem. During Winter Quarter 2012, the author measured the Precor elliptical trainer's onboard battery voltage as 12.16V.

Next, this project requires determining LTC1871's operating switching frequency and duty cycle range, as designated by (5.1) to (5.3), taken from Texas Instruments'

SEPIC design datasheet: [27] (<http://www.ti.com/lit/an/slyt309/slyt309.pdf>)

$$T = \text{One switching cycle period} = \frac{1}{100 \text{ kHz}} = 10\mu\text{s} \quad (5.1)$$

$$\alpha = \text{Duty cycle for which switch is closed} \quad (5.2)$$

$$1 - \alpha = \text{Duty cycle for which switch is open} \quad (5.3)$$

This project selects 100 kHz as the LTC1871's switching frequency as it provides a reasonable tradeoff between component sizes and component power dissipation. Higher switching frequencies result in greater component power dissipation but reduce the inductances required for L_1 and L_2 , while lower switching frequencies result in lower component power dissipation, but increase the inductances required for L_1 and L_2 . Because this project's SEPIC's design requires high power output (288W), component power dissipation is the more crucial issue. Hence, the low 100 kHz switching frequency was selected. The LTC1871 is capable of a 50 kHz minimum switching frequency, but at that level the SEPIC requires larger inductors, resulting in higher overall converter cost.

Equations (5.4) to (5.9) govern the SEPIC's operating duty cycle. V_D is D_1 's forward voltage drop when it conducts. This project's SEPIC's design calculations assume 0.7V for V_D . D_{MAX} corresponds to the SEPIC's absolute maximum duty cycle and D_{MIN} corresponds to its absolute minimum duty cycle.

$$D_{MAX} = \frac{V_{OUT} + V_D}{V_{IN(min)} + V_{OUT} + V_D} = \frac{36 + 0.7}{5 + 36 + 0.7} = 88\% \quad (5.4)$$

$$\alpha_{MAX} = D_{MAX} = 0.88 \quad (5.5)$$

$$1 - \alpha_{MAX} = 1 - 0.88 = 0.12 \quad (5.6)$$

$$D_{MIN} = \frac{V_{OUT} + V_D}{V_{IN(max)} + V_{OUT} + V_D} = \frac{36 + 0.7}{65 + 36 + 0.7} = 36\% \quad (5.7)$$

$$\alpha_{MIN} = D_{MIN} = 0.36 \quad (5.8)$$

$$1 - \alpha_{MIN} = 1 - 0.36 = 0.64 \quad (5.9)$$

Therefore, the duty cycle range for the LTC1871 PWM controller for this project's SEPIC is 36% to 88%, which is well within the LTC1871's operating specifications.

Next, this project's SEPIC's amplification factor range (the ratio between its output voltage to its input voltage) needed to be determined. The amplification factor helps simplify calculating the values rest of the converter's crucial components. Equations (5.10) and (5.11) determine the SEPIC's amplification factor range, assuming 100% converter efficiency.

$$A_{i(min.)} = \frac{V_{OUT} + V_D}{V_{IN(max.)}} = \frac{\alpha_{MIN}}{1 - \alpha_{MIN}} = \frac{0.36}{0.64} = 0.5625 \quad (5.10)$$

$$A_{i(max.)} = \frac{V_{OUT} + V_D}{V_{IN(min.)}} = \frac{\alpha_{MAX}}{1 - \alpha_{MAX}} = \frac{0.88}{0.12} = 7.33\bar{3} \quad (5.11)$$

Thus, this converter's amplification factor ranges from 0.5625 to 7.333.

The first crucial SEPIC component values that require consideration are switching inductors L_1 and L_2 . The inductances of both inductors should be large enough to keep the SEPIC operating in CCM, but not so large to cause resonance effects because of their adjacency to C_1 , C_{IN} and C_2 . Equations (5.12) to (5.19) determine the SEPIC inductors' requirements. In order to determine the proper inductance values for L_1 and L_2 , (5.12) first assumes 100% converter efficiency and that peak to peak ripple currents through L_1 and L_2 are approximately 30% of the maximum input current at the minimum input voltage. Furthermore, (5.12) also assumes that the SEPIC drives a 0.5A maximum load at its minimum input voltage (5V). ΔI_L corresponds to the inductors' peak to peak ripple current and L_1 and L_2 correspond to their inductances. I_{L1} corresponds to L_1 's peak

current and I_{L2} corresponds to L_2 's peak current. I_{L1} (5V input) corresponds to L_1 's average current at 5V SEPIC input, and I_{L1} (65V input) corresponds to its average current at 65V SEPIC input. I_{L2} (5V input) corresponds to L_2 's average current at 5V SEPIC input, and I_{L2} (65V input) corresponds to its average current at 65V SEPIC input.

$$\Delta I_L = I_{IN} \times 30\% = I_{OUT} \times \frac{V_{OUT}}{V_{IN(min)}} \times 30\% = 0.5A \times \frac{36V}{5V} \times 30\% = 1.08A \quad (5.12)$$

$$L_1 = L_2 = L = \frac{V_{IN(min)} D_{MAX}}{\Delta I_L f_{sw}} = \frac{5 \times 88\%}{(1.08A)(100 \text{ kHz})} = 40.74\mu H \quad (5.13)$$

$$I_{L1(peak)} = I_{IN(max)} + \frac{\Delta I_L}{2} = I_{IN(max)} \times \left(1 + \frac{30\%}{2}\right) = 25A \times \left(1 + \frac{30\%}{2}\right) = 28.75A \quad (5.14)$$

$$I_{L2(peak)} = I_{OUT} + \frac{\Delta I_L}{2} = I_{OUT} \times \left(1 + \frac{30\%}{2}\right) = 8A \times \left(1 + \frac{30\%}{2}\right) = 9.2A \quad (5.15)$$

$$\overline{I_{L1(5V \text{ input})}} = A_{i(max)} I_{OUT(min \text{ load})} = (7.33\overline{3})(0.5) = 3.667A \quad (5.16)$$

$$\overline{I_{L1(65V \text{ input})}} = A_{i(min)} I_{OUT(max \text{ load})} = (0.5625)(8A) = 4.5A \quad (5.17)$$

$$\overline{I_{L2(5V \text{ input})}} = \overline{I_{OUT(min \text{ load})}} = 0.5A \quad (5.18)$$

$$\overline{I_{L2(65V \text{ input})}} = \overline{I_{OUT(max \text{ load})}} = 8A \quad (5.19)$$

Winding both SEPIC inductors on the same core (magnetically coupling) halves the required critical inductance:

$$L'_1 = L'_2 = \frac{L}{2} = \frac{V_{IN(min)} D_{MAX}}{2 \Delta I_L f_{sw}} = \frac{5(0.88)}{2(1.08A)(100 \text{ kHz})} = 20.37\mu H \quad (5.20)$$

Because no non-custom commercial coupled inductor physically exists for the inductance and saturation current level that this project's SEPIC requires, this converter uses two discrete inductors. This project's SEPIC uses 60μH for both L_1 and L_2 . This inductance value is significantly higher than but not too different from the calculated critical inductance value in (5.13). This project's SEPIC uses Vishay's IHV28BZ60 power inductor for both L_1 and L_2 . The Vishay IHV28BZ60 has a 60μH nominal inductance with 10% tolerance, 85mΩ ESR and a 28A saturation current limit. While this is slightly below the 28.75A saturation requirement calculated in (5.14), no commercially available, non-custom inductor exists that meets this specification, along with a similar inductance

value ($\pm 10\mu\text{H}$). Furthermore, preventative measures later implemented in this project's SEPIC limit its peak input current to 25A. Future variations of this SEPIC design may implement inductors with higher nominal inductance and saturation current ratings. According to Vishay, the IHV28BZ60 does not lose more than 10% of its rated nominal inductance value if its core saturates [36]. This means that if the IHV28BZ60 saturates, it only loses up to $6\mu\text{H}$ from its $60\mu\text{H}$ nominal inductance value, decreasing its inductance to $54\mu\text{H}$ in the worst case scenario. This value is still well above the calculated critical inductance value from (5.13). Thus, Vishay's IHV28BZ60 is a reasonable choice for L_1 and L_2 .

The SEPIC's next crucial component is its switching MOSFET (S_1). Equations (5.21) to (5.23) determine the SEPIC's switching MOSFET's requirements. $V_{DS(\text{peak})}$ corresponds to the switching MOSFET's peak drain-source voltage, I_{S1} corresponds to its peak drain current, and $I_{S1(\text{RMS})}$ corresponds to its RMS drain current.

$$V_{DS(\text{peak})} > 1.15(V_{IN(\text{max.})} + V_{OUT} + V_D) = 1.15(65V + 36V + 0.7V) \Rightarrow 116.955V \quad (5.21)$$

$$I_{S1} = I_{L1(\text{peak})} + I_{L2(\text{peak})} = 28.75A + 9.2A = 37.95A \quad (5.22)$$

$$I_{S1(\text{RMS})} = I_{OUT} \sqrt{\frac{(V_{IN(\text{min.})} + V_{OUT} + V_D) \times (V_{OUT} + V_D)}{V_{IN(\text{min.})}^2}} = 0.5A \sqrt{\frac{(5+36+0.7) \times (36+0.7)}{5^2}} = 3.912A \quad (5.23)$$

Equations (5.24) to (5.26) determine the SEPIC's switching diode's requirements. $I_{D(\text{peak})}$ corresponds to the switching diode's peak forward current, $I_{D(\text{max.})}$ corresponds to its average forward current and V_R corresponds to its required peak reverse voltage rating.

$$I_{D(\text{peak})} = I_{S1(\text{peak})} = 37.95A \quad (5.24)$$

$$\overline{I_{D(\text{max.})}} = \overline{I_{OUT(\text{max.})}} = 8A \quad (5.25)$$

$$V_R > 1.15(V_{IN(\text{max.})} + V_{OUT}) = 1.15(65V + 36V) \rightarrow 116.15V \quad (5.26)$$

Equations (5.27) to (5.30) determine the SEPIC's coupling capacitor's requirements. C_1 corresponds to the coupling capacitor's minimum required capacitance value, $I_{C1(RMS)}$ corresponds to its RMS current rating, V_{C1} corresponds to its average voltage and ΔV_{C1} corresponds to its peak to peak voltage ripple. Equation (5.27) assumes that the coupling capacitor's peak to peak voltage ripple is 7.5% of the SEPIC's maximum input voltage (65V).

$$C_1 \geq \frac{I_{OUT} \times D_{MAX}}{\Delta V_{C1(peak-peak)} \times f_{sw}} \geq \frac{I_{OUT} \times D_{MAX}}{7.5\% \times V_{IN(max)} \times f_{sw}} \geq \frac{8A \times 0.88}{0.075 \times 65V \times 100kHz} \geq 14.44\mu F \quad (5.27)$$

$$I_{C1(RMS)} = I_{OUT} \times \sqrt{\frac{V_{OUT} + V_D}{V_{IN(min)}}} = I_{OUT} \times \sqrt{\frac{D_{MAX}}{1 - D_{MAX}}} = 8A \times \sqrt{\frac{0.88}{1 - 0.88}} = 21.664A \quad (5.28)$$

$$\overline{V_{C1}} = V_{IN} = 5V \text{ to } 65V \quad (5.29)$$

$$\Delta V_{C1} = \frac{I_{OUT} D_{MAX}}{C_1 f_{sw}} = \frac{8A(0.88)}{(15\mu F)(100kHz)} = 4.693V \quad (5.30)$$

15 μ F is the next higher available standard capacitor value, thus C_1 is 15 μ F. This project's SEPIC uses a non-polarized capacitor because the voltage across C_1 reverses polarity during each half-switching cycle. C_1 's RMS current requirement is in reality much lower than in (5.28), since at 5V input the SEPIC will not support an 8A output load.

Equations (5.31) to (5.33) determine the SEPIC's output filter capacitor's requirements. $I_{Cout(RMS)}$ corresponds to the output filter capacitor's RMS current rating, ESR refers to its equivalent series resistance rating and C_{OUT} refers to its minimum required capacitance value. Equations (5.31) and (5.32) also assume a 2% maximum output filter capacitor voltage ripple.

$$I_{Cout(RMS)} = I_{OUT} \times \sqrt{\frac{V_{OUT} + V_D}{V_{IN(min)}}} = I_{OUT} \times \sqrt{\frac{D_{MAX}}{1 - D_{MAX}}} = 8A \times \sqrt{\frac{0.88}{1 - 0.88}} = 21.664A \quad (5.31)$$

$$ESR \leq \frac{\Delta V_{OUT} \times 0.5}{I_{L1(peak)} + I_{L2(peak)}} = \frac{36 \times 0.02 \times 0.5}{28.75A + 9.2A} = 0.00949\Omega = 9.49m\Omega \quad (5.32)$$

$$C_{OUT} \geq \frac{I_{OUT} D_{MAX}}{f_{sw} \Delta V_{OUT}} = \frac{(8A)(0.88)}{100 \text{ kHz} \times 36 \times 0.02} = 97.77 \mu F \quad (5.33)$$

Again, likewise with C_1 , C_{OUT} 's RMS current requirement is in reality much lower than in (5.31), since at 5V input the SEPIC will not support an 8A output load. In order to account for high ESR levels in a real electrolytic capacitor, C_{OUT} is 500 μ F, which is over five times the minimum required capacitance value. Furthermore, adding two additional low-ESR ceramic capacitors in parallel at the SEPIC's output (10 μ F each) helps offset ESR from the large electrolytic capacitor. Therefore the SEPIC's total output capacitance is 520 μ F.

Recalculating the output voltage ripple percentage with the selected C_{OUT} value yields:

$$\% V_{OUT} \text{ ripple} = \frac{I_{OUT} D_{MAX}}{V_{OUT} C_{OUT} f_{sw}} \times 100\% = \frac{(8A)(0.88)}{36V \times 520 \mu F \times 100 \text{ kHz}} \times 100\% = 0.376\% \quad (5.34)$$

Thus increasing the C_{OUT} value to 520 μ F reduces the SEPIC's output voltage ripple to 0.376%.

Equations (5.35) and (5.36) determine the SEPIC's input filter capacitor's requirements. $I_{Cin(RMS)}$ corresponds to the input filter capacitor's RMS current rating and C_{IN} corresponds to its minimum required capacitance value.

$$I_{Cin(RMS)} = \frac{\Delta I_L}{\sqrt{12}} = \frac{1.08A}{\sqrt{12}} = 0.3118A \quad (5.35)$$

$$C_{IN} = \frac{C_{OUT}}{10} = \frac{97.77 \mu F}{10} = 9.777 \mu F \quad (5.36)$$

Likewise with the SEPIC's output filter capacitor, C_{IN} is 33 μ F (a standard capacitor value higher than the minimum calculated C_{IN}) in order to account for high ESR levels in a real electrolytic capacitor. C_{OUT} is much larger than C_{IN} because the SEPIC's output current

characteristics are much worse than that of its input's. Now that this project has selected its basic SEPIC components, it requires components for biasing its PWM controller IC.

5.2 PWM Controller Component Selection

Selecting proper values for the SEPIC's components ensures proper operation at the specified 5-65V input range. However, its PWM controller also requires proper external components in order to properly bias its internal analog and digital circuitry such that it correctly controls the SEPIC's switch. This section concentrates on selecting the proper components for the LTC1871 PWM controller used for the SEPIC's switch.

Figure 5.1 shows the LTC1871 controller's internal schematic [35].

Figure 5.1: LTC1871 Internal Schematic [35]

The MSOP-10 package LTC1871 PWM controller consists of 10 pins that require external biasing components. In numerical order, those pins are RUN, I_{TH}, FB, FREQ, MODE/SYNC, GND (signal ground), GATE, INTV_{CC}, V_{IN} and SENSE. The first pin requiring component selection is the RUN pin.

5.2.1 Programming Thresholds with the LTC1871's RUN Pin

A resistor divider network consisting of $R_{\text{RUN1_MAIN}}$ and $R_{\text{RUN2_MAIN}}$ (from figure 5.7) supplies a reference voltage (1.248V according to figure 5.1) to the LTC1871's RUN pin, establishing a turn-on threshold voltage for the controller IC. Along with the internal comparator, this circuit sets the input voltage thresholds at which the LTC1871 controller turns on and off. $V_{\text{IN(OFF)}}$ is the LTC1871's turn-off threshold and $V_{\text{IN(ON)}}$ is its turn-on threshold. The RUN pin requires a 1.248V reference voltage to turn on the LTC1871 controller, along with 100mV hysteresis on the reference voltage for noise immunity [35].

The voltage divider resistor values don't actually matter since the Precor elliptical trainer's on-board battery, a constant, 12V DC voltage source, biases the PWM controller. Therefore, the resistor values were left as the same values from Linear Technology's LTC1871 LTSpice macromodel test fixture circuit (274k Ω for $R_{\text{RUN1_MAIN}}$ and 133k Ω for $R_{\text{RUN2_MAIN}}$). Equations (5.37) and (5.38) determine the LTC1871's turn-off and turn-on input threshold voltages using Linear's test fixture values.

$$V_{\text{IN(OFF)}} = 1.248V \left(1 + \frac{R_{\text{RUN1_MAIN}}}{R_{\text{RUN2_MAIN}}} \right) = 1.248 \left(1 + \frac{274k\Omega}{133k\Omega} \right) = 3.819V \quad (5.37)$$

$$V_{\text{IN(ON)}} = 1.348V \left(1 + \frac{R_{\text{RUN1_MAIN}}}{R_{\text{RUN2_MAIN}}} \right) = 1.348 \left(1 + \frac{274k\Omega}{133k\Omega} \right) = 4.125V \quad (5.38)$$

This suffices for a 12V (DC) battery input to the LTC1871's V_{IN} pin. The next LTC1871 controller pin that requires component selection is its I_{TH} pin.

5.2.2 I_{TH} Pin

The LTC1871's I_{TH} pin functions as a compensation pin for the internal error amplifier. The LTC1871's internal error amplifier at this pin requires an external compensation circuit. Keeping the error amplifier uncompensated results in higher phase shifts (than a compensated amplifier) from various high frequency poles and zeros [37], which in turn results in unwanted harmonic oscillations. Those oscillations result in undesirable error amplifier operation, which in turn result in undesirable SEPIC MOSFET switching. Linear Technology's LTC1871 LTSpice macromodel test fixture circuit implements an external pole-zero compensation circuit [37], providing the error amplifier with frequency compensation. Figure 5.2 shows the pole-zero compensation circuit from Linear's LTC1871 LTSpice test fixture circuit.

Figure 5.2: LTC1871 Test Fixture Pole-Zero Compensator

C_3 and R_4 from figure 5.2 connect directly to the LTC1871's I_{TH} pin. Figure 5.3 shows an equivalent circuit model of the LTC1871's error amplifier using a Norton-equivalent output circuit [37].

Figure 5.3: Error Amplifier Equivalent Circuit

R_1 is the error amplifier's Norton equivalent output resistance and because the amplifier's output voltage does not vary with output current, R_1 is very large (i.e. infinite). A Thévenin-equivalent circuit connects R_1 in series with a controlled voltage source (rather than a controlled current source) and sets R_1 to 0Ω .

Figure 5.4 shows the error amplifier's magnitude response at each of its pole frequencies (f_1 , f_2 and f_3) [37].

Figure 5.4: Error Amplifier Magnitude Response

$f_{1(new)}$ in figure 5.4 corresponds to the amplifier's first pole frequency after applying compensation.

The pole-zero compensator uses a large capacitor (C_2 in figure 5.3) in series with a resistor (R_4 in figure 5.3), creating a zero frequency f_z (much larger than that of $f_{1(new)}$) that cancels out the error amplifier's second pole frequency f_2 [37]. Therefore, $f_z = f_2$. This series combination also creates an additional pole frequency (f_4) much larger than that of the created f_z [37]. The resistor value is much lower than that of the error amplifier's input resistance. C_3 's capacitance is also much smaller than that of C_2 's, lowering the amplifier's first pole frequency from f_1 to $f_{1(new)}$ [37]. The lowered first pole

frequency $f_{1(new)}$ then dominates the amplifier's magnitude response until the new, compensated cut-off frequency f_3 [37].

This compensator circuit greatly attenuates any high frequency harmonics that cause undesirable error amplifier operation and hence Linear Technology requires it for proper LTC1871 operation. The compensator's component values were left as nearly the same values from Linear Technology's LTC1871 recommended base SEPIC application circuit shown in figure 5.5 [35].

Figure 5.5: Linear Technology's Recommended LTC1871 Base SEPIC [35]

Equations (5.39) to (5.41) determine the compensator's pole frequencies using the selected components from figure 5.5 [37]. R_1 denotes the error amplifier's Norton-equivalent output resistance (which is very large).

$$f_{1(new)} \cong \frac{1}{2\pi R_1 C_{c1}} = \frac{1}{2\pi (R_1 \text{ large})(6800pF)} \cong 0 \text{ Hz} \quad (5.39)$$

$$f_2 = f_z = \frac{1}{2\pi R_C C_{c1}} = \frac{1}{2\pi (33k)(6800pF)} = 704.974 \text{ Hz} \quad (5.40)$$

$$f_4 \cong \frac{1}{2\pi R_C C_{c2}} = \frac{1}{2\pi (33k)(47pF)} \cong 102 \text{ kHz} \quad (5.41)$$

The selected compensator values suffice at blocking out higher frequency harmonics because attenuation begins at close to DC. Therefore these components suffice for

biasing the LTC1871's I_{TH} pin for this project's SEPIC. The next pin, FB, determines the SEPIC's output voltage and also requires external components.

5.2.3 Output Voltage Programming with the FB Pin

Regulating S_1 's duty cycle with the SEPIC's input voltage requires the LTC1871 controller knowing the SEPIC's set output voltage. A feedback voltage divider resistor network connected to the SEPIC's output sets the SEPIC's output voltage. The pull-down resistor (R_{FB2} in figure 5.7) provides the 1.23V that the LTC1871's FB pin requires for properly regulating the SEPIC's output at the desired voltage. Equation (5.42) determines the SEPIC's output voltage.

$$V_{OUT} = 1.23V \left(1 + \frac{R_{FB1}}{R_{FB2}} \right) \quad (5.42)$$

R_{FB2} is 4k Ω , thus a 36V SEPIC output requires the following resistance value for R_{FB1} :

$$R_{FB1} = R_{FB2} \left(\frac{V_{OUT}}{1.23} - 1 \right) = 4k\Omega \left(\frac{36}{1.23} - 1 \right) = 113.07k\Omega \quad (5.43)$$

R_{FB1} is 115k Ω , as it is the next highest commercially available standard resistor value.

After selecting components that determine the SEPIC's output voltage, the LTC1871 controller needs components to set the SEPIC's switching frequency.

5.2.4 Switching Frequency Programming with the FREQ Pin

Figure 5.6 below, taken from the LTC1871's datasheet, shows the relationship between the LTC1871's switching frequency for S_1 and the chosen frequency set resistor (R_T) [35]:

Figure 5.6: LTC1871 Controller Frequency Resistor Values and Corresponding Frequencies [35]

R_T is the same as R_{FREQ} in this project's SEPIC schematic in figure 5.7. For a 100 kHz switching frequency, a 220kΩ resistor suffices for R_{FREQ} , as it is a commonly available value. According to figure 5.6 it also yields a switching frequency very close to that of the desired 100 kHz. The next step in biasing the LTC1871 controller after selecting its frequency set resistor is controlling its MODE/SYNC pin.

5.2.5 MODE/SYNC Pin Operation

Leaving the MODE/SYNC pin open (unconnected) or connected to ground pulls its voltage low and enables burst-mode operation. The internal 50kΩ internal resistor connected to the MODE/SYNC pin (in figure 5.5 above) pulls it low. Tying the MODE/SYNC pin to a voltage source above 2V or an external clock disables LTC1871 burst-mode operation and enables continuous mode operation. This project's SEPIC desires continuous PWM operation, as burst mode is desirable only if the SEPIC drives a

very light load (i.e. low output current) for most of its operation. Thus, this project's SEPIC directly connects the LTC1871's MODE/SYNC pin to the INTV_{CC} pin to provide a 5.2V source ($> 2V$) for the LTC1871 to operate in continuous mode. Burst mode reduces PWM frequency at light loads and maximizes efficiency at light loads [LTC1871 datasheet]. Operating the SEPIC in burst mode is similar to operating it in DCM, and as mentioned in section 4.2 DCM is only desirable for light loads. Therefore, continuous LTC1871 controller operation yields better overall SEPIC efficiency over burst mode operation. The next section discusses proper LTC1871 controller grounding.

5.2.6 GND Pin

The LTC1871's GND pin connects directly to the SEPIC's ground node. Proper, low-noise operation however, requires it connected to a signal ground plane separate from that of a power ground plane. Chapter 9 (PCB Layout) discusses this requirement in greater detail.

5.2.7 GATE Pin

The LTC1871's GATE pin feeds the switching signal into S_1 's gate and hence connects directly to S_1 's gate. The next section discusses proper INTV_{CC} pin biasing.

5.2.8 INTV_{CC} Regulator Bypassing and Operation (INTV_{CC} Pin)

The LTC1871's INTV_{CC} regulator is a 5.2V voltage supply that powers the LTC1871's gate driver and logic circuitry shown in figure 5.1. This regulator supplies up to 50mA of current. At the very minimum, the INTV_{CC} regulator requires a physically adjacent 4.7 μ F capacitor connected to LTC1871's INTV_{CC} pin. This pin requires this capacitor for bypassing the INTV_{CC} regulator to ground from the high transient currents

that the LTC1871's gate driver requires. The SEPIC schematic in figure 5.7 denotes this bypass capacitor as C_{VCC} . After selecting components for biasing the $INTV_{CC}$ pin, biasing the LTC1871's V_{IN} pin is the next crucial step in the LTC1871 component selection process.

5.2.9 V_{IN} Pin

The LTC1871's V_{IN} pin connects to the Precor elliptical trainer's onboard 12V battery's positive terminal. By default its desired terminal is the SEPIC's input (which is directly from the elliptical trainer's output), however because the LTC1871's V_{IN} pin can only tolerate up to a 36V maximum input voltage, it is instead connected to the elliptical trainer's 12V battery. After connecting the LTC1871's V_{IN} pin, the last pin that requires component selection is the SENSE pin.

5.2.10 Current Sense Resistor Selection and SENSE Pin

The current sense resistor (denoted as R_{SNS} in the SEPIC's schematic in figure 5.7) limits the peak current flowing through S_1 . It is not necessary for low power applications (in which V_{IN} is less than 36V, according to Linear Technology's LTC1871 datasheet [35]) because the LTC1871 controller determines S_1 's drain current limit using S_1 's drain-source on resistance ($R_{DS\ ON}$). However, adding a discrete sense resistor further reduces S_1 's drain current stress by adding resistance. S_1 's peak calculated current is 37.95A from (5.22), but requires limiting to 30A in order to keep L_1 from saturating. The LTC1871's SENSE pin operates on a 150mV threshold. Equation (5.44) determines the current sense resistor value required for limiting S_1 's peak drain current to 30A. V_{SNS} denotes the LTC1871's SENSE pin voltage.

$$R_{SNS} = \frac{V_{SNS}}{I_{D(S1,MAX.)}} = \frac{150mV}{30A} = 0.005\Omega = 5m\Omega \quad (5.44)$$

Selecting the R_{SNS} value completes the LTC1871 biasing component selection process.

Before beginning the simulation process however, this project's SEPIC's design requires simulation models for S_1 and D_1 . The next section describes the selection process for S_1 and D_1 .

5.3 Switching MOSFET and Diode Selection

Maximizing converter efficiency requires selecting a power MOSFET with low gate charge and a low drain-source on resistance ($R_{DS\ ON}$). A low $R_{DS\ ON}$ is desirable because it helps minimize I^2R losses across the MOSFET when high current flows through its drain. Low gate charge is also desirable because less gate charge results in less energy (and hence power) lost across the switching MOSFET during turn-on and turn-off when its total gate capacitance charges and discharges [38]. Furthermore, this project's SEPIC's switching MOSFET also needs to be able to withstand the peak drain-source voltage and drain current requirements outlined in (5.21) to (5.23). Infineon's IPP110N20N3 N-channel power MOSFET satisfies all the above requirements and furthermore, LTSpice has a simulation model available for it. The IPP110N20N3 power MOSFET has a 200V maximum V_{DS} rating, 11m Ω maximum $R_{DS\ ON}$ rating and 65nC total gate charge [39]. It can also withstand up to 88A continuous drain current and 352A pulsed drain current at 25°C ambient operating temperature and 63A continuous current at 100°C ambient operating temperature [39]. Because it satisfies the requirements outlined in (5.21) to (5.23) and also has a low gate charge and low $R_{DS\ ON}$, Infineon's IPP110N20N3 is a suitable switching MOSFET for this project's SEPIC.

For this project's SEPIC's switching diode, the MBR20100CT Schottky diode seems suitable. The MBR20100CT has the highest combined peak reverse voltage and average forward rectified current ratings out of all the pre-packaged diode models available in LTSpice, hence its selection. It also has a 100V peak reverse voltage rating and a 10A (for a single diode) average forward rectified current rating. However, as calculated in (5.25) and (5.26), while the MBR20100CT's average forward rectified current rating is sufficient for the switching diode's required 8A average forward current, its peak reverse voltage rating is not adequate. Using two MBR20100CTs connected in series is thus more feasible, as doing so decreases its peak reverse voltage. However, for initial simulation purposes, only one is used. After selecting its switching MOSFET and diode, the base SEPIC design is ready for simulation.

5.4 Base SEPIC Simulation Implementation and Output Voltage Results

Figure 5.7 shows the base SEPIC design schematic after selecting the proper components based on the criteria from sections 5.1 to 5.2.10.

Figure 5.7: Proposed SEPIC Base Design

The initial base SEPIC design is ready for simulation. This SEPIC's first characteristic requiring investigation is its output voltage. Figure 5.8 shows the base SEPIC's output

voltage in green. This first simulation runs the SEPIC under a full load condition. This project specifies full load condition as 65V input while driving an 8A load.

Figure 5.8: Base SEPIC Output Voltage

Average output voltage is 36.577V and peak-to-peak ripple voltage is 0.162V, corresponding to a 0.44% peak-to-peak output voltage ripple. This initial base SEPIC design outputs the required 36V average voltage specified and meets the output voltage ripple specifications outlined in section 1.2. This report's next subsection analyzes other crucial node waveforms in this initial SEPIC design for ensuring that they closely match those in figure 4.1 and [26, 27]. Having matching simulation waveforms as from figure 4.1 and [26, 27] ensures that this initial SEPIC design functions correctly.

5.4.1 Base SEPIC Simulation Switching Waveform Analysis at 65V Input

Other crucial SEPIC operation waveforms besides output voltage also require inspection for ensuring proper converter operation. These waveforms include Q_1 's

switching trajectory and its drain current ($I_d(Q_1)$), current through switching inductors L_1 and L_2 (I_{L1} and I_{L2}), voltage across and current through coupling capacitor C_5 ($V(N003)$ - $V(N004)$ and I_{C5}) and switching diode current (I_{D1}). Figure 5.9 shows these waveforms with the base SEPIC running in full load and steady-state operation. The light green waveform denotes SEPIC output voltage, blue denotes Q_1 gate voltage (switching trajectory), dark green denotes Q_1 drain current, red denotes I_{L1} , teal denotes I_{L2} , $V(N003)$ - $V(N004)$ denotes coupling capacitor voltage, magenta denotes coupling capacitor current, and gray denotes switching diode current.

Figure 5.9: Base SEPIC Crucial Component Waveforms Maximum Input Voltage, Full Load and Steady-state

Each waveform is the proper shape according to figure 4.1 and [26, 27]. The SEPIC's switching period determined from the Q_1 switching trajectory waveform in figure 5.9 is 9μs, which corresponds to a 111 kHz switching frequency. This switching frequency suffices, as it is close enough to the 100 kHz desired SEPIC switching frequency when using a commonly available 220kΩ resistor as the LTC1871 controller's frequency set

resistor. At the SEPIC's specified 65V maximum input voltage and 8A maximum load while in steady-state operation, peak drain current through Q_1 is 16.39A, peak current through L_1 is 6.648A, peak current through L_2 is 10.3A, peak voltage across the coupling capacitor is 67.8V, peak current through the coupling capacitor is 9.93A, and peak switching diode current is 16.22A. Average switching diode current is 8.011A, which nearly matches that of the SEPIC's average load current. The coupling capacitor's ripple voltage is 2V peak-to-peak and its RMS current is 6.292A. Finally, average drain current through Q_1 is 4.8A, average current through L_1 is 4.42A and average current through L_2 is 8.4A. Furthermore, neither the current through L_1 nor the current through L_2 decrease to 0A during the entire SEPIC steady-state operation period, thus this SEPIC operates in CCM at full load. The L_1 and L_2 current waveforms are both also perfect triangular waves, indicating that neither inductor saturates in this SEPIC's steady-state operation.

Overall, these voltage and current values do not place any major electrical stress on the crucial SEPIC components, provided that the physical components can tolerate these electrical stress levels. Of course, all passive components in this base SEPIC simulation are ideal (with no ESR), except for the 85m Ω ESR set in L_1 and L_2 's parameters. Furthermore, at 65V input, the SEPIC only operates at 36% duty cycle with a 0.5625 voltage gain from input to output. Its input current (from the elliptical trainer) is lower than that of the output load current. Assuming that the output load current remains constant at 8A, lower SEPIC input voltages require higher input current and thus place higher electrical stress on components connected to the SEPIC's input node – specifically

Q_1 and L_1 . Thus, these voltage and current values increase at lower SEPIC input voltages and hence are not the highest overall values facing the SEPIC components.

Other waveforms that require analysis are the SEPIC's switch drain-source voltage and inductor voltages. Figure 5.10 shows these waveforms with the base SEPIC running in steady-state operation. The light green waveform denotes SEPIC output voltage, blue denotes Q_1 gate voltage (switching trajectory, the magenta waveform and $V(N003)$ - $V(SENSE)$ denote Q_1 drain-source voltage, the red waveform and $V(IN)$ - $V(N003)$ denote L_1 voltage V_{L1} and the teal waveform and $V(N004)$ denote L_2 voltage V_{L2} .

Figure 5.10: Base SEPIC Switch Drain-Source Voltage and Inductor Voltages Maximum Input Voltage, Full load and Steady-state

With 65V input, the SEPIC's input voltage is greater than its output voltage, resulting in V_{L1} and V_{L2} 180° out of phase with each other, as specified in [27]. If the SEPIC's input voltage is less than its output voltage, then V_{L1} and V_{L2} would both be in phase with each other (i.e. 0° phase relative to each other). Peak switch drain-source voltage is 105.63V, peak L_1 voltage is 64.47V and peak L_2 voltage is 37.16V. L_1 and L_2 's minimum voltages are the antipodes of each other's peak voltages – minimum L_1 voltage is -39.74V and minimum L_2 voltage is -66.45V. Average switch drain-source voltage is 64.91V and average L_1 voltage is near 0V. Average L_2 voltage is -1.125V. L_2 's average voltage is not quite close to 0V because as mentioned in sections 4.3.1 and 4.3.2, L_2 alternatively charges (via coupling capacitor C_5) and discharges when the SEPIC's switch conducts and constantly discharges to the load when the SEPIC's switch does not conduct.

As specified in section 1.2, this project's SEPIC must maintain CCM operation down to 25% of full load at its output, i.e. a 2A load at its output. Thus, the remaining

waveforms that require analysis are the L_1 and L_2 current waveforms when the SEPIC operates with a 2A load at its output. Figure 5.11 shows these waveforms with the base SEPIC running in steady-state operation with 65V input. The green waveform denotes SEPIC output voltage, blue denotes L_1 current I_{L1} and red denotes L_2 current I_{L2} .

Figure 5.11: Base SEPIC Inductor Currents at Maximum Input Voltage, 25% Load and Steady-state

Minimum L_1 current is -1.5A and minimum L_2 current is 0.775A. The current through both SEPIC inductors never reaches 0A for any significant time period relative to the SEPIC's 9µs switching period, thus this base SEPIC meets the specification noted in section 1.2 requiring CCM converter operation down to 25% of output full-load current. After confirming that the base SEPIC outputs the correct voltage and that all crucial waveforms are the proper shape, the next characteristic requiring investigation is its overall efficiency.

5.4.2 Base SEPIC Efficiency Analysis

The ratio between the SEPIC's output power to its input power determines its overall efficiency. Higher efficiency often translates to lower power losses across components and less stress on those components. Lower efficiency means that the SEPIC needs to draw more current from its voltage source (the Precor Elliptical trainer) in order to deliver the same amount of power to the load than if the efficiency was higher. Higher efficiency is more desirable than lower efficiency because it results in lower EHFEM system lifetime costs, as outlined in section 2.3.1. Therefore, this project also focuses on maximizing efficiency on the SEPIC design as much as possible. Figure 5.12 shows the base SEPIC's steady-state input and output power waveforms at 65V input and with an 8A load. The green waveform denotes output power and the blue waveform denotes input power.

Figure 5.12: Base SEPIC Input and Output Power at Maximum Input Voltage, Full load and Steady-state

The base SEPIC's average output power at 65V input and with an 8A load is 292.62W and average input power is 328.11W. Thus, the overall converter efficiency at full load and steady-state is 89.2%. While this overall converter efficiency falls within the SEPIC specification described in section 1.2, operating at near 100% converter efficiency is more desirable. The components in a real SEPIC are far less ideal than those used in this simulation model. This translates to lower overall converter efficiency on a real SEPIC. Therefore, an investigation on power dissipation (loss) in this SEPIC is necessary. One major component in which such losses occur is in the SEPIC's switching MOSFET, denoted as Q_1 in the schematic in figure 5.7.

Figure 5.13 shows the base SEPIC's steady-state output voltage and power dissipation across Q_1 at 65V input and with an 8A load. The green waveform denotes the base SEPIC's output voltage and the blue waveform denotes Q_1 's power dissipation.

Figure 5.13: Base SEPIC Output Voltage and Switch Power Dissipation Waveforms

Average power loss across Q_1 at full load and steady-state is 2.862W, which is less than 1% of the SEPIC's 288W output. However, the 1.6kW power transients are undesirable for Q_1 for prolonged SEPIC operation, as these high power transients reduce Q_1 's operating life and that of the overall converter as well. Thus this base SEPIC requires additional circuitry to reduce the high power transients and power dissipation losses across Q_1 .

After running the simulation, the LTSpice simulation program also returned a warning message shown in figure 5.14 informing that D_1 's reverse voltage and average forward current reached near their limits during SEPIC steady-state operation.

Figure 5.14: Base SEPIC Switching Diode Reverse Voltage and Forward Current Stress Warning

Editing D_1 allows for viewing the diode's operating details in the SEPIC's steady-state operation mode. Figure 5.15 shows these details.

Figure 5.15: Base SEPIC Switching Diode Operation in Converter Steady-State Mode

D_1 's peak reverse voltage reaches 104V, which is above its 100V maximum rating [40]. Furthermore, D_1 also dissipates 6.08W, requiring very effective (and potentially expensive) cooling on the diode. Therefore using one MBR20100CT switching diode in this SEPIC is not sufficient for prolonged SEPIC operation.

Overall, without any further design modifications this initial base SEPIC design is not feasible for implementation into a high-power energy conversion system such as the EHFEM project. The power transients across Q_1 reduce its operating life, requiring frequent converter servicing and component replacement throughout the SEPIC's lifetime if implemented into the EHFEM system. Furthermore, using one MBR20100CT for this project's SEPIC's switching diode is also unfeasible since its reverse voltage rises above

its maximum rated value. Therefore, this SEPIC requires additional circuitry that reduces the power transients (and hence overall dissipation losses) across Q_1 as well as additional series-connected MBR20100CT diodes, as prolonged operation is preferable to reduce overall EHFEM system lifetime costs. A snubber cell is one type of circuit that helps reduce switching power transients and overall dissipation losses in a DC-DC converter [41].

CHAPTER 6: SEPIC SNUBBER CIRCUIT

6.1 Snubber Introduction and Design

As determined in section 5.4.2, this project's SEPIC's switch theoretically dissipates 2.862W with 1.6kW transients at 65V input while driving an 8A load, which is an unreasonable power dissipation stress on the switch even with a heatsink attached to it. Even with a proper cooling mechanism those high power transients may destroy the switch after prolonged SEPIC operation. Furthermore, if this SEPIC drives an 8A load at lower input voltages, the SEPIC's switch's drain current increases because the load requires more power from the input source (the Precor elliptical trainer) in order to maintain a steady 36V SEPIC output and 8A load current. This drain current increase results in even greater power dissipation and higher power transients across the SEPIC's switch than at maximum input voltage and maximum load current. Thus, this project's SEPIC requires a mechanism for reducing such power dissipation stress and losses. Such losses usually result from the switch's drain-source voltage not decreasing to 0V fast enough when the switch turns on and current begins to flow through its drain (resulting in nonzero $V \times I$ losses) as well as its drain current not decaying quickly enough to 0A when the switch turns off and its drain-source voltage begins increasing. Thus, this project's SEPIC requires an additional circuit that slows down switch drain current rise during switch turn-on and drain-source voltage rise during turn off, such that the switch turns on when its drain-source voltage reaches 0V and turns off when its drain current reaches 0A. Numerous circuits can perform this task, such as a snubber cell or an active clamp [41]. The purpose of adding either of these circuits to the SEPIC is reducing the power

dissipation stress on its switch as well as electromagnetic interference (EMI) noise from it. This in turn reduces overall power dissipation losses in the converter and improves overall converter efficiency.

A snubber cell, or simply known as a snubber, is more feasible for this project's SEPIC over an active clamp circuit because unlike an active clamp circuit, a snubber can be created from purely passive components and does not require any additional switches (such as MOSFETs), which in turn require additional control circuitry [42-45].

There are many available snubber topologies – some use purely passive components while others employ both passive and active components, known as passive snubbers and active snubbers. Like active clamp circuits, active snubbers require additional switches and those switches require additional control circuitry, adding to the SEPIC's complexity and overall cost. Furthermore, switch control synchronization problems are a likely occurrence if more than one switch is used in an active snubber [41]. Passive snubbers, on the other hand, have simple structures and are thus easy for implementing into a DC-DC converter, while also having a low cost [46]. Passive snubbers are further broken into two sub-categories – those that employ resistors and those that do not. The most well-known type of snubber that employs resistors is a resistor-capacitor-diode (RCD) snubber [46]. However, RCD snubbers and resistive snubbers in general have poor switch power dissipation reduction performance because they incur additional power dissipation losses in their resistors (I^2R losses) [46].

An alternative passive snubber topology instead replaces resistors with an inductor, thus employing only “theoretically lossless” components [47-51]. The main

advantages of this snubber topology are its low cost, high reliability and its effective performance in reducing switch power dissipation stress and losses. Thus, the snubber used in this project's SEPIC employs this lossless passive component topology, reducing component costs and snubber power dissipation losses. This snubber circuit only uses inductors, capacitors and diodes. Figure 6.1 shows this lossless passive snubber used in this project's SEPIC [41].

Figure 6.1: Lossless Passive Snubber Cell [41]

Node A denotes the wire connection to the SEPIC's main switching diode (D_1)'s anode, while K denotes D_1 's cathode location. Node A' corresponds to D_2 's anode. This snubber is actually a combination of two sub-circuits – one that reduces power dissipation across the SEPIC's switch during its turn-on stage and another that performs the same task during the SEPIC's switch's turn-off stage. These sub-circuits are the snubber's turn-on and turn-off circuits. Figures 6.2 and 6.3 show the snubber's turn-on and turn-off circuits [41].

Figure 6.2: Snubber Turn-On Circuit [41]

Figure 6.3: Snubber Turn-Off Circuit [41]

Combining the circuits in figures 6.2 and 6.3 results in the complete snubber in figure 6.1. The entire snubber consists of an inductor (L_s), two capacitors (C_b and C_s) and three diodes (D_2 , D_3 and D_4). Figure 6.4 shows the snubber implemented into a basic SEPIC [41].

Figure 6.4: Basic SEPIC with Lossless Passive Snubber Cell [41]

The snubber turn-on circuit in figure 6.2 employs an inductor connected in series with D_1 , which limits D_1 's reverse recovery current increase when S_1 turns on. L_1 already helps limit S_1 's drain current rise during S_1 's turn-on stage. Diodes D_3 and D_4 and buffer capacitor C_b absorb the reverse recovery energy absorbed in both L_s and C_s . C_b then transfers its absorbed energy to the SEPIC's output, completing the energy recovery process. This component combination helps S_1 achieve zero-current switching (ZCS) turn-on.

The snubber turn-off circuit in figure 6.3 employs a capacitor (C_s) connected in parallel with D_1 , which limits S_1 's drain-source voltage (V_{DS}) rise during its turn-off stage. An additional diode, D_2 , provides isolation between L_2 and C_s , preventing S_1 , C_1 and L_2 from accidentally charging C_s and resulting in switch turn-on losses from the charge when there is a low-current load or no load at the SEPIC's output. This component combination helps S_1 achieve zero-voltage switching (ZVS) turn-off (by restricting S_1 's V_{DS} rise during turn-off) and also helps D_1 achieve both ZVS turn-on and turn-off.

Compared to DC-DC converters using resonant soft-switching topologies, this snubber only handles small switching transient energy instead of all the energy from the converter's main power path. Thus, the energy circulated in the snubber is much smaller than the energy circulated through the soft-switching circuits in a conventional resonant ZCS/ZVS DC-DC converter [41]. Thus, this project's SEPIC as a whole is expected to have greater efficiency than the previously selected off-the-shelf resonant ZCS/ZVS DC-DC converter used by the EHFEM group in [1] in this project while operating with a 5-65V input range, when implemented with a passive lossless snubber. The next section describes the snubber's operation during S_1 turn-off, turn-on and intermediate stages.

6.2 Snubber Operation Analysis

The snubber's principal operation consists of eight separate stages. Figure 6.5 shows key waveforms in a boost converter for each snubber operation stage. Because a SEPIC operates in a similar manner to that of a boost converter [20, 21], these waveform shapes are similar for a SEPIC. The only differences lie in waveform magnitudes – in a SEPIC, a secondary inductor (L_2) also supplies energy to the snubber and the load (with the energy provided to it from coupling capacitor C_1), thus while the current waveforms through snubber components are the same shape as that of a boost converter, their magnitudes are higher. Therefore these general waveform shapes are also valid for a SEPIC and are thus valid for analyzing each snubber operation stage for this project's SEPIC [41].

Figure 6.5: Key Waveforms of Boost Converter with Lossless Passive Snubber [41]

In figure 6.5, V_g denotes S_1 's gate voltage (and its switching trajectory), V_{S1} denotes S_1 's drain-source voltage and I_{S1} denotes S_1 's drain current. All other waveforms correspond to voltages across and currents through each respective snubber component. The snubber's first stage (stage 1) operation consists of the time period from t_0 to t_1 , when S_1 turns on. Stage 2 consists of t_1 to t_2 , stage 3 consists of t_2 to t_3 , stage 4 consists of t_3 to t_4 , stage 5 consists of t_4 to t_5 (when S_1 turns off), stage 6 occurs during t_5 to t_6 and stage 7

occurs during t_6 to t_7 . The snubber's final operation stage, stage 8, occurs during t_7 to when S_1 turns on again at t_0 . Time t_0 designates the beginning of a switching period.

For the purpose of this operation analysis, the SEPIC's output filter capacitance C_2 is large enough such that the SEPIC's output voltage V_O is constant and ripple-free. C_2 supplies energy to the load for maintaining load current when S_1 conducts and D_1 does not conduct. As long as C_2 's capacitance is large enough, C_2 does not affect this snubber cell's operation and this analysis treats it as an open circuit (hence neglecting it). The next section describes the snubber's first operation stage.

6.2.1 Snubber Stage 1 Operation

Figure 6.6 shows the SEPIC implemented with the snubber, in the snubber's first operation stage [41].

Figure 6.6: Stage 1 Snubber Operation [41]

The red arrows indicate the current flow direction through each branch. The snubber's first operation stage begins with S_1 turning on at time t_0 . While S_1 turns on, D_1 still conducts because it is not ideal; reverse recovery current still flows through it. Snubber inductor L_s restricts the rate of rise of current through S_1 , achieving zero-current switching (ZCS) MOSFET turn-on. During S_1 turn on, I_{L1} , the current through L_1 ,

increases in the positive direction (from the main voltage source V_s to S_1) while I_{L2} , the current through L_2 , increases in the negative direction. V_s is responsible for the I_{L1} increase. As S_1 turns on, its drain to source voltage decreases to 0V, allowing the voltage across the converter's coupling capacitor, V_{C1} , to equal V_s . Once S_1 's drain-source voltage reaches 0V, the voltage across L_2 , V_{L2} equals $-V_s$. C_1 supplies the energy to increase I_{L2} in the negative direction, increasing the stored energy in L_2 . Equations (6.1) to (6.4) describe the current through L_s (I_{Ls}), C_1 (I_{C1}) and L_2 (I_{L2}) during this operation stage. Equations (6.1) and (6.2) first show the fundamental time-domain capacitor current and inductor voltage equations necessary for understanding the I_{Ls} equations in (6.3) to (6.4) [33, 52].

$$I_C = C \frac{dV_C}{dt} \quad (6.1)$$

$$V_L = L \frac{dI_L}{dt} \quad (6.2)$$

$$I_{Ls}(t) = -I_{C1}(t_0) + I_{L2}(t_0) - \frac{V_o}{L_s}(t - t_0) \quad (6.3)$$

$$I_{Ls}(t) = -C_1 \left(\frac{[V_{Cs}(t) - V_{Cs}(t_0)]}{t - t_0} \right) + I_{L2}(t_0) - \frac{V_o}{L_s}(t - t_0) \quad (6.4)$$

Once the SEPIC reaches time t_1 , the snubber's second operation stage begins.

6.2.2 Snubber Stage 2 Operation

At time t_1 , D_1 's reverse recovery phenomenon finishes. Once D_1 turns off, D_3 turns on since V_{Cs} and V_{Cb} both equal 0V. All energy storage components in the snubber (L_s , C_b , C_s) charge through the first resonant path V_o - C_s - D_3 - C_b - L_s - C_1 - S_1 . The reverse voltage across D_1 increases to $V_{Cs} + V_{Cb}$, and charging C_s and C_b limits the reverse voltage's rise rate. Figure 6.7 shows the SEPIC implemented with the snubber, in the snubber's second operation stage [41].

Figure 6.7: Stage 2 Snubber Operation [41]

While S_1 conducts and D_1 does not conduct, C_2 supplies energy to the load for maintaining the load current. Because C_2 's capacitance is large enough for ensuring a constant, ripple-free output voltage, C_2 does not affect the snubber's operation.

Equations (6.5) to (6.8) give the snubber inductor current (I_{Ls}), snubber capacitor voltage (V_{Cs}), and buffer capacitor voltage (V_{Cb}) during this operation stage.

$$I_{Ls}(t) = -\frac{V_o}{Z_1} \sin(\omega_1(t - t_1)) - I_{rr} \cos(\omega_1(t - t_1)) \quad (6.5)$$

$$V(t) = I_{rr} Z_1 \sin(\omega_1(t - t_1)) - V_o \cos(\omega_1(t - t_1)) + V_o \quad (6.6)$$

$$V_{Cs}(t) = \frac{C_b}{C_s + C_b} V(t) \quad (6.7)$$

$$V_{Cb}(t) = \frac{C_s}{C_s + C_b} V(t) \quad (6.8)$$

Where:

$$I_{rr} = \frac{V_o}{L_s} (t_1 - t_0) + I_{L2}(t_0) + I_{C1}(t_0) - I_{L1}(t_0) \quad (6.9)$$

$$Z_1 = \sqrt{\frac{L_s(C_s + C_b)}{C_s C_b}} \quad (6.10)$$

$$\omega_1 = \sqrt{\frac{C_s + C_b}{L_s C_s + C_b}} \quad (6.11)$$

The snubber slightly increases S_1 's peak drain current requirement (which was 37.95A from (5.24) above). By adding the snubber cell, S_1 's peak drain current requirement becomes its original determined peak current summed with $I_{Ls,p}$, the snubber inductor's peak current. $I_{Ls,p}$ occurs when $V_{Cs} + V_{Cb} = V_o$.

$$I_{Ls,p} = \frac{\sqrt{V_O^2 + (I_{rr}Z_1)^2}}{Z_1} \quad (6.12)$$

The first resonance stage stops at time t_2 when $V_{Cs}(t_2) = V_O$, when diode D_2 turns on. The snubber inductor current at time t_2 is:

$$I_{Ls}(t_2) = \frac{\sqrt{(I_{rr}Z_1)^2 + V_O^2 - V_O \left(\frac{C_s}{C_b}\right)^2}}{Z_1} \quad (6.13)$$

From (6.13), the energy stored in L_s and C_s is:

$$E_{Ls}(t_2) + E_{Cb}(t_2) = \frac{1}{2}L_s I_{Ls}^2(t_2) + \frac{1}{2}C_b V_O^2(t_2) = \frac{1}{2}L_s I_{rr}^2 + \frac{1}{2}C_s V_O^2 \quad (6.14)$$

Once the SEPIC reaches time t_2 , the snubber's third operation stage begins.

6.2.3 Snubber Stage 3 Operation

Once V_{Cs} charges to V_O at time t_2 , D_2 turns on and C_s becomes open, beginning a second resonance stage. I_{Ls} charges C_b through the L_s - L_2 - D_2 - D_3 - C_b resonant path. L_s and C_b perform one-way resonance because of D_2 and D_3 's orientations; current cannot flow through C_b to D_2 and D_3 . Figure 6.8 shows the SEPIC implemented with the snubber, in the snubber's third operation stage [41].

Figure 6.8: Stage 3 Snubber Operation [41]

I_{Ls} and V_{Cb} are:

$$I_{Ls}(t) = \frac{C_s V_O}{C_b Z_2} \sin(\omega_2(t - t_2)) - I_{S2} \cos(\omega_2(t - t_2)) \quad (6.15)$$

$$V_{Cb}(t) = I_{S2} Z_2 \sin(\omega_2(t - t_2)) + \frac{C_s}{C_b} V_O \cos(\omega_2(t - t_2)) \quad (6.16)$$

Where:

$$I_{S2} = \frac{V_O}{Z_1} \sin(\omega_1(t_2 - t_1)) + I_{rr} \cos(\omega_1(t_2 - t_1)) \quad (6.17)$$

$$Z_2 = \sqrt{\frac{L_s}{C_b}} \quad (6.18)$$

$$\omega_2 = \sqrt{\frac{1}{L_s C_b}} \quad (6.19)$$

The second resonance stage stops at time t_3 , when $I_{Ls}(t_3)$ reduces to 0A. At this stage the energy stored in L_s is transfers to C_b . Thus, the energy stored in C_b is:

$$\frac{1}{2} C_b (V_{Cb}(t_3))^2 = E_{Cb}(t_3) = E_{Ls}(t_2) + E_{Cb}(t_2) = \frac{1}{2} L_s I_{rr}^2 + \frac{1}{2} C_s V_O^2 \quad (6.20)$$

Where $E_{Cb}(t_3)$ is the stored energy in C_b at time t_3 , $E_{Cb}(t_2)$ is the stored energy in C_b at time t_2 , and $E_{Ls}(t_2)$ is the stored energy in L_s at time t_2 . Buffer capacitor voltage at t_3 $V_{Cb}(t_3)$ is:

$$V_{Cb}(t_3) = \sqrt{\frac{L_s I_{rr}^2 + C_s V_O^2}{C_b}} \quad (6.21)$$

The buffer capacitor voltage reaches its peak value at time t_3 . Thus, peak buffer capacitor voltage, $V_{Cb,p}$, is:

$$V_{Cb,p} = V_{Cb}(t_3) \quad (6.22)$$

D_1 freewheels during stages 1-3. The maximum reverse voltage imposed across it is:

$$V_{R(D1,max.)} = V_{Cb}(t_3) + V_O \quad (6.23)$$

At time t_3 , the snubber's fourth operation stage begins.

6.2.4 Snubber Stage 4 Operation

Once I_{L_s} reduces to 0A at time t_3 , D_2 and D_3 simultaneously turn off. Because no current flows through C_b , it maintains a constant voltage. The total energy transferred to C_b is the sum of the energy absorbed in L_s and C_s , from (6.20). Figure 6.9 shows the SEPIC implemented with the snubber, in the snubber's third operation stage [41].

Figure 6.9: Stage 4 Snubber Operation [41]

S_1 remains turned on and D_2 remains turned off until time t_4 , when the snubber's fifth operation stage begins.

6.2.5 Snubber Stage 5 Operation

At time t_4 , the SEPIC's switch S_1 turns off. During this stage, the SEPIC's primary inductor current $I_{L1}(t_4)$ flows through D_2 to discharge C_s to the output. L_2 also discharges through the same path. The voltage polarity across C_s causes D_3 and D_4 to remain reverse-biased during this stage. Figure 6.10 shows the SEPIC implemented with the snubber, in the snubber's fifth operation stage.

Figure 6.10: Stage 5 Snubber Operation [41]

During this stage the drain-source voltage across S_1 is:

$$V_{DS(S1)} = V_O - V_{Cs} - V_{C1} \quad (6.24)$$

V_{Cs} discharges from V_O to 0V, and this in turn helps slow down S_1 's drain-source voltage rise. During stage 4, V_{Cs} is:

$$V_{Cs}(t) = V_O - \frac{I_{L1}(t_4) + I_{L2}(t_4)}{C_s} (t - t_4) \quad (6.25)$$

Equation (6.25) assumes that I_{L1} and I_{L2} are both constant during this stage. D_3 and D_4 remain turned off until time t_5 , when the snubber's sixth operation stage begins.

6.2.6 Snubber Stage 6 Operation

At time t_5 , V_{Cs} finishes discharging to 0V and in turn causes D_3 and D_4 to turn on. L_s voltage V_{Ls} equals V_{Cb} and thus I_{L1} and I_{L2} both increase, discharging C_b to the output. Figure 6.11 shows the SEPIC implemented with the snubber, in the snubber's sixth operation stage [41].

Figure 6.11: Stage 6 Snubber Operation [41]

This phenomenon is similar to the resonance cycle in stage 2. I_{Ls} and V_{Cb} are:

$$I_{Ls}(t) = \frac{V_{Cb}(t_2)}{Z_2} \sin(\omega_2(t - t_5)) \quad (6.26)$$

$$V_{Cb}(t) = V_{Cb}(t_2) \cos(\omega_2(t - t_5)) \quad (6.27)$$

D_2 , D_3 and D_4 conduct until time t_6 , when the snubber's second to last operation stage begins.

6.2.7 Snubber Stage 7 Operation

At time t_6 , I_{L1} and I_{L2} both increase to their respective $I_{L1}(t_6)$ and $I_{L2}(t_6)$ values. D_2 and D_3 turn off because of C_b 's voltage polarity, and L_1 and L_2 now discharge to the output through the L_s - C_b - D_4 path. During this stage C_b also discharges to the output. Figure 6.12 shows the SEPIC implemented with the snubber, in the snubber's seventh operation stage [41].

Figure 6.12: Stage 7 Snubber Operation [41]

V_{Cb} during this stage is:

$$V_{Cb}(t) = V_{Cb}(t_6) - \left[\frac{I_{L1}(t_6) + I_{L2}(t_6)}{C_b} \right] (t - t_6) \quad (6.28)$$

D_1 turns on during the next stage, and C_b 's discharging helps slow D_1 's turn-on process, reducing its power dissipation at turn-on. D_4 conducts until the final snubber operation stage.

6.2.8 Snubber Stage 8 Operation

Figure 6.13 shows the SEPIC's snubber in its final operation stage [41].

Figure 6.13: Stage 8 Snubber Operation [41]

During the snubber's final operation stage, C_b discharges to 0V at time t_7 . Once C_b completely discharges to the output, the snubber's energy recovery process finishes. C_b 's discharge turns off D_4 and turns on D_1 . Turning off D_4 helps prevent reversely charging C_s . L_1 and L_2 continue discharging through the L_s - D_1 path (as in a normal SEPIC without this snubber cell) until S_1 turns on again at the end of this stage. Once S_1 turns on again, the snubber's operation cycle restarts at stage 1 and continues through all eight stages as long as the SEPIC is operating with at least 5V input voltage. After completing its operation analysis, the next step in designing the snubber is selecting the proper component values for it.

6.3 Snubber Component Selection

During stage 6, D_2 and D_3 should naturally turn off before C_b discharges to 0V, otherwise the residual current from those two diodes turns on D_2 , D_3 and D_4 during the entire S_1 switching period. Turning on D_2 , D_3 and D_4 during the entire S_1 switching period is undesirable and results in the snubber operating incorrectly. Thus, the snubber component values must obey the following inequality:

$$\frac{1}{2}L_s I_F^2 < \frac{1}{2}L_s I_{rr}^2 + \frac{1}{2}C_s V_O^2 \quad (6.29)$$

From (6.29), I_F is the main switching diode's average forward current, which is the same as this project's SEPIC's maximum load current (8A). Either I_{rr} or C_s needs to be large in order to satisfy this inequality. However, C_s needs to be small for minimizing main switching MOSFET current stress as well as main switching diode reverse voltage stress. 10nF is a reasonable value and selected for C_s .

Equation (6.9) determines I_{rr} . Obtaining I_{rr} requires determining t_0 and t_1 from the SEPIC's simulation results. Time t_0 occurs when S_1 's (Q_1 in the SEPIC's schematic in figure 5.7) gate begins conducting, and time t_1 occurs when D_1 turns off (when its reverse voltage magnitude becomes large). Figure 6.14 shows the converter's diode reverse voltage trajectory in gray, I_{L2} in teal, I_{C1} in magenta, I_{L1} in red and Q_1 's gate switching trajectory in blue.

Figure 6.14: SEPIC Operation During $t_0 < t < t_1$ Period

Time t_0 occurs when Q_1 's gate begins switching on (VN007 in figure 6.14). Time t_1 occurs when D_1 turns off and its reverse voltage begins rising (ABS(V(OUT)-V(N004) in figure 6.14). From there, the obtained I_{L2} , I_{C1} and I_{L1} values substitute into (6.30):

$$I_{rr} = \frac{36V}{L_s} (36.04\mu s - 36.001\mu s) + 6.608A + (-2.59A) - 2.615A = \frac{36V}{L_s} (0.039\mu s) + 1.403A \quad (6.30)$$

Plugging in 10nF for C_s into (6.29) and solving for L_s yields the following:

$$\frac{1}{2} L_s (8A)^2 < \frac{1}{2} L_s \left[\frac{36V}{L_s} (0.039\mu s) + 1.403A \right]^2 + \frac{1}{2} (10nF) (36V)^2 \rightarrow 0 < L_s < 360.6 nH \quad (6.31)$$

Ideally L_s should be as close to the maximum value as possible in order for minimizing diode reverse-recovery losses. Thus, this project selects 360nH for L_s , which is the next lower rated commercially available inductor. This result determines the C_b capacitance. Keeping V_{Cb} at 25% of V_O or below ($V_{Cb} \leq 7.2V$) when the converter's output is 36V is preferable. The selected maximum V_{Cb} value (7.2V) as well as the obtained L_s value plug into (6.21) for obtaining C_b .

$$7.2V = \sqrt{\frac{(360nH) \left[\frac{36V}{360nH} (0.039\mu s) + 1.403A \right]^2 + (10nF)(36V)^2}{C_b}} \rightarrow C_b = 445 nF \quad (6.32)$$

The next higher commercially available capacitor value is 470nF. However, when accounting for reverse recovery energy as well as capacitance tolerances, C_b should be higher than 470nF. Thus, this project selects C_b as 560nF, which is the next higher commercially available capacitor value after 470nF.

Finally, proper snubber cell operation requires that its resonant frequency determined from (6.19) be much greater than the SEPIC's switching frequency (100 kHz). Equations (6.33) and (6.34) determine the snubber's resonant frequency.

$$\omega_2 = \sqrt{\frac{1}{(360nH)(560nF)}} = 2.227 \times 10^6 rad/s \quad (6.33)$$

$$f_2 = \frac{2.227 \times 10^6}{2\pi} = 354.5 kHz \quad (6.34)$$

The snubber's 354.5 kHz resonant frequency is much greater than the SEPIC converter's 100 kHz switching frequency as well as the 111 kHz switching frequency measured in simulation in section 5.4, thus it should work properly with this project's converter. After selecting proper snubber components, this project implements this passive lossless snubber into the simulation SEPIC circuit.

6.4 SEPIC Simulation Results with Snubber

Figure 6.15 shows the SEPIC with the lossless passive snubber:

Figure 6.15: SEPIC Simulation Circuit with Passive Lossless Snubber

This implementation uses MBR20100CTs as snubber diodes because of the MBR20100CT's fast reverse recovery time. Also for the SEPIC's main switching diode (D_1), this implementation uses four MBR20100CTs (two in series and two in parallel – denoted as D_1 , D_5 , D_9 and D_{10} in figure 6.15) for preventing D_1 from exceeding its maximum average forward current and reverse voltage ratings. As mentioned in section 5.4, using only one MBR20100CT for the SEPIC's main switching diode resulted in the diode exceeding its electrical limits. This implementation also adds feedback and sense node capacitors (C_{FB} and C_{SNS}) for ensuring clean signals for the LTC1871 controller's crucial feedback and sense pins. Simulation results for this SEPIC implementation use a 65V input voltage and 8A output load (full load), and this project analyzes them in steady-state operation. Figure 6.16 shows the SEPIC's output voltage in green and power dissipation transients across Q_1 in blue.

Figure 6.16: SEPIC Output Voltage and Q_1 Power Dissipation with Snubber at Full Load and Steady-state Operation

Average SEPIC output voltage is 36.572V, which suffices for the converter's 36V output requirement. The snubber reduces Q_1 's peak power dissipation transients to 180W. These transients were nearly 1.6kW (from figure 5.13) before implementing the snubber. A 1 μ s 180W transient for is much safer for the IPP110N20N3 power MOSFET (Q_1) than a 1.6kW transient lasting the same time span. Overall average Q_1 power dissipation also reduces to 1.8W (from 2.862W without the snubber), which is a dramatic improvement from the previous implementation from section 5.4. Figure 6.17 shows power dissipation across the SEPIC's main switching diode (D_1).

Figure 6.17: SEPIC Switching Diode Power Dissipation with Snubber at Full Load and Steady-state Operation

Using multiple MBR20100CTs as well as a passive lossless snubber greatly reduces power dissipation across D_1 . Its power dissipation reduces to 2.08W (from 6.08W in section 5.4, before this project implemented the snubber). Furthermore, it does not reach close to its average forward current limit or exceed its peak reverse voltage limit, unlike the SEPIC implementation from section 5.4. However, one snubber diode requires further attention. That diode is D_4 from figure 6.15 and is one of the snubber's diodes. The LTSpice simulation produced the warning message shown in figure 6.18:

Figure 6.18: Snubber Diode D_4 Reverse Voltage and Forward Current Stress Warning

D_4 's reverse voltage reaches 105V, which exceeds the MBR20100CT's 100V peak reverse voltage rating. Therefore a single MBR20100CT is not sufficient for tolerating the snubber's output diode stress requirements. Thus, this SEPIC's final implementation uses two MBR20100CTs for its snubber output diode. Now, this project requires an efficiency analysis of its SEPIC when implemented with the designed snubber.

6.4.1 Snubbed SEPIC Efficiency Analysis

Likewise with section 5.4.2, this project performs efficiency analysis of the SEPIC with snubber circuit at full-load conditions and steady-state operation. Figure 6.19 shows the snubbed SEPIC's output voltage in green and its input voltage in blue.

Figure 6.19: SEPIC with Snubber Input and Output Power at Full Load and Steady-state Operation

Average converter output power is 292.58W and average input power is 316.69W. This translates to a 92.4% overall converter efficiency. This converter efficiency is a slight improvement from the 89.2% efficiency (described in section 5.4.2) without a snubber. This efficiency gain is still crucial for reducing the EHFEM system's lifecycle costs and this snubber can be fabricated from very low-cost components because of its simplicity. Therefore this project implements the passive lossless snubber circuit designed in this chapter into its SEPIC.

Overall, while adding a snubber increases the SEPIC's cost and overall PCB real estate, the benefits that it offers outweigh its disadvantages.

Implementing a snubber in this project's SEPIC reduces the magnitude of lethal power dissipation transients imposed across its switching MOSFET and diodes, which are the SEPIC's crucial switching components. However, this snubber circuit only protects components connected to the SEPIC's switching nodes. The SEPIC's input node

requires protection as well from large input ripple voltages that exceed the maximum 65V input specification described in section 1.2. As mentioned earlier in section 3.2, the Precor EFX 546i does not output a flat DC voltage; however, this project earlier did not account for this problem in great depth. Therefore during exercise this project's SEPIC receives a rippled DC voltage at its input. The next chapter and section describes those large ripple input voltages in greater detail.

CHAPTER 7: PRECOR EFX 546i OUTPUT VOLTAGE RIPPLE AND SEPIC INPUT PROTECTION

7.1 Precor EFX 546i Output Voltage Ripple

In Spring Quarter 2011, EHFEM team members Greg Hollister, Alvin Hilario and Nicholas Lovgren described the Precor EFX 546i elliptical trainer outputting a large ripple voltage in addition to its DC voltage [10, 53-54]. That ripple voltage was a sawtooth shape and reached a maximum 32V peak-to-peak amplitude when the elliptical user exercised at 180 strides per minute and level 17 training resistance level setting [10]. The ripple voltage also reached a 40.7 kHz peak frequency at 100 strides per minute and level 10 training resistance (its minimum frequency was 31.3 kHz at 150 strides per minute and level 15 training resistance) [10]. The overall ripple voltage frequency decreases with the elliptical training resistance level, and its amplitude increases with training resistance level. Therefore, if the elliptical trainer outputs a 65V average voltage into this project's SEPIC, its peak will be above the maximum 65V input specification. Inputting voltages above 65V into the SEPIC may damage its components. Additionally, input currents above 25A caused by the input ripple voltage can also damage SEPIC components. Therefore this project's SEPIC requires additional protection at its input node for either clamping input voltages greater than 65V and currents greater than 25A, or completely shutting off operation. An overvoltage and overcurrent protection circuit implemented into the SEPIC's input suffices for those requirements. Section 7.2 describes this project's SEPIC's input overvoltage and overcurrent protection circuit in greater detail.

7.2 Input Overvoltage and Overcurrent Protection Circuit

Dedicated controller ICs for overvoltage and overcurrent protection exist on the consumer market and are readily available from major electronic component suppliers. Therefore implementing an overvoltage and overcurrent protection circuit in this project's SEPIC is much easier with a dedicated controller IC than with discrete components. This project selects Linear Technology's LT4356-1 surge stopper IC as the core of its SEPIC input overvoltage and overcurrent protection circuit. It tolerates up to 100V input voltage and clamps short overvoltage and overcurrent transients to user-defined values set by external biasing components [55]. If those overvoltage and overcurrent transients persist for an extended time period, then the LT4356-1 shuts down voltage and current flow to the rest of the SEPIC circuit via a pass transistor (a power MOSFET). Likewise with the selected LTC1871 PWM controller, the LT4356-1 has an MSOP-10 package available, thereby allowing for a small PCB footprint. Furthermore, like the LTC1871, the LT4356-1 also has an LTSpice model available for simulation. Figure 7.1 shows a typical LT4356-1 application [55].

4A, 12V Overvoltage Output Regulator

Figure 7.1: Typical LT4356-1 Application [55]

This application from Linear's LT4356-1 datasheet protects a DC-DC converter from input overvoltage and overcurrent transients. A resistor divider network connected to the LT4356-1's FB pin determines the maximum input clamping voltage, and a current sense resistor (10m Ω) determines the maximum converter input current. The IRLR2908 pass transistor shuts down voltage and current flow to the DC-DC converter if it detects an overvoltage or overcurrent condition for a prolonged time period. The 0.1 μ F capacitor connected to the LT4356-1's TMR pin determines that time period. This overall LT4356-1 configuration is similar to the one that this project will implement for its SEPIC. Figure 7.2 shows the LT4356-1's internal schematic [55].

Figure 7.2: LT4356-1 Internal Schematic [55]

The A_{OUT} and IN^+ pins control an internal auxiliary amplifier. The LT4356-1 surge stopper IC is available in DFN-12, MSOP-10 and SOIC-16 packages. The LT4356-1 used in this project's SEPIC is in the MSOP-10 package. The LT4356-1's MSOP-10 package does not contain the internal auxiliary amplifier, hence the A_{OUT} and IN^+ pins are also absent from the IC. These two pins are present in LTSpice's LT4356-1 simulation model. However, this project does not use these pins and hence grounds these pins in simulation models. Likewise with the LTC1871 PWM controller, the MSOP-10 package LT4356-1 surge stopper IC consists of 10 pins that require external biasing components.

In numerical order, those pins are FB, OUT, GATE, SNS, V_{CC} , !SHDN, !FLT, EN, GND and TMR. The first pins requiring component selection are the IC's feedback (FB), output (OUT), V_{CC} and current sense (SNS) pins. The LT4356-1's FB pin controls the protection circuit's output voltage limit (its OUT pin), which is the SEPIC's input voltage limit (65V). Its V_{CC} pin connects to an external voltage source (in this project's case, the Precor elliptical trainer's output) for supplying the input voltage that it requires. The LT4356-1's SNS pin connects to a current sense resistor for controlling the protection circuit's input current limit for the SEPIC.

7.2.1 FB, OUT, V_{CC} and SNS Pin Operation

As mentioned in section 7.2, the LT4356-1's FB pin determines the maximum SEPIC input clamping voltage using a resistor divider network. The SEPIC's maximum specified input voltage is 65V from the Precor elliptical trainer, therefore this project designs its LT4356-1 protection circuit such that any SEPIC input voltage transients greater than 65V clamp to this value. This is the same voltage at the LT4356-1's OUT pin, which for this project connects directly to the SEPIC's clamped input voltage node. Equation (7.1) determines the required voltage divider resistors for setting the clamping voltage.

$$V_{REG} = \frac{1.25V(R_{RUN1_OVP} + R_{RUN2_OVP})}{R_{RUN2_OVP}} = 65V \quad (7.1)$$

V_{REG} denotes the converter's maximum desired input voltage (determined as 65V from section 1.2) and R_{RUN1_OVP} and R_{RUN2_OVP} denote the resistors used for the resistor divider network. R_{RUN1_OVP} is the resistor connected to the SEPIC's input and R_{RUN2_OVP} is the resistor connected to ground. They are the 102k Ω and 4.99k Ω resistors in figure 7.1.

Linear Technology recommends limiting the current through R_{RUN1_OVP} and R_{RUN2_OVP} to $250\mu A$ during an overvoltage or overcurrent fault condition [55]. Equation (7.2) uses this R_{RUN2_OVP} current value for determining R_{RUN2_OVP} .

$$R_{RUN2_OVP} = \frac{1.25V}{250\mu A} = 5k\Omega \quad (7.2)$$

$4.99k\Omega$ is the nearest commercially available standard resistor value, thus $R_{RUN2_OVP} = 4.99k\Omega$. Therefore, R_{RUN1_OVP} is the following value:

$$R_{RUN1_OVP} = \frac{(V_{REG}-1.25V)R_{RUN2_OVP}}{1.25V} = \frac{(65V-1.25V)5k\Omega}{1.25V} = 254.49k\Omega \quad (7.3)$$

$255k\Omega$ is the nearest commercially available standard resistor value, thus $R_{RUN1_OVP} = 255k\Omega$. After selecting the proper resistors for input voltage clamping, the LT4356-1 requires selecting components for SEPIC overcurrent protection. As mentioned in section 7.2, the LT4356-1 determines the SEPIC's maximum input current via a current sense resistor connected to between its V_{CC} and SNS pins. The V_{CC} pin directly connects to the Precor elliptical trainer's output for turning on the LT4356-1 IC and can tolerate up to 100V (as well as up to 60V reverse voltage input) [55]. That tolerance suffices for this project's SEPIC's 5-65V input range. The SNS pin senses the voltage across the current sense resistor and if that voltage exceeds 50mV for a prolonged time period, the LT4356-1 shuts off voltage and current flow to the SEPIC [55]. Figure 7.5 denotes this current sense resistor as R_{SNS_OVP} . Equation (7.4) determines the current sense resistor value for limiting the SEPIC's input current to this project's desired value. This project desires limiting the SEPIC's input current to 25A.

$$R_{SNS_OVP} = \frac{50mV}{I_{IN(max)}} = \frac{50mV}{25A} = 2m\Omega \quad (7.4)$$

Earlier in the design process, this project limited SEPIC main switching transistor Q_1 's drain current to 30A through a current sense resistor. However, limiting the SEPIC's input current to 25A prevents L_1 from saturating (because L_1 connects directly to the SEPIC's input node) without significantly compromising the SEPIC's 36V output regulation ability. According to its datasheet, the Vishay IHV28BZ60 inductor saturates when 28A flows through its terminals [36]. In addition to the LT4356-1's output resistor divider and current sense resistor, another crucial component that controls SEPIC shutdown and turn-on during and after a fault condition is the input pass transistor. The next section describes in greater detail biasing the pass transistor.

7.2.2 GATE Pin

The LT4356-1's GATE pin controls the input pass transistor (the IRLR2908 N-channel MOSFET in figure 7.1) that shuts down voltage and current flow to the SEPIC during an overvoltage or overcurrent condition. The GATE pin doesn't require any additional compensation components, but transient input voltage steps greater than 5V/ μ s require connecting a capacitor from the GATE pin to ground. This capacitor helps prevent the pass transistor from self-enhancement [55]. Figure 7.3 shows Linear's LT4356-1 sample test fixture circuit and it includes this additional gate capacitor (C_3).

Figure 7.3: Linear Technology's LT4356-1 Sample Test Fixture Circuit

Figure 7.5 denotes this capacitor as CGATE_OVP. Connecting a series resistor with the pass transistor's gate also helps further dampen any oscillations that may occur, ensuring stable overvoltage and overcurrent regulation. Linear's sample test fixture circuit selects 10Ω for this resistor and $0.0068\mu\text{F}$ (6800pF) for its gate capacitor and these values suffice for this project's protection circuit because they are small values that would not cause any signal underdamping. The next section describes the LT4356-1's input undervoltage shutdown control.

7.2.3 !SHDN Pin and Low Battery Threshold Detection

The LT4356-1 surge stopper should operate over the full specified 5V-65V range for this project, and operating at input voltages lower than 5V does not harm any SEPIC components. Thus this project's SEPIC does not require low battery threshold detection and this project connects the LT4356-1's !SHDN pin directly to the protection circuit's input node. The LT4356-1's !SHDN pin has the same voltage tolerances as the V_{CC} pin, thus tying it to the protection circuit's input suffices. The LT4356-1's other shutdown condition is during a prolonged overvoltage or overcurrent fault condition. Physically

indicating (such as through an LED) when such a fault condition occurs helps simplify troubleshooting the protection circuit if needed. The next section describes implementation of fault condition indicators for this project's SEPIC's protection circuit.

7.2.4 !FLT Pin

The LT4356-1's active-low !FLT pin pulls to a logic low during an overvoltage or overcurrent fault condition. This pin connects to any type of fault indicator, such as an LED, and can drive that indicator by sinking up to 3mA from an external voltage source. This project indicates an input overvoltage or overcurrent fault by connecting an LED to the !FLT pin, using the Precor elliptical trainer's onboard 12V battery as its voltage source. This project uses Lumex's SML-LX1206SRC-TR 1206 package red LED as its fault indicator. Simulations, however, use Nichia's NSCW100 LED as its model because of its availability in LTSpice. Figure 7.5 denotes this LED as FLT_LED. Additionally, this project also requires a series resistor connected to the LED for limiting its current to 3mA, as its typical forward voltage drop is 1.7V [56]. Figure 7.5 denotes this series resistor as R_{FLT} . Equations (7.5) and (7.6) determine the fault LED's maximum current:

$$I_{FLT_LED} = \frac{V_{BAT} - V_{FLT_LED}}{R_{FLT}} \quad (7.5)$$

$$R_{FLT} = \frac{V_{BAT} - V_{FLT_LED}}{I_{FLT_LED}} = \frac{12V - 1.7V}{3mA} = 3.43k\Omega \quad (7.6)$$

V_{BAT} denotes the Precor elliptical trainer's battery voltage, V_{FLT_LED} denotes to the LED's forward voltage, I_{FLT_LED} is its current and R_{FLT} is the required resistance value for the LED's current-limiting resistor. 3.6k Ω is the next higher available standard resistor value, thus this project selects 3.6k Ω for R_{FLT} . Equation (7.7) recalculates the current flowing through the fault LED using this selected resistor value:

$$I_{FLT_LED} = \frac{V_{BAT} - V_{FLT_LED}}{R_{FLT}} = \frac{12V - 1.7V}{3.6k\Omega} = 2.861mA \quad (7.7)$$

This current value suffices for the fault LED, as it is below 3mA. After selecting components for the LT4356-1's !FLT pin for driving a fault indicator, the next pin requiring programming is the enable (EN) pin.

7.2.5 EN Pin

The LT4356-1's EN pin is its open collector enable output [55]. Linear Technology's LT4356-1 datasheet recommends connecting this pin directly to the DC-DC converter's main PWM controller EN, RUN or !SHDN pin. Thus, this project ties this pin to the LTC1871 controller's RUN pin. The next section describes proper grounding for the LT4356-1 surge stopper.

7.2.6 GND Pin

Likewise with the LTC1871 PWM controller, the LT4356-1's GND pin connects directly to the SEPIC's ground node. Proper, low-noise operation however, requires it connected to a signal ground plane separate from that of a power ground plane. Chapter 9 (PCB Layout) discusses this requirement in greater detail. After properly connecting the LT4356-1's GND pin, the final pin requiring biasing components is its fault timer (TMR) pin.

7.2.7 Fault Timer Control and TMR Pin

The LT4356-1's TMR pin sets the protection circuit's fault condition early warning, fault and cool-down time periods [55]. An external biasing capacitor determines these time periods. Figure 7.5 denotes this capacitor as C_{TMR} . The current charging up this timing capacitor connected to this pin during fault conditions depends on the voltage

difference between the LT4356-1's V_{CC} and OUT pins. When the voltage across the TMR pin charges to 1.25V, the !FLT pin pulls low, indicating that the LT4356-1 detects an input overvoltage or overcurrent fault condition. If the condition persists longer, the protection circuit's pass transistor shuts off once the timer capacitor further charges to 1.35V. During this charging period, the LT4356-1 clamps the SEPIC input voltage to 65V or the input current to 25A, depending on what occurs first. The LT4356-1 shuts down voltage and current flow to the SEPIC's input once the timer capacitor charges to 1.35V. Once the fault condition disappears, no current flows through the TMR pin (thus preventing the timer capacitor from charging) and a $2\mu A$ current pulls the TMR pin down, discharging the timer capacitor. Once the timer capacitor discharges to the TMR pin's 0.5V retry threshold, the GATE pin pulls high, turning back on the protection circuit's pass transistor and allowing voltage and current to flow into the SEPIC as long as no further prolonged fault condition persists [55]. Linear Technology recommends $1\mu s$ early warning time for fault detection, thus this project uses $1\mu s$ as its early warning time [55]. This value is 10 times lower than this project's SEPIC's $10\mu s$ switching period, thus this value suffices for proper protection circuit operation. Equation (7.8) determines the C_{TMR} value required for $1\mu s$ early warning time for fault detection:

$$C_{TMR} = \frac{1\mu s \cdot 5\mu A}{100mV} = 50pF \quad (7.8)$$

47pF is the nearest commercially available standard capacitor value, thus this project selects C_{TMR} as 47pF. Equation (7.9) determines the total fault time allowed for SEPIC input voltage and current clamping. Equation (7.9) denotes this fault time as t_{OC} .

$$t_{OC} = \frac{47pF \cdot 0.85V}{45.5\mu A} = 0.878\mu s \quad (7.9)$$

This value suffices for 31.3 kHz input ripple from the Precor elliptical. Therefore, for this project's SEPIC, the LT4356-1 clamps input overvoltage and overcurrent transients with less than 0.878μs duration and shuts off voltage and current flow if those transients persist for a longer period. During the clamping time period, the protection circuit's pass transistor must withstand a large power dissipation because a significant voltage difference exists between the pass transistor's drain and source. This difference is between the input voltage (output voltage from the Precor elliptical trainer) and the clamped voltage, and results in V×I dissipation. Figure 7.5 denotes the protection circuit's pass transistor as Q_OVP.

7.2.8 Pass Transistor Selection

As described in section 7.2.1, 65V input to the LT4356-1 is the minimum set threshold for the IC's input overvoltage condition. Input overcurrent conditions may occur at lower input voltages. However, at those lower input voltages Q_OVP dissipates less power because its drain-source voltage (V_{DS}) is lower. Thus, this project sets 65V as the baseline for determining the required Q_OVP power dissipation during a fault condition. V_{CC(max.)} equals the maximum SEPIC input voltage V_{IN(max.)} because the LT4356-1's V_{CC} pin connects to the SEPIC input. Equation (7.10) determines the maximum power dissipation level that Q_OVP must withstand during voltage or current clamping. Equation (7.10) denotes this power dissipation level as P_{Q_OVP}.

$$P_{Q_OVP} = \frac{V_{CC(max.)} \cdot 50mV}{R_{SNS_OVP}} = \frac{65V \cdot 50mV}{2m\Omega} = 1.625kW \quad (7.10)$$

Q_OVP must withstand this power dissipation level for the duration equal to t_{oc} (determined as $0.878\mu s$ in (7.9) above). This project selects IXYS's IXFK230N20T for Q_OVP, which has a 1.67kW absolute maximum power dissipation rating [57].

Preventing Q_OVP's destruction requires the IXFK230N20T's safe operating area falling within the boundaries set by both t_{oc} and P_{Q_OVP} . However, the operating voltage and current require extraction from P_{Q_OVP} in because the IXFK230N20T's safe operating area plot consists of only operating voltage plotted against operating current [57]. The SEPIC's maximum operating input voltage ($V_{CC\ max}$) is 65V; now only the current (I_{Q_OVP}) requires extraction from P_{Q_OVP} :

$$P_{Q_OVP} = V_{CC(max.)} \cdot I_{Q_OVP} \quad (7.11)$$

$$I_{Q_OVP} = \frac{P_{Q_OVP}}{V_{CC(max.)}} = \frac{1.625kW}{65V} = 25A \quad (7.12)$$

Figure 7.4 shows the IXFK230N20T's safe operating area, with the obtained maximum operating voltage and current values from (7.1) and (7.12) marked in blue [57].

Figure 7.4: IXYS IXFK230N20T Safe Operating Area [57]

From figure 7.4, the IXYS IXFK230N20T withstands 65V operating voltage and 25A operating current for longer than 1ms when operating at its 175°C maximum junction temperature, which is well above the required 0.878μs duration. Thus the IXYS IXFK230N20T qualifies as a suitable transistor for Q_OVP. After selecting all the proper biasing components for the LT4356-1 surge stopper, this protection circuit is ready for implementation into the simulation SEPIC circuit.

7.3 Protected SEPIC Simulation Results

Figure 7.5 shows the complete SEPIC circuit implemented with the snubber from chapter 6 and the input overvoltage and overcurrent protection from section 7.2.

Figure 7.5: Complete SEPIC Schematic

This finalized SEPIC implementation adds ceramic capacitors in parallel with the large electrolytic capacitors at the elliptical input and output nodes, for reducing overall ESR that the electrolytic capacitors add. Furthermore, it also adds additional capacitors ($4.7\mu\text{F}$ each) in parallel with the coupling capacitor for reducing its ripple voltage and RMS current requirement. Ideally the coupling capacitor's ripple voltage should be low, but not so low that it results in unstable converter operation [27]. This implementation also uses two output snubber diodes (denoted as DSNUB3 and DSNUB4 in figure 7.5) for preventing their reverse voltages from exceeding 100V. It also increases input filter capacitor C_{IN} to $100\mu\text{F}$ (from $33\mu\text{F}$ in previous implementations) after taking the elliptical trainer's output voltage ripple into consideration. The LT4356-1 protection circuit's output also contains an additional filter capacitor (denoted as CCLAMP in figure 7.5) for suppressing any ripple voltages that may occur at the protection circuit's output.

This implementation also adds an input transient suppression diode (a 1N5378B 100V zener diode, denoted as TSD in figure 7.5) for protecting the LT4356-1 IC from potential input voltage ripples that exceed 100V. As described in section 7.1, EHFEM team members measured a 32V peak-to-peak ripple from the elliptical trainer's output at higher output voltages, which corresponds to an 81V peak input voltage for a 65V average output, and although the EHFEM team members did not measure the elliptical trainer's output ripple voltage at 65V, voltage ripples greater than 100V may occur. Also, this implementation also renames all component reference designators to those more understandable to the end user. Unlike previous implementations, this project also simulates this SEPIC at 40°C ambient operating temperature, as this temperature will very likely be the same as when a future project encloses this SEPIC inside a confined chassis. Figure 7.6 shows the protected SEPIC's output voltage in green and power dissipation transients across Q_1 in blue, while operating at full-load and steady-state conditions.

Figure 7.6: Protected SEPIC Output Voltage and Q_1 Power Dissipation at Full Load and Steady-state Operation

The protected SEPIC's characteristics in steady-state remain unchanged from the unprotected SEPIC's in section 6.4. Its steady-state efficiency also remains unchanged at 92.4%. The LT4356-1 protection circuit, however, only shows its true advantages at the SEPIC's input during an overvoltage or overcurrent fault condition. Figure 7.7 shows such an example. This next simulation in figure 7.7 changes the SEPIC's main source voltage to a 32V peak-to-peak, 31.3 kHz sawtooth wave with 65V DC offset (81V peak voltage and 49V minimum voltage). This input waveform uses the SEPIC's 65V maximum DC voltage input specified in section 1.2 as well as the lowest recorded ripple frequency recorded from [10] for testing the LT4356-1 protection circuit's voltage clamping abilities using the longest possible overvoltage transient. Figure 7.7 shows the SEPIC's input voltage in green, LT4356-1 output voltage in blue, as well as Q_{OVP} 's power dissipation in red.

Figure 7.7: Protected SEPIC Input Voltage, Clamped Input Voltage and Q_OVP Power Dissipation at Full Load and Steady-state Operation

Using $100\mu\text{F}$ for C_{IN} effectively quells the input voltage ripple, as the LT4356-1 input protection circuit clamps input voltage transients, as shown in the LT4356-1's output voltage waveform in figure 7.7. However, Q_OVP dissipates 86.223W average in steady-state, which requires a very expensive heatsink for cooling. Therefore, this project's SEPIC requires design modifications to its input protection circuit. The only design modification that the input protection circuit requires is changing C_{IN} 's capacitance value. Equation (6.1) from section 6.2 is crucial for understanding how capacitance value affects voltage ripple. Voltage ripple is inversely proportional to capacitance value, and increasing C_{IN} to $1000\mu\text{F}$ should reduce input ripple voltage by a factor of 10. Equations

(7.13) and (7.14) modify (6.1) and calculate the SEPIC's input peak to peak ripple voltage using the previously defined 100μF as well as new 1000μF C_{IN} value as well as the SEPIC's specified 6.5A maximum input current and the 31.3 kHz minimum input voltage ripple frequency determined from section 7.1.

$$\Delta V_C = \frac{I_C \Delta t}{C} = \frac{I_C}{C \times f} = \frac{6.5A}{(100\mu F)(31.3kHz)} = 2.08V \quad (7.13)$$

$$\Delta V_C = \frac{I_C \Delta t}{C} = \frac{I_C}{C \times f} = \frac{6.5A}{(1000\mu F)(31.3kHz)} = 0.208V \quad (7.14)$$

Therefore, increasing C_{IN} to 1000μF decreases the peak to peak input voltage ripple by tenfold and a 0.208V peak to peak input voltage ripple suffices as a stable input voltage. This peak to peak input voltage ripple reduction should help decrease Q_OVP's power dissipation stress. Future SEPIC revisions will increase C_{IN}'s capacitance value again if necessary. This project now re-simulates the waveforms from figure 7.7 using the new 1000μF input capacitor. Figure 7.8 shows these results, with the SEPIC's input voltage in green, LT4356-1 output voltage in blue, as well as Q_OVP's power dissipation in red.

Figure 7.8: Protected SEPIC Input Voltage, Clamped Input Voltage and Q_OVP Power Dissipation at Full Load and Steady-state Operation with 1000μF Input Capacitor

With a 1000μF input capacitor Q_OVP's average input voltage clamping power dissipation drops to nearly 12W in steady-state, which is an acceptable level for this converter's application, however, such power dissipation still requires a heatsink with low thermal resistance. The LT4356-1 input protection circuit only clamps the input to 64V because of non-ideal V_{DS} drops across Q_OVP. With a high time-varying input voltage to the SEPIC, C_{IN} 's RMS current also exceeds 50A according to the same simulation from figure 7.8. One way to remedy this problem is connecting C_{IN} in series with the Precor elliptical trainer's onboard 10Ω resistor load (with the load grounded at one end). This project's SEPIC's PCB layout in Chapter 9 includes an additional

connector for C_{IN} , allowing for a connection either to the Precor elliptical trainer's 10Ω load or ground. Doing so reduces C_{IN} 's RMS current to a more acceptable 1A while still maintaining the same LT4356-1 performance as the simulation from figure 7.8. For the remaining simulations as well as synthetic converter hardware tests described in chapter 10 (i.e. using a constant DC power supply as an input source instead of the Precor elliptical), C_{IN} connects to ground for simplicity. The next section details the SEPIC's finalized design before implementing it in PCB layout.

7.4 Finalized SEPIC Simulation Results

Figure 7.9 shows the final SEPIC schematic before this project lays it out on a PCB in chapter 9.

Figure 7.9: Finalized SEPIC Schematic Before PCB Layout

This finalized implementation adds 0.1 μ F ceramic bypass capacitors to the SEPIC's elliptical input, 12V battery input, switching and output nodes. These bypass capacitors prevent resonance effects from occurring at the SEPIC's 100-111 kHz switching frequency, and also prevent high frequency harmonics from causing interference with circuit operation. An example of such resonance effects is that the large input and output capacitors no longer function as capacitors – instead they behave like inductors (with increasing impedance at higher frequencies rather than the opposite) and lose their high frequency signal passing functions [58]. This implementation also renames the SEPIC's main switching transistor Q_1 to Q_MAIN . Furthermore, this implementation also adds a resistor divider network consisting of R_{VCC1} and R_{VCC2} at the LTC1871 controller's $INTV_{CC}$ pin, preventing large inrush current spikes at start-up that may damage the LTC1871 controller. This project selects R_{VCC1} and R_{VCC2} such that the LTC1871's $INTV_{CC}$ pin still maintains its required 5.2V operation voltage during normal controller operation while simultaneously drawing low current from the Precor elliptical trainer's onboard 12V battery. This implementation also considers using IXYS's IXYS's IXFK230N20T as an alternate transistor for Q_MAIN because of its high power tolerance. Further report sections refer to the IPP110N20N3 as Q_MAIN and the IXFK230N20T as Q_MAIN_ALT (Q_MAIN alternate). Also not shown in figure 7.9 are fuses protecting the converter's main input, battery input and output nodes. UL, IEEE 1547, NEC and PG&E safety guidelines require these fuses and they are included as off-PCB devices during hardware testing. Table B-1 shows this project's SEPIC component

bill of materials (BOM) with all component part numbers for this finalized converter,
sans PCB costs.

Table B-1: Finalized SEPIC Component BOM

Controller I.C.s:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
LTC1871HMS (10-MSOP)	1	4.83	4.83	U1	10-MSOP	N/A	N/A
LT4356-1HMS (10-MSOP)	2	3.70	7.40	U2	10-MSOP	N/A	N/A
Total Cost (\$)			12.23				
Transistors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Infineon IPP110N20N3	1	8.06	8.06	Q_MAIN	TO-220-3	2.45W	4.01W
IXYS IXFK230N20T	1	9.50	9.50	Q_OVP	TO-264	1.625kW	1.625kW
Total Cost (\$)			17.56				
Diodes:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
MBR20100CT (Schottky)	7	0.88	6.16	D_MAIN1, D_MAIN2, D_MAIN3, D_MAIN4, DSNUB2, DSNUB3, DSNUB4	TO-220-3	31.11A, 110V	27.72A, 56.46V
Microsemi APT30S20BG (Schottky)	1	3.67	3.67	DSNUB1	TO-247	39.43A, 15.95V	27.89A, 9.58V
1N5378B (Zener)	1	0.57	0.57	TSD	Through-Hole	100V	100V
SML-LX1206SRC-TR (LED)	1	0.43	0.43	FLT_LED	1206	1.2V min. forward drop	1.2V min. forward drop
Total Cost (\$)			10.83				
Fuses:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Radio Shack 270-270-1234 30A Blade Fuse Holder	2	2.99	5.98	Fuse Holder for Input and Output Fuse	N/A	65V, 25A	5V, 25A
Cooper Bussmann 270-1085 30A Blade Fuse	1	0.73	0.73	Input Fuse	Blade Fuse	65V, 25A	5V, 25A
Cooper Bussmann 270-1081 10A Blade Fuse	1	0.73	0.73	Output Fuse	Blade Fuse	36V, 8A	5V, 0.5A
Bourns SF-0402F200-2 2A Fuse	1	0.68	0.68	Elliptical Battery Fuse	0402	12V, 2A	12V, 2A
Total Cost (\$)			8.12				

Table B-1 (Continued)

Resistors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
274k, 1/8W, 1%, ERJ-6ENF2743V	1	0.07	0.07	RRUN1_MAIN	0805	0.521mW	0.522mW
255k, 1/8W, 1%, ERJ-6ENF2553V	1	0.07	0.07	RRUN1_OVP	0805	16.289mW	0.094mW
220k, 1/8W, 1%, ERJ-6ENF2203V	1	0.07	0.07	RFREQ	0805	0.002mW	0.002mW
133k, 1/8W, 1%, ERJ-6ENF1333V	1	0.07	0.07	RRUN2_MAIN	0805	0.115mW	0.122mW
115k, 1/8W, 1%, ERJ-6ENF1153V	1	0.07	0.07	RFB1	0805	11.12mW	12.28mW
33.2k, 1/8W, 1%, ERJ-6ENF3322V	1	0.07	0.07	RTH	0805	28.32mW	0.028mW
4.99k, 1/8W, 1%, ERJ-6ENF4991V	1	0.07	0.07	RRUN2_OVP	0805	0.319mW	0.002mW
3.92k, 1/8W, 1%, RMCF0805FT3K92	1	0.04	0.04	RFB2	0805	0.387mW	0.427mW
3.6k, 1/8W, 1%, ERJ-6ENF3601V	1	0.07	0.07	RFLT	0805	32.4mW	32.4mW
887, 1/8W, 1%, ERJ-6ENF8870V	1	0.07	0.07	RVCC1	0805	< 125mW	< 125mW
680, 1/8W, 5%, ERJ-6GEYJ681V	1	0.04	0.04	RVCC2	0805	< 125mW	< 125mW
10, 1/8W, 1%, ERJ-6ENF10R0V	1	0.07	0.07	RGATE_OVP	0805	13.05mW	0W
0.005, 5W, 1%, OARSXPR005FLF	1	1.42	1.42	RSNS	SMT (Custom)	410mW	1.132W
0.002, 5W, 1%, WSLP39212L000FEB	1	3.07	3.07	RSNS_OVP	SMT (Custom)	64mW	1.000W
Total Cost (\$)			5.27				
Inductors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Vishay IHV28BZ60 (60μH, I _{SAT} =28A)	2	20.30	40.60	L1, L2	Through-Hole	30.68A (max.), 65V	22.37A (max.), 52.22V
Vishay IHLP4040DZERR36M01 (360nH, I _{SAT} =60A)	1	2.52	2.52	LSNUB	SMT (Custom)	29.3A (max.), 106.25V	17.85A (max.), 55.35V
Total Cost (\$)			43.12				

Table B-1 (Continued)

Capacitors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
470μ, Electrolytic, 400V, B43504A9477M	1	13.34	13.34	COUT	Through-Hole	36V, 5.43A (RMS)	36V, 2.368A (RMS)
1000μ, Electrolytic, 100V, UPW2A102MHD	1	3.03	3.03	CIN	Through-Hole	65V, 1.886pA (RMS)	5V, 227.67fA (RMS)
47μ, Electrolytic, 100V, AFK476M2AH32T-F	1	2.35	2.35	CCLAMP	SMT (Custom)	65.68V, 119.05mA (RMS)	5V, 42.067mA (RMS)
33μ, Electrolytic, 35V, EEE-HA1V330WP	1	0.63	0.63	CBAT	SMT (Custom)	12V, 206.56fA (RMS)	12V, 442.09fA (RMS)
15μ, Ceramic, 250V, KHD251E156M99C0B00	1	25.78	25.78	CINT	Through-Hole	77.05V, 3.67A (RMS)	21.22V, 1.404A (RMS)
10μ, Ceramic, 50V, UMK325C7106MM-T	3	0.96	2.88	COUT2, COUT3, COUT4	1210	36V, 2.758A (RMS)	36V, 2.016A (RMS)
10μ, Ceramic, 25V, TMK316B7106KL-TD	2	0.58	1.16	CBAT2, CBAT3	1206	12V, 963.63fA (RMS)	12V, 1.211pA (RMS)
4.7μ, Ceramic, 100V, C4532X7S2A475M	5	1.75	8.75	CINT2, CINT3, CINT4, CINT5, CINT6	1812	77.05V, 1.22A (RMS)	21.22V, 1.22A (RMS)
4.7μ, Ceramic, 16V, EMK212B7475KG-T	1	0.38	0.38	CVCC	0805	5.23V, 153.1mA (RMS)	5.23V, 170.84mA (RMS)
0.56μ, Ceramic, 25V, C0805C564K3RACTU	1	0.84	0.84	CSNUB1	0805	11.33V, 3.629A (RMS)	11.00V, 1.577A (RMS)
0.1μ, Ceramic, 100V, CGA3E3X7S2A104K	7	0.30	2.10	CIN2, CIN3, CCLAMP2, CCLAMP3, CCLAMP4, CBAT4, CBAT5	0603	65V, 1.000mA (RMS)	5V, 0.285mA (RMS)
0.1μ, Ceramic, 100V, HMK212B7104KG-T	2	0.30	0.60	CINT7, CINT8	0805	77.05V, 47.184mA (RMS)	21.22V, 17.59mA (RMS)
0.1μ, Ceramic, 50V, C1608X7R1H104M	2	0.16	0.32	COUT_CPH1, COUT_CPH2	0603	36V, 63.00mA (RMS)	36V, 46.00mA (RMS)
10000p, Ceramic, 250V, C2012X7R2E103K	1	0.33	0.33	CSNUB2	0805	103.44V, 1.619A (RMS)	56.06V, 756.43mA (RMS)
6800p, Ceramic, 100V, C1608X8R2A682K	1	0.36	0.36	CGATE_OVP	0603	77.47V, 4.796mA (RMS)	11.331V, 0.006mA (RMS)
6800p, Ceramic, 25V, C1005X7R1E682K	1	0.09	0.09	CTH2	0402	1.232V, 0.110mA (RMS)	1.232V, 0.030mA (RMS)
56p, Ceramic, 10V, C0402C560J8GACTU	1	0.35	0.35	CTH1	0402	1.232V, 0.390mA (RMS)	1.232V, 0.009mA (RMS)
47p, Ceramic, 10V, 0603ZA470DAT2A	5	0.30	1.50	CFB, CSNS, CTMR, CRUN_OVP, CRUN_MAIN	0603	1.5V, 0.028mA (RMS)	1.307V, 0.020mA (RMS)
Total Cost (\$)			64.79				
Component Count	68						
Total Converter Cost (\$)	161.92						

The BOM in table B-1 determines components based on required absolute maximum electrical ratings based on simulation results using the schematic from figure 7.9. Using those maximum electrical ratings, this project determined the most appropriate parts and selected them from component suppliers such as Digi-Key and Mouser Electronics. The most expensive components for this SEPIC are its switching inductors (L_1 and L_2) as well as its coupling capacitor (C_{INT}). This project's SEPIC's switching inductors and coupling capacitors require low ESR as well as high current tolerance (RMS current tolerance on the capacitor). The capacitor selected for C_{INT} , a Nippon Chemi-Con KHD251E156M99C0B00, is a ceramic capacitor with 250V and 4A RMS current tolerance [59], which suffices for this project's SEPIC. Because it is a ceramic capacitor, its ESR is also inherently low. Using an electrolytic capacitor with similar voltage and RMS current ratings for C_{INT} reduces this SEPIC's cost, but requires more parallel ceramic capacitors for offsetting the electrolytic capacitor's high ESR. Selecting all components also helps determine each component's parasitic electrical values such as ESR or parallel capacitance. This project then implemented those parasitic values into the SEPIC's components in the schematic in figure 7.9, for determining how much those parasitic values affect overall converter performance. Because ceramic capacitors typically do not indicate ESR values in their datasheets, this project uses 0.02Ω as the ESR value for all of its SEPIC's ceramic capacitors. Typical ceramic capacitor ESR values maximize at 0.015Ω , and this project uses a conservative 0.02Ω value for all of its SEPIC's ceramic capacitors [60]. Figures 7.10 and 7.11 show these results. Figure 7.10 shows the finalized SEPIC's output voltage waveform in green and Q_{MAIN} 's power

dissipation at full load (65V input, 8A load) and in steady-state, with the parasitic component values implemented. Figure 7.11 shows the SEPIC's output power in green and input power in blue. Likewise with the SEPIC implementation in section 7.3, this project also runs these simulations at 40°C ambient operating temperature.

Figure 7.10: Finalized SEPIC Output Voltage and Q_{MAIN} Power Dissipation with Parasitic Component Values

Figure 7.11: Finalized SEPIC Converter Input and Output Power with Parasitic Component Values

From figure 7.10, average SEPIC output voltage is 37.29V, which is slightly higher than the required 36V, but nonetheless suffices. The increased output voltage results from changing one of the SEPIC's output feedback resistors, R_{FB2} , to 3.92k Ω from 4k Ω . This project changed R_{FB2} 's value for cost reduction and component availability purposes, because 0805 3.92k Ω resistors were available at a much lower price than 4k Ω resistors. Peak-to-peak output voltage ripple is 2.959V, which corresponds to a 7.94% output voltage ripple. This value is below the 10% output voltage ripple requirement specified in section 1.2, therefore this value suffices. Average Q_MAIN power dissipation in steady-state is 2.29W, increased from 1.8W from the design in section 6.4. The parasitic component values caused this power dissipation increase across Q_MAIN, however 2.29W suffices with adequate cooling on Q_MAIN. From figure 7.11, average SEPIC output power is 291.31W and average input power is 312.17W. This translates to a 93.3% converter efficiency at full load and steady-state, with all parasitic values implemented

into the SEPIC's components. This is still above the 75% efficiency specification outlined in section 1.2, therefore this simulated efficiency value suffices. Therefore, even after adding parasitic component values to this project's SEPIC, it still performs better in simulation than the specifications outlined in section 1.2. This project next implements takes the SEPIC from figure 7.9 and determines cooling requirements for components that dissipate more than 1W power, as such components experience shortened lifespan and decrease overall converter lifespan if not adequately cooled.

CHAPTER 8: THERMAL CONSIDERATIONS

8.1 Semiconductor Device Power Dissipation

Because the transistors and diodes in this project's SEPIC are far from ideal, they still dissipate power in the form of heat, as witnessed in previous chapters. Operating such components without heat removal devices result in shortened component lifespan as well as shortened overall SEPIC lifespan. Shortened overall SEPIC lifespan results in frequent component replacement and consequently, increased the EHFEM system's lifecycle costs. Heatsinks are the most common and least expensive type of heat removal devices for semiconductor devices. They also save physical PCB space without using any PCB copper pads (in the case of thermal vias). Therefore, this project employs heatsinks for removing heat from its SEPIC's transistors and diodes for ensuring prolonged converter operation. Determining the maximum power dissipation in Q_MAIN, the main switching diodes (D_MAINx) and snubber diodes (D_SNUBx) requires knowing the lowest possible input voltage at which the SEPIC maintains 36V output and drive an 8A load (288W output). From simulation, 27V was the minimum input voltage at which the SEPIC maintained a 36V output with an 8A load. Realistically, the SEPIC will not drive an 8A load at 27V input because it must maintain a 10Ω input resistance, but this conservative estimate ensures that selected heatsinks more than adequately cool components requiring heatsinking and that the SEPIC does not suffer premature failure. Increasing C_{IN} to $1000\mu F$ in section 7.4 dramatically reduced the average power dissipation across Q_OVP during an overvoltage input condition to 10.56W, from 60.02W with C_{IN} at $100\mu F$.

Tables C-1 and C-2 determine from simulation the average steady-state power dissipation across the SEPIC's main semiconductor components. Table C-1 determines average power dissipation when the SEPIC operates with 65V input and 8A load, and table C-2 determines peak power dissipation when the SEPIC operates with 27V input and 8A load.

Table C-1: SEPIC Semiconductor Component Average Steady-state Power Dissipation with 65V Input and 8A Load

SEPIC Component	Average Power Dissipation (W)
D_MAIN1 to 4	3.436
D_SNUB1	1.871
D_SNUB2	3.48
D_SNUB3 and 4	3.498
Q_MAIN	2.2
Q_OVP	0.25

Table C-2: SEPIC Semiconductor Component Average Steady-state Power Dissipation with 27V Input and 8A Load

SEPIC Component	Average Power Dissipation (W)
D_MAIN1 to 4	1.584
D_SNUB1	1.932
D_SNUB2	3.939
D_SNUB3 and 4	4.084
Q_MAIN	6.0
Q_OVP	0.53

This project's heatsink selection process considers the higher power dissipation values of the two operation cases above. Also this project rounds up all power dissipation values in thermal calculations, for further conservative thermal resistance requirement estimates.

The next section determines each component's heatsink's required thermal resistance ratings.

8.2 Heatsink Thermal Resistance Ratings

Equations (8.1) and (8.2) determine the required thermal resistance rating for each SEPIC semiconductor component's heatsink. Both equations also assume 40°C ambient converter operation temperature.

$$P_{MAX} = \frac{T_{JMAX} - T_A}{R_{\theta JC} + R_{\theta CS} + R_{\theta HA}} \quad (8.1)$$

Where:

$$\begin{aligned} P_{MAX} &= \text{Maximum component power dissipation [W]} \\ T_{JMAX} &= \text{Maximum junction operating temperature [°C]} \\ T_A &= \text{Ambient air temperature [°C]} \\ R_{\theta JC} &= \text{Junction to case thermal resistance [°C/W]} \\ R_{\theta CS} &= \text{Case to heatsink thermal resistance [°C/W]} \\ R_{\theta HA} &= \text{Heatsink to ambient (air) thermal resistance [°C/W]} \\ R_{\theta HA} &= \frac{T_{JMAX} - T_A}{P_{MAX}} - R_{\theta JC} - R_{\theta CS} \end{aligned} \quad (8.2)$$

Equation (8.3) calculates Q_OVP's required heatsink thermal resistance, using IXYS's IXFK230N20T datasheet thermal values [57]. Although from tables C-1 and C-2 Q_OVP only dissipates 0.53W average power in the worst case scenario, it must withstand 12W average should the SEPIC encounter an input overvoltage and overcurrent transient and require Q_OVP clamping that transient.

$$R_{\theta HA} = \frac{175^{\circ}\text{C} - 40^{\circ}\text{C}}{12\text{W}} - 0.09^{\circ}\text{C/W} - 0.15^{\circ}\text{C/W} = 11.01^{\circ}\text{C/W} \quad (8.3)$$

Equation (8.4) calculates Q_MAIN_ALT's required heatsink thermal resistance if the project ever necessitated its use. Likewise with (8.3), (8.4) calculates the required heatsink thermal resistance using IXYS's IXFK230N20T datasheet thermal values [57].

$$R_{\theta HA} = \frac{175^{\circ}\text{C} - 40^{\circ}\text{C}}{6\text{W}} - 0.09^{\circ}\text{C/W} - 0.15^{\circ}\text{C/W} = 22.26^{\circ}\text{C/W} \quad (8.4)$$

Equation (8.5) calculates Q_MAIN's required heatsink thermal resistance, using Infineon's IPP110N20N3 datasheet thermal values [39].

$$R_{\theta HA} = \frac{175^{\circ}\text{C} - 40^{\circ}\text{C}}{6\text{W}} - 0.5^{\circ}\text{C}/\text{W} = 22.5^{\circ}\text{C}/\text{W} \quad (8.5)$$

Infineon's IPP110N20N3 datasheet does not specify a case-to-heatsink thermal resistance, thus this project omits $R_{\theta CS}$ from Q_MAIN's heatsink thermal resistance calculation. Like IXYS's IXFK230N20T TO-264 power MOSFET, this project expects the IPP110N20N3's case-to-heatsink thermal resistance to be fairly small and thus be neglects it.

Equation (8.6) calculates DSNUB1's required heatsink thermal resistance, using Microsemi's APT30S20BG datasheet thermal values [61].

$$R_{\theta HA} = \frac{150^{\circ}\text{C} - 40^{\circ}\text{C}}{2\text{W}} - 0.58^{\circ}\text{C} = 54.42^{\circ}\text{C}/\text{W} \quad (8.6)$$

Microsemi's APT30S20BG datasheet does not specify a case-to-heatsink thermal resistance, thus this project omits $R_{\theta CS}$ from DSNUB1's heatsink thermal resistance calculation. Like IXYS's IXFK230N20T TO-264 power MOSFET, this project expects the APT30S20BG's case-to-heatsink thermal resistance to be fairly small and thus neglects it.

For all diodes using Micro Commerical's MBR20100CT Schottky diode (D_MAIN1 to 4 and DSNUB2 to 4), Micro Commerical's datasheet does not specify either a junction-to-case thermal resistance or a case-to-heatsink thermal resistance [40]. This project omits $R_{\theta CS}$ from the heatsink thermal resistance calculation for such diodes, however this project still uses the typical TO-220-3 package transistor $3^{\circ}\text{C}/\text{W}$ thermal resistance value for $R_{\theta JC}$ [62].

Equation (8.7) calculates D_MAIN1, D_MAIN2, D_MAIN3 and D_MAIN4's required heatsink thermal resistance.

$$R_{\theta HA} = \frac{150^{\circ}\text{C}-40^{\circ}\text{C}}{3.5\text{W}} - 3^{\circ}\text{C}/\text{W} = 28.43^{\circ}\text{C}/\text{W} \quad (8.7)$$

Equation (8.8) calculates DSNUB2's required heatsink thermal resistance.

$$R_{\theta HA} = \frac{150^{\circ}\text{C}-40^{\circ}\text{C}}{4\text{W}} - 3^{\circ}\text{C}/\text{W} = 24.5^{\circ}\text{C}/\text{W} \quad (8.8)$$

Equation (8.9) calculates DSNUB3 and DSNUB4's required heatsink thermal resistance.

$$R_{\theta HA} = \frac{150^{\circ}\text{C}-40^{\circ}\text{C}}{4.1\text{W}} - 3^{\circ}\text{C}/\text{W} = 23.83^{\circ}\text{C}/\text{W} \quad (8.9)$$

After calculating component heatsink thermal resistance ratings, this project selects commercially available heatsinks that properly suit the SEPIC's semiconductor components as well as have lower than calculated thermal resistance ratings.

8.3 Heatsink Selection and Component Temperature Recalculation

For Q_OVP and Q_MAIN_ALT, this project selects Ohmite's MV-102-55E TO-247/TO-264 heatsink. Figure 8.1 shows the heatsink's physical appearance [63].

Figure 8.1: Ohmite MV-102-55E TO-247 and TO-264 Heatsink [63]

Ohmite's MV-102-55E heatsink dissipates heat from up to four simultaneous TO-247 or TO-264 components, and has a $6^{\circ}\text{C}/\text{W}$ natural thermal resistance [63]. When attached to a 400 LFM active cooling device, its thermal resistance decreases to $2^{\circ}\text{C}/\text{W}$ [63]. Even without an attached active cooling device, the MV-102-55E's thermal resistance specification is much lower than the required calculated thermal resistance for Q_OVP and Q_MAIN_ALT under the highest stress operating conditions, thus it is reasonable for those components. The smaller version of this heatsink that holds only one component, the MV-xxx-27E series, has a similar natural thermal resistance rating and is more ideal for this project's SEPIC. However, the MV-xxx-27E was unavailable on Digi-Key when

this project created its SEPIC's bill of materials, thus this project uses the larger -55E heatsink instead.

For Q_MAIN, this project selects Aavid Thermalloy's 531202B02500G TO-220-3 heatsink. Figure 8.2 shows the heatsink's physical appearance [64].

Figure 8.2: Aavid Thermalloy 531202B02500G TO-202 and TO-220 Heatsink [64]

Aavid Thermalloy's 531202B02500G heatsink supports TO-202 and TO-220 components and has a $7.5^{\circ}\text{C}/\text{W}$ natural thermal resistance [64]. When attached to a 600 LFM active cooling device, its thermal resistance decreases to $4^{\circ}\text{C}/\text{W}$ [64]. Likewise with the MV-102-55E used for Q_OVP and Q_MAIN_ALT, the 531202B02500G's thermal resistance specification without any active cooling is much lower than the required calculated thermal resistance for Q_MAIN under the highest stress operating conditions, thus it is reasonable for Q_MAIN.

This project selects Ohmite's WA-T247-101E TO-247 heatsink for DSNUB1.

Figure 8.3 shows the heatsink's physical appearance [65].

Figure 8.3: Ohmite WA-T247-101E TO-247 Heatsink [65]

Ohmite WA-T247-101E heatsink supports TO-247 components and has a $7^{\circ}\text{C}/\text{W}$ natural thermal resistance [65]. When attached to a 500 LFM active cooling device, its thermal resistance oddly increases to $8^{\circ}\text{C}/\text{W}$ [65]. Thus, active cooling is not required for the WA-T247-101E and in many cases adding active cooling worsens thermal performance. Its natural thermal resistance is much lower than the required calculated thermal resistance for DSNUB1 under the highest stress operating conditions, thus it is reasonable for DSNUB1.

For the remaining D_MAIN and DSNUB diodes, this project selects Aavid Thermalloy's 574502B00000G TO-220-3 heatsink. Figure 8.4 shows the heatsink's physical appearance [66].

Figure 8.4: Aavid Thermalloy 574502B00000G TO-220 Heatsink [66]

Aavid Thermalloy's 574502B00000G heatsink cools TO-220 components and has a 21.2°C/W natural thermal resistance. When attached to a 400 LFM active cooling device, its thermal resistance oddly decreases to 8°C/W [66]. Although this heatsink has the worst thermal performance out of all the selected ones, its installation process is simple and it does not require any additional clips, bolts or screws. Its natural thermal resistance is still lower than the required calculated thermal resistance for the DMAIN1-4 diodes and DSNUB2-4 diodes under the highest stress operating conditions, thus it is reasonable for those components.

This project now determines each component's operating junction temperature by using (8.10) for recalculating maximum junction temperatures. (8.10) uses thermal resistance ratings gathered from the selected heatsinks.

$$T_{JMAX} = P_{MAX}(R_{\theta JC} + R_{\theta CS} + R_{\theta HA}) + T_A \quad (8.10)$$

Equation (8.11) calculates Q_OVP's maximum junction temperature with the Ohmite MV-102-55E heatsink.

$$T_{JMAX} = 12W(0.09 + 0.15 + 6) + 40^\circ\text{C} = 114.88^\circ\text{C} \quad (8.11)$$

Equation (8.12) calculates Q_MAIN_ALT's maximum junction temperature with the Ohmite MV-102-55E heatsink.

$$T_{JMAX} = 6W(0.09 + 0.15 + 6) + 40^{\circ}\text{C} = 77.44^{\circ}\text{C} \quad (8.12)$$

Equation (8.13) calculates Q_MAIN's maximum junction temperature with Aavid Thermalloy's 531202B02500G heatsink.

$$T_{JMAX} = 6W(0.5 + 7.5) + 40^{\circ}\text{C} = 88^{\circ}\text{C} \quad (8.13)$$

Equation (8.14) calculates DSNUB1's maximum junction temperature with the Ohmite WA-T247-101E heatsink.

$$T_{JMAX} = 2W(0.58 + 7) + 40^{\circ}\text{C} = 55.16^{\circ}\text{C} \quad (8.14)$$

Equation (8.15) calculates D_MAIN1, D_MAIN2, D_MAIN3 and D_MAIN4's maximum junction temperature with Aavid Thermalloy's 574502B00000G heatsink.

$$T_{JMAX} = 3.5W(3 + 21.2) + 40^{\circ}\text{C} = 124.7^{\circ}\text{C} \quad (8.15)$$

Equation (8.16) calculates DSNUB2's maximum junction temperature with Aavid Thermalloy's 574502B00000G heatsink.

$$T_{JMAX} = 4W(3 + 21.2) + 40^{\circ}\text{C} = 136.8^{\circ}\text{C} \quad (8.16)$$

Equation (8.17) calculates DSNUB3 and DSNUB4's maximum junction temperature with Aavid Thermalloy's 574502B00000G heatsink.

$$T_{JMAX} = 4.1W(3 + 21.2) + 40^{\circ}\text{C} = 139.22^{\circ}\text{C} \quad (8.17)$$

Snubber diodes DSNUB2, DSNUB3 and DSNUB4 operate close to their maximum specified junction temperatures (150°C) under full load conditions, however, this project selected their heatsinks for saving physical space. Saving physical space reduces

converter costs by preventing the SEPIC's PCB from becoming too physically large.

Each selected heatsink suffices for properly cooling the SEPIC's semiconductor components and help allow for prolonged SEPIC use. After selecting all the heatsinks for its SEPIC, this project can now design its SEPIC's PCB layout.

CHAPTER 9: PCB LAYOUT

9.1 SEPIC First Revision PCB Layout Design

Poor PCB layouts jeopardize the functionality of well-designed circuits. The PCB layout for this project's SEPIC followed design guidelines outlined in [67-70] for suppressing any undesirable side-effects such as capacitive crosstalk or resonance. This project's SEPIC's PCB also keeps high power and high frequency switching components such as diodes, MOSFETs and inductors physically as far away as possible from the low-power control components (e.g. the LTC1871 controller). Figure 9.1 shows a boost converter layout example featured in Linear Technology's LTC1871 datasheet [35].

Figure 9.1: Linear Technology's Suggested LTC1871 Boost Converter Layout [35]

Because a SEPIC's overall layout is very similar to that of a boost converter [26, 27], this project uses figure 9.1 as a guideline for determining this project's SEPIC's component

and trace placement. The SEPIC's first revision PCB layout is similar to that of Linear Technology's suggested LTC1871 boost converter layout in figure 9.1. However, this project reverses the SEPIC's physical input voltage (V_{IN}) location (it is on the left side instead of the right side of the PCB as in figure 9.1) for accommodating the LT4356-1 input protection circuit. This project also minimizes PCB manufacturing costs by manufacturing its SEPIC layout on a two-layer PCB without any silkscreen or solder masking. Silkscreen and solder masking are necessary for a production level or retail SEPIC board for easing troubleshooting as well as for providing additional safety to the end user (from the lack of exposed pads). However, this project does not require silkscreen and solder masking for a prototyping board, which is the PCB type this project uses and tests with the rest of the SEPIC components. Figure 9.2 shows this project's SEPIC first revision PCB layout:

Figure 9.2: SEPIC First Revision PCB Layout

Red designates the PCB's top copper layer while green designates its bottom copper layer. Blue designates its silkscreen layer, but does not show up on the manufactured PCB. Because of this, L_1 and L_2 can freely extend past the PCB's physical borders, as the Vishay IHV28BZ60 physically stands upward. Figures 9.3, 9.4 and 9.5 show the SEPIC's individual top and bottom copper and silkscreen layers.

Figure 9.3: SEPIC First Revision PCB Layout Top Copper Layer

Figure 9.4: SEPIC First Revision PCB Layout Bottom Copper Layer

Figure 9.5: SEPIC First Revision PCB Layout Silkscreen Layer

The PCB's bottom copper layer consists almost entirely of solid ground planes, with separate planes for the elliptical battery, low-power control circuit components such as the LTC1871 controller and high power components such as L_2 . This preliminary layout does not yet have each ground plane bridged together. The top copper layer consists of narrow (0.010" to 0.020" width) traces for the control circuit components and wide traces (0.200" and wider) for the high power components. The high power components require

wide traces for effective high current flow capability, namely at the SEPIC's input, output and switching (Q_{MAIN} and C_{INT} connection) nodes. Additionally, the PCB also contains holes for mounting test points so that this project's author can easily measure critical node voltages. The silkscreen layer also labels each test point; for example, $Q_{\text{MAIN_D}}$ corresponds to the test point located at Q_{MAIN} 's drain node. This project, however, would remove these test point mounting holes in a final production version of its SEPIC's PCB, as the test points are only necessary for PCB prototyping and testing purposes. Inductor current paths contain break points for inserting wire loops for easing the inductor current measurement process. This PCB contains wire loop insertion points at L_1 , L_2 and L_{SNUB} 's inputs. This layout also provides additional placeholder footprints for additional output capacitors. Furthermore, it also provides a placeholder footprint for $Q_{\text{MAIN_ALT}}$, in case the IPP110N20N3 power MOSFET used for Q_{MAIN} is not suitable for this SEPIC during hardware testing. The layout also contains break points at the SEPIC's elliptical input, 12V battery input and output nodes for soldering fuses, in order to abide by UL, IEEE, PG&E and NEC safety requirements.

One major flaw with this preliminary layout is that it neglects heatsink spacing as well as chassis mounting holes and voltage source connection points with the proper mechanical dimensions. Furthermore, this layout revision is also missing R_{VCC1} and R_{VCC2} 's footprints. This project, however, did not consider inserting R_{VCC1} and R_{VCC2} into the SEPIC's schematic until the PCB layout's second revision. The SEPIC's second revision layout addresses these problems by including heatsink footprints in its silkscreen

layer, R_{VCC1} and R_{VCC2} footprints and chassis mounting holes and holes for inserting banana binding posts to crucial voltage source, load and ground connection points.

9.2 SEPIC Second Revision PCB Layout Design

The SEPIC's second revision PCB layout includes all proper heatsink footprints, semiconductor components and chassis mounting holes and device connection points. Present are eight chassis mounting holes located on the PCB corners and edges. Each chassis mounting hole has a 0.125" diameter and supports #5-40 and #5-44 size screws [71]. Some other major differences with the SEPIC's second revision PCB layout (compared to its first revision) are that this revision replaces wide power path traces with solid copper planes for trace editing ease, and widens physically long low power (small signal traces) at their power component connection end points. Such traces are 0.030" and wider for minimizing parasitic resistance caused by long narrow trace lengths. Figure 9.6 shows the SEPIC's second revision PCB layout.

Figure 9.6: SEPIC Second Revision PCB Layout

This layout revision includes footprints for all components from the finalized SEPIC schematic in figure 7.9, as well as heatsink mounting holes and connection points for voltage sources and electronic loads. Additionally, this layout also includes a break point at the LT4356-1 protection circuit's output (as well as an alternate C_{IN} connection point) in case it ever requires bypassing for troubleshooting purposes during the hardware testing process. The layout also retains inductor break points for soldering wire loops.

Furthermore, this layout also contains a connection point for connecting C_{IN} in series with the Precor elliptical trainer's 10Ω resistor coils, for reducing RMS current as this report denotes in section 7.5. For synthetic hardware tests (using a power supply as the SEPIC's main input voltage source instead of the elliptical), this project uses a solderable jumper for connecting C_{IN} 's negative terminal to ground. The Q_MAIN and Q_MAIN_ALT heatsink footprints overlap each other because for proper operation, this SEPIC would only use either transistor, but not both simultaneously. The production-level PCB layout would obviously remove one of the footprints depending on which transistor yields better hardware test results. This layout revision also removes solderable fuse connection points at the elliptical input and output nodes, as for fuse replacement ease this project decided on using external blade fuses and blade fuse holders connected to those nodes. Therefore, no one should ever connect this board to a voltage source or load without connecting blade fuses to the elliptical input and output nodes. Although the LT4356-1 circuit protects the SEPIC's input, one major flaw that this project realized late into implementation is that there this converter does not have any output protection (such as an output current limiter circuit). Therefore this SEPIC requires fuses, not only to conform to UL, IEEE, PG&E and NEC safety requirements, but also for end-user safety. This layout still requires a solderable fuse for the 12V battery input node, however. Figures 9.7, 9.8 and 9.9 show the second revision SEPIC's individual top and bottom copper and silkscreen layers:

Figure 9.7: SEPIC Second Revision PCB Layout Top Copper Layer

Figure 9.8: SEPIC Second Revision PCB Layout Bottom Copper Layer

This second revision PCB layout also bridges each ground plane together, based on component return current flow direction.

Likewise with L_1 and L_2 , Q_OVP's heatsink can freely extend beyond the PCB's physical borders, as it physically stands upward and only requires two mounting holes for

physical security. Furthermore, as mentioned in section 9.1, the prototype PCB used for this SEPIC's hardware testing does not have the silkscreen layer.

This second revision board also measures $5.275'' \times 5.800''$, corresponding to a 30.595 in.^2 area. It is quite large compared to the previously used off-the-shelf Vicor Maxi 28V DC-DC converter (which is $4.60'' \times 2.20''$), but is more functional with the EHFEM system than the Vicor converter. This PCB also costs \$124.84 for two boards, with \$84.24 for a single board. This results in a \$259.07 total converter cost, as table D-1 shows. Again, it is more expensive even without a chassis when compared to the Vicor Maxi 28V DC-DC converter (which costs \$208.00 [1]), but again, its functionality compensates for its cost. Table D-1 shows this finalized SEPIC's BOM with the PCB and heatsink costs included.

Table D-1: Finalized SEPIC BOM with PCB and Heatsink Costs

PCBs	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Custom 2-layer 5.8" x 5.275"	1	84.42	84.42	PCB	2-layer	N/A	N/A
Total Cost (\$)			84.42				
Controller I.C.s:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
LTC1871HMS (10-MSOP)	1	4.83	4.83	U1	10-MSOP	N/A	N/A
LT4356-1HMS (10-MSOP)	2	3.70	7.40	U2	10-MSOP	N/A	N/A
Total Cost (\$)			12.23				
Transistors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Infiniteon IPP110N20N3	1	8.06	8.06	Q_MAIN	TO-220-3	2.45W	4.01W
IXYS IXFK230N20T	1	9.50	9.50	Q_OVP	TO-264	1.625kW	1.625kW
Total Cost (\$)			17.56				
Diodes:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
MBR20100CT (Schottky)	7	0.88	6.16	D_MAIN1, D_MAIN2, D_MAIN3, D_MAIN4, DSNUB2, DSNUB3, DSNUB4	TO-220-3	31.11A, 110V	27.72A, 56.46V
Microsemi APT30S20BG (Schottky)	1	3.67	3.67	DSNUB1	TO-247	39.43A, 15.95V	27.89A, 9.58V
1N5378B (Zener)	1	0.57	0.57	TSD	Through-Hole	100V	100V
SML-LX1206SRC-TR (LED)	1	0.43	0.43	FLT_LED	1206	1.2V min. forward drop	1.2V min. forward drop
Total Cost (\$)			10.83				
Heatsinks:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
531202B02500G (TO-220)	1	1.38	1.38	Q_MAIN heatsink	TO-220	N/A	N/A
MV-102-55E (TO-247 and TO-264)	1	7.39	7.39	Q_OVP heatsink	TO-264	N/A	N/A
WA-T247-101E (TO-247)	1	2.07	2.07	DSNUB1 heatsink	TO-247	N/A	N/A
574502B00000G (TO-220)	7	0.27	1.89	D_MAIN1, D_MAIN2, D_MAIN3, D_MAIN4, DSNUB2, DSNUB3, DSNUB4	TO-220	N/A	N/A
Total Cost (\$)			12.73				

Table D-1 (Continued)

Fuses:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Radio Shack 270-270-1234 30A Blade Fuse Holder	2	2.99	5.98	Fuse Holder for Input and Output Fuse	N/A	65V, 25A	5V, 25A
Cooper Bussmann 270-1085 30A Blade Fuse	1	0.73	0.73	Input Fuse	Blade Fuse	65V, 25A	5V, 25A
Cooper Bussmann 270-1081 10A Blade Fuse	1	0.73	0.73	Output Fuse	Blade Fuse	36V, 8A	5V, 0.5A
Bourns SF-0402F200-2 2A Fuse	1	0.68	0.68	Elliptical Battery Fuse	0402	12V, 2A	12V, 2A
Total Cost (\$)			8.12				
Resistors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
274k, 1/8W, 1%, ERJ-6ENF2743V	1	0.07	0.07	RRUN1_MAIN	0805	0.521mW	0.522mW
255k, 1/8W, 1%, ERJ-6ENF2553V	1	0.07	0.07	RRUN1_OVP	0805	16.289mW	0.094mW
220k, 1/8W, 1%, ERJ-6ENF2203V	1	0.07	0.07	RFREQ	0805	0.002mW	0.002mW
133k, 1/8W, 1%, ERJ-6ENF1333V	1	0.07	0.07	RRUN2_MAIN	0805	0.115mW	0.122mW
115k, 1/8W, 1%, ERJ-6ENF1153V	1	0.07	0.07	RFB1	0805	11.12mW	12.28mW
33.2k, 1/8W, 1%, ERJ-6ENF3322V	1	0.07	0.07	RTH	0805	28.32mW	0.028mW
4.99k, 1/8W, 1%, ERJ-6ENF4991V	1	0.07	0.07	RRUN2_OVP	0805	0.319mW	0.002mW
3.92k, 1/8W, 1%, RMCF0805FT3K92	1	0.04	0.04	RFB2	0805	0.387mW	0.427mW
3.6k, 1/8W, 1%, ERJ-6ENF3601V	1	0.07	0.07	RFLT	0805	32.4mW	32.4mW
887, 1/8W, 1%, ERJ-6ENF8870V	1	0.07	0.07	RVCC1	0805	< 125mW	< 125mW
680, 1/8W, 5%, ERJ-6GEYJ681V	1	0.04	0.04	RVCC2	0805	< 125mW	< 125mW
10, 1/8W, 1%, ERJ-6ENF10R0V	1	0.07	0.07	RGATE_OVP	0805	13.05mW	0W
0.005, 5W, 1%, OARSXPR005FLF	1	1.42	1.42	RSNS	SMT (Custom)	410mW	1.132W
0.002, 5W, 1%, WSLP39212L000FEB	1	3.07	3.07	RSNS_OVP	SMT (Custom)	64mW	1.000W
Total Cost (\$)			5.27				
Inductors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Vishay IHV28BZ60 (60μH, I _{SAT} =28A)	2	20.30	40.60	L1, L2	Through-Hole	30.68A (max.), 65V	22.37A (max.), 52.22V
Vishay IHLP4040DZERR36M01 (360nH, I _{SAT} =60A)	1	2.52	2.52	LSNUB	SMT (Custom)	29.3A (max.), 106.25V	17.85A (max.), 55.35V
Total Cost (\$)			43.12				

Table D-1 (Continued)

Capacitors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
470μ, Electrolytic, 400V, B43504A9477M	1	13.34	13.34	COUT	Through-Hole	36V, 5.43A (RMS)	36V, 2.368A (RMS)
1000μ, Electrolytic, 100V, UPW2A102MHD	1	3.03	3.03	CIN	Through-Hole	65V, 1.886pA (RMS)	5V, 227.67fA (RMS)
47μ, Electrolytic, 100V, AFK476M2AH32T-F	1	2.35	2.35	CCLAMP	SMT (Custom)	65.68V, 119.05mA (RMS)	5V, 42.067mA (RMS)
33μ, Electrolytic, 35V, EEE-HA1V330WP	1	0.63	0.63	CBAT	SMT (Custom)	12V, 206.56fA (RMS)	12V, 442.09fA (RMS)
15μ, Ceramic, 250V, KHD251E156M99C0B00	1	25.78	25.78	CINT	Through-Hole	77.05V, 3.67A (RMS)	21.22V, 1.404A (RMS)
10μ, Ceramic, 50V, UMK325C7106MM-T	3	0.96	2.88	COUT2, COUT3, COUT4	1210	36V, 2.758A (RMS)	36V, 2.016A (RMS)
10μ, Ceramic, 25V, TMK316B7106KL-TD	2	0.58	1.16	CBAT2, CBAT3	1206	12V, 963.63fA (RMS)	12V, 1.211pA (RMS)
4.7μ, Ceramic, 100V, C4532X7S2A475M	5	1.75	8.75	CINT2, CINT3, CINT4, CINT5, CINT6	1812	77.05V, 1.22A (RMS)	21.22V, 1.22A (RMS)
4.7μ, Ceramic, 16V, EMK212B7475KG-T	1	0.38	0.38	CVCC	0805	5.23V, 153.1mA (RMS)	5.23V, 170.84mA (RMS)
0.56μ, Ceramic, 25V, C0805C564K3RACTU	1	0.84	0.84	CSNUB1	0805	11.33V, 3.629A (RMS)	11.00V, 1.577A (RMS)
0.1μ, Ceramic, 100V, CGA3E3X7S2A104K	7	0.30	2.10	CIN2, CIN3, CCLAMP2, CCLAMP3, CCLAMP4, CBAT4, CBAT5	0603	65V, 1.000mA (RMS)	5V, 0.285mA (RMS)
0.1μ, Ceramic, 100V, HMK212B7104KG-T	2	0.30	0.60	CINT7, CINT8	0805	77.05V, 47.184mA (RMS)	21.22V, 17.59mA (RMS)
0.1μ, Ceramic, 50V, C1608X7R1H104M	2	0.16	0.32	COUT_CPH1, COUT_CPH2	0603	36V, 63.00mA (RMS)	36V, 46.00mA (RMS)
10000p, Ceramic, 250V, C2012X7R2E103K	1	0.33	0.33	CSNUB2	0805	103.44V, 1.619A (RMS)	56.06V, 756.43mA (RMS)
6800p, Ceramic, 100V, C1608X8R2A682K	1	0.36	0.36	CGATE_OVP	0603	77.47V, 4.796mA (RMS)	11.331V, 0.006mA (RMS)
6800p, Ceramic, 25V, C1005X7R1E682K	1	0.09	0.09	CTH2	0402	1.232V, 0.110mA (RMS)	1.232V, 0.030mA (RMS)
56p, Ceramic, 10V, C0402C560J8GACTU	1	0.35	0.35	CTH1	0402	1.232V, 0.390mA (RMS)	1.232V, 0.009mA (RMS)
47p, Ceramic, 10V, 0603ZA470DAT2A	5	0.30	1.50	CFB, CSNS, CTMR, CRUN_OVP, CRUN_MAIN	0603	1.5V, 0.028mA (RMS)	1.307V, 0.020mA (RMS)
Total Cost (\$)			64.79				
Component Count	79						
Total Converter Cost (\$)	259.07						

After receiving the PCB via ExpressPCB, this project's author performed an exhaustive continuity test for ensuring that the PCB contained proper connection points and that there were no undesirable or potentially fatal bridge points. As expected, the PCB passed the continuity test upon first receipt. After verifying the PCB, this project's author soldered SEPIC components onto it and began the hardware testing process.

CHAPTER 10: HARDWARE TESTING

10.1 Component Soldering Process

This SEPIC's component soldering process only took five days. This project's author began by first soldering the physically smaller low-power components such as the LTC1871 controller and gradually moved on to larger, high-power components such as L_1 and L_2 . Lack of fine solder tips in Cal Poly's power electronics laboratory impeded the soldering process, but for the most part it was not difficult because this SEPIC did not have components requiring a reflow oven. Furthermore, during this SEPIC's initial soldering process, this project's author accidentally removed a thin copper trace connecting the LTC1871 controller's FREQ pin, thus the author had to spend an additional three hours re-bridging the connection using a very thin wire. That solution sufficed, however, as the LTC1871's FREQ pin is not very sensitive to wire width, as long as the wire does not greatly modify the resistance the pin senses. Figures 10.1 and 10.2 show this project's SEPIC with all components soldered.

Figure 10.1: First Revision SEPIC, Top View

Figure 10.2: First Revision SEPIC, Angled View

After completing the SEPIC's component soldering process, this project begins its first hardware testing round.

10.2 No-Load Testing

This project's first set of hardware tests consist of running the SEPIC with no load (open-circuit output). These tests ensure that both the SEPIC's LTC1871 and LT4356-1 controllers function, as well as crucial SEPIC components such as Q_MAIN and L_1 and L_2 . Figure 10.3 shows the block diagram for this project's SEPIC's no-load tests.

Figure 10.3: SEPIC No-Load Test Block Diagram

These no load tests utilize a TPS-4000 dual rail power supply acting as a 12V source for biasing the SEPIC's LTC1871 controller, as well as a GW Instek DC power supply acting as the SEPIC's main input source and a BK Precision 150W electronic load for obtaining output voltage measurements. The GW Instek power supply has a 3A and 190W limit, however, those pitfalls do not matter for no-load tests. Figure 10.4 shows the SEPIC's no-load test wiring set up.

Figure 10.4: SEPIC No-Load Test Wiring

For this project's no-load tests, this project uses a 30A input protection fuse but no output protection fuse. The LT4356-1 protection circuit described in Chapter 7 limits the SEPIC's peak pulsed input current to 25A, thus the 30A input protection fuse suffices. However, these no-load tests do not require an output protection fuse because these tests do not cause any current flow through the SEPIC's output node. Figure 10.5 shows instrument readouts with all sources energized and the SEPIC running.

Figure 10.5: SEPIC No-Load Test at 36V Input

From left to right are the BK Precision electronic load, TPS-4000 power supply and the GW Instek power supply. With no load, the SEPIC outputs 37.38V through the entire 5-65V input range from the GW Instek power supply. This is slightly above the 36V output requirement described in section 1.2, but this extra voltage allows for headroom when the SEPIC drives a load, as typically its output voltage should reduce by a small margin when it drives a load. At around 63V input, however, the SEPIC's LT4356-1 protection LED turned on, indicating an overvoltage input condition. The LT4356-1 circuit was very sensitive to input voltage and at 65.1V input it shut off voltage and current flow to the SEPIC. When the input voltage reduced to 63V, the SEPIC functioned again and regulated a 37.38V output. This confirmed that the LT4356-1 protection circuit was

properly functioning. Therefore, this project's SEPIC functions with no load and this project's author properly connected the SEPIC's components during the soldering process.

10.3 Load Testing

The next SEPIC hardware testing round determines the SEPIC's load regulation abilities. As mentioned in earlier chapters, this project does not require the SEPIC driving its specified 8A maximum load current at all input voltages – only when it receives the maximum 65V input from the Precor elliptical trainer. The first load test set uses an electronic load in constant current mode for determining the SEPIC's load regulation abilities. Later tests use the electronic load in constant voltage mode for determining the SEPIC's maximum output load current for maintaining 36V output over this project's specified 5-65V input range.

10.3.1 Electronic Load in Constant Current Mode

Figure 10.6 shows the SEPIC's initial load test block diagram.

Figure 10.6: SEPIC Initial Load Test Block Diagram

The initial load tests use the same setup as the no-load tests, except this time this project uses a 7.5A blade fuse for protection. No commercially-available 8A blade fuses exist (as only rating increments of 0.5A exist for such blade fuses) therefore later higher current tests use 10A blade fuses. This project uses the oscilloscopes and current probe amplifiers in the next section for troubleshooting purposes. Because the GW Instek power supply has a 3A limit, this first set of load tests only tests loads up to 2.3A. Furthermore, these tests use an HP 300W electronic load because this SEPIC's 288W maximum output power rating exceeds the BK Precision electronic load's (used for the no-load tests) 150W rating. Tables E-1 to E-16 show these test results at each input voltage and output load current.

Table E-1: SEPIC Initial Load Tests with 0.2A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.2A Load Tests:							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
10	1.019	29.73	0.2	10.19	5.946	58.351	9.814
12	1.077	37.34	0.2	12.924	7.468	57.784	11.142
15	0.848	37.34	0.2	12.72	7.468	58.711	17.689
20	0.621	37.34	0.2	12.42	7.468	60.129	32.206
25	0.487	37.34	0.2	12.175	7.468	61.339	51.335
27	0.454	37.34	0.2	12.258	7.468	60.923	59.471

Table E-2: SEPIC Initial Load Tests with 0.25A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.25A Load Tests:							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
12	1.21	33.96	0.25	14.52	8.49	58.471	9.917
15	1.038	37.34	0.25	15.57	9.335	59.955	14.451
20	0.76	37.34	0.25	15.2	9.335	61.414	26.316
25	0.6	37.34	0.25	15	9.335	62.233	41.667
27	0.56	37.34	0.25	15.12	9.335	61.739	48.214

Table E-3: SEPIC Initial Load Tests with 0.3A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.3A Load Tests:							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	0.666	33.96	0.3	17.982	10.188	56.657	40.541
36	0.512	37.34	0.3	18.432	11.202	60.775	70.313

Table E-4: SEPIC Initial Load Tests with 0.35A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.35A Load Tests:							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	0.772	33.96	0.35	20.844	11.886	57.024	34.974
36	0.591	37.34	0.35	21.276	13.069	61.426	60.914

Table E-5: SEPIC Initial Load Tests with 0.4A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.4A Load Tests:							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	0.876	33.96	0.4	23.652	13.584	57.433	30.822
36	0.672	37.34	0.4	24.192	14.936	61.739	53.571

Table E-6: SEPIC Initial Load Tests with 0.45A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.45A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	1	33.96	0.45	27	15.282	56.600	27.000
36	0.751	37.34	0.45	27.036	16.803	62.150	47.936

Table E-7: SEPIC Initial Load Tests with 0.5A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.5A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	1.085	33.96	0.5	29.295	16.98	57.962	24.885
36	0.831	37.34	0.5	29.916	18.67	62.408	43.321

Table E-8: SEPIC Initial Load Tests with 0.6A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.6A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	1.325	33.96	0.6	35.775	20.376	56.956	20.377
36	0.99	37.34	0.6	35.64	22.404	62.862	36.364

Table E-9: SEPIC Initial Load Tests with 0.7A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.7A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	1.526	33.96	0.7	41.202	23.772	57.696	17.693
36	1.15	37.34	0.7	41.4	26.138	63.135	31.304

Table E-10: SEPIC Initial Load Tests with 0.8A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.8A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	1.727	33.96	0.8	46.629	27.168	58.264	15.634
36	1.305	37.34	0.8	46.98	29.872	63.585	27.586

Table E-11: SEPIC Initial Load Tests with 0.9A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

0.9A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	1.923	33.96	0.9	51.921	30.564	58.866	14.041
36	1.458	37.34	0.9	52.488	33.606	64.026	24.691

Table E-12: SEPIC Initial Load Tests with 1A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

1A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
18	1.776	21.18	1	31.968	21.18	66.254	10.135
27	2.119	33.96	1	57.213	33.96	59.357	12.742
36	1.61	37.34	1	57.96	37.34	64.424	22.360
50	1.155	37.18	1	57.75	37.18	64.381	43.290

Table E-13: SEPIC Initial Load Tests with 1.5A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

1.5A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
27	2.613	31.57	1.5	70.551	47.355	67.122	10.333
36	2.3	36.27	1.5	82.8	54.405	65.707	15.652
50	1.705	37.15	1.5	85.25	55.725	65.367	29.326

Table E-14: SEPIC Initial Load Tests with 2A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

2A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
36	2.85	34.57	2	102.6	69.14	67.388	12.632
50	2.24	37.04	2	112	74.08	66.143	22.321

Table E-15: SEPIC Initial Load Tests with 2.2A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

2.2A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
50	2.204	34.39	2.2	110.2	75.658	68.655	22.686

Table E-16: SEPIC Initial Load Tests with 2.3A Constant Current Electronic Load, TPS-4000 and GW Instek Power Supplies

2.3A Load Tests							
V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
50	2.293	32.9	2.3	114.65	75.67	66.001	21.805

V_{IN} denotes the SEPIC's main input voltage, I_{IN} denotes its input current (read from the GW Instek power supply), V_{OUT} denotes its output voltage, I_{OUT} denotes its output current, P_{IN} denotes its input power, P_{OUT} denotes its output power, η denotes its overall

efficiency, and R_{IN} denotes its input resistance at the specified input voltage and output load current. These initial load tests do not take 12V battery power dissipation into account because this project's author neglected using a method for accurately measuring the TPS-4000 power supply's current. However, the TPS-4000 power supply's current was very small during these load tests, therefore overall converter efficiency would not change by any significant margin if the battery's power dissipation was included. Therefore, these tests safely neglect the 12V battery's power dissipation. These efficiency results, however, are below the project's specifications outlined in section 1.2 and require improvement. Also, at input voltages above 52V, the converter does not tolerate very high current loads (only up to 0.09A maximum), thus this project discards those results. Therefore this initial converter has load regulation problems at input voltages above 52V. For the most part, overall converter efficiency increases with increasing input voltage and load current. Furthermore, these varying load currents do not allow for the SEPIC maintaining a constant 10Ω input resistance, therefore this project does not expect these same results if the electronic load was in constant voltage mode (instead of constant current mode as in this initial load test). However, this project must address the load regulation problem at input voltages above 52V, otherwise this SEPIC will not function with the entire EHFEM system. The next load test set uses the HP electronic load in constant voltage mode for determining the maximum load current that the SEPIC drives while maintaining 36V output over the SEPIC's specified input range.

10.3.2 Electronic Load in Constant Voltage Mode

Table E-17 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V). Table E-18 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-17: SEPIC Initial Load Tests with Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)
5	0.89	36.02	0.05	4.450	1.801	40.472	5.618
10	0.996	36.02	0.14	9.960	5.043	50.631	10.040
12	1.185	36.02	0.22	14.220	7.924	55.727	10.127
15	1.453	36.02	0.35	21.795	12.607	57.844	10.323
18	1.745	36.02	0.53	31.410	19.091	60.779	10.315
20	1.939	36.02	0.67	38.780	24.133	62.232	10.315
25	2.402	36.02	1.08	60.050	38.902	64.782	10.408
27	2.565	36.02	1.26	69.255	45.385	65.533	10.526
30	2.796	36.02	1.54	83.880	55.471	66.131	10.730
36	2.241	36.02	1.46	80.676	52.589	65.186	16.064
40*	3.065*	36.02*	2.55*	122.600*	91.851*	74.919*	13.051*
45	2.753	36.02	2.32	123.885	83.566	67.455	16.346
50	2.285	36.06	2.09	114.250	75.365	65.965	21.882
52	2.297	36.06	2.19	119.444	78.971	66.116	22.638
60	0.112	36.01	0.09	6.720	3.241	48.228	535.714
* GW-Instek Current Limited							

Table E-18: SEPIC Initial Load Tests with 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)	Training Resistance Level
8.08	0.797	36.02	0.09	6.440	3.242	50.340	10.138	2
11.54	1.119	36.02	0.19	12.913	6.844	52.998	10.313	3
13.49	1.295	36.02	0.27	17.470	9.725	55.671	10.417	4
15.88	1.502	36.02	0.38	23.852	13.688	57.386	10.573	5
18.27	1.718	36.02	0.52	31.388	18.730	59.674	10.634	6
21	1.99	36.02	0.72	41.790	25.934	62.059	10.553	7
23.3	2.202	36.02	0.9	51.307	32.418	63.185	10.581	8
26.2	2.452	36.02	1.15	64.242	41.423	64.479	10.685	9
29.1	2.695	36.02	1.43	78.425	51.509	65.679	10.798	10
31.8	2.517	36.02	1.46	80.041	52.589	65.703	12.634	11
34.3	2.112	36.02	1.29	72.442	46.466	64.142	16.241	12
35.1	2.174	36.02	1.38	76.307	49.708	65.141	16.145	13
39.9*	3.071*	36.03*	2.62*	122.533*	94.399*	77.039*	12.993*	14
44.8	2.738	36.04	2.3	122.662	82.892	67.577	16.362	15
46.4	2.138	36.04	1.81	99.203	65.232	65.756	21.703	16
47.6	2.197	36.02	1.91	104.577	68.798	65.787	21.666	17
48.3	2.232	36.06	1.96	107.806	70.678	65.560	21.640	18
58.2	0.11	36.02	0.09	6.402	3.242	50.637	529.091	19
59.8	0.112	36.02	0.09	6.698	3.242	48.402	533.929	20
* GW-Instek Current Limited								

This test does not include input voltages above 60V because beyond that input level the converter cannot drive sufficient load current (greater than 1A). Asterisks indicate input voltages for which the GW Instek power supply reached or exceeded its current limit, therefore maximum load current and efficiency results for such input voltages are not completely accurate. Overall, as long as the SEPIC maintains a constant 36V output, it maintains a 10 Ω input resistance even at the elliptical trainer's lowest output power level. However, this only holds true up to around 40V input. Beyond that, the SEPIC loses its input resistance maintenance high load current maintenance abilities. Overall converter efficiency at the tested input voltages is also below specification. Running the SEPIC in

CCM explains its low efficiency at lower input voltage and hence while driving low-current loads, as mentioned in section 4.2. The converter shows decent efficiency levels during the typical workout's 15-30V input range; however, such levels are still below what this project desires. However, these results show that the SEPIC can handle 120W input as well as 60V input from the elliptical trainer. 60V was the maximum voltage that the EHFEM group in [1] obtained from a 160 strides per minute exercise test, and 160 strides per minute falls within the typical pace range at which Cal Poly Recreational Center users exercise. The next section investigates possible faults for causing the SEPIC's poor load regulation at input voltages above 52V.

10.3.3 First Troubleshooting Phase

This project now examines what SEPIC components could cause these faults – beginning with the SEPIC's switching inductors. Figures 10.7 to 10.10 show primary inductor L_1 's current waveforms under various input voltages and load levels. The figures' respective captions indicate SEPIC input voltage and load level. Secondary inductor L_2 's current waveforms follow a similar shape (with different magnitudes), therefore this project does not include L_2 's current waveforms in this report.

Figure 10.7: L_1 Current at 5V Input, 0.1A Load

Figure 10.8: L_1 Current at 18V Input, 1A Load

Figure 10.9: L_1 Current at 36V Input, 1.5A Load

Figure 10.10: L_1 Current at 50V Input, 2.3A Load

Figure 10.7 shows the SEPIC not regulating a 36V output (5V input with a 0.1A load) – the inductor waveform is distorted. Figures 10.8 to 10.10 are more realistic waveforms, as the SEPIC regulates a 36V output with the conditions in the mentioned figures.

Furthermore, figures 10.8 to 10.10 show “dead time” between inductor charging cycles. These waveform shapes indicate DCM operation because of the dead time, even though the current levels do not drop to 0A for any significant time period relative to the SEPIC’s 100 kHz switching frequency. Ideally, the inductor waveforms should not have any dead time in between charging periods. However, because Q_MAIN’s switching trajectory affects inductor operation, thus Q_MAIN may not be switching properly. This report’s next paragraph addresses and investigates this problem in greater detail.

However, these L_1 current waveforms in figures 10.7 to 10.10 also show that L_1 never saturates (and hence L_2 never saturates, either) for the 5-50V input and the 0.2-2.3A load current range, therefore the SEPIC’s L_1 and L_2 saturation current rating suffices.

This project’s next investigation interest area is Q_MAIN’s switching trajectory. Figures 10.11 to 10.13 show Q_MAIN’s gate voltage waveforms (with respect to signal ground) under various input voltages and load levels.

Figure 10.11: Q_MAIN Switching Trajectory at 5V Input and 0.1A Load

Figure 10.12: Q_MAIN Switching Trajectory at 36V Input and 1.46A Load

Figure 10.13: Q_MAIN Switching Trajectory at 50V Input and 2.3A Load

Figures 10.10 to 10.13 show that Q_MAIN's gate switching waveforms are nowhere close to ideal. Although Q_MAIN switches at 110 kHz, the LTC1871 controller has a problem maintaining its duty cycle. Figure 10.13 shows that at 50V input and 2.3A load, Q_MAIN's gate voltage reaches metastable states, resulting in unstable SEPIC operation. Furthermore, figures 10.10 to 10.13 show high frequency harmonics (greater than 100 kHz) in Q_MAIN's switching trajectories, which may contribute to the SEPIC's abnormal operation at higher input voltages and higher current loads. Ideally Q_MAIN's gate switching waveform should be close to what figure 5.10 shows. This project now revisits the finalized SEPIC converter's (shown in figure 7.9) simulations because unlike the simulation this project performed on the initial basic SEPIC design in chapter 5, the simulations on the finalized SEPIC design in chapter 7 neglected inspecting inductor current waveforms as well as gate switching waveforms with all parasitic elements added to the SEPIC's passive components. Figure 10.14 shows all pertinent waveforms in the finalized SEPIC from figure 7.9, with the SEPIC running under full load and steady-state operation conditions. In figure 10.14, V(out) denotes the SEPIC's output voltage, V(n013) denotes Q_MAIN's gate voltage, I(L1) denotes L₁'s current, I(D_main1) denotes the SEPIC's main switching diode D_MAIN1's current, V(sw) denotes Q_MAIN's drain voltage, V(fb) denotes the feedback node's voltage, and V(sense) denotes the sense node and Q_MAIN's source voltage, which is proportional to Q_MAIN's drain current.

Figure 10.14: Finalized SEPIC Output Voltage, Q_{MAIN} Switching Trajectory, L_1 Current and Feedback and Sense Node Voltages from LTSpice Simulation

As figure 10.14 shows, the same problem with Q_{MAIN} 's switching trajectory shows up in simulation results. Other crucial switching waveforms also do not match their shapes from figure 5.10. Additionally, the SEPIC's feedback voltage also oscillates, indicating that its node filter capacitor C_{FB} may not be large enough. This project then determined that Q_{MAIN} 's abnormal gate switching trajectory in turn caused abnormal L_1 current (from figures 10.7 to 10.10 and 10.14), D_{MAIN1} current and Q_{MAIN} drain current waveform shapes. Therefore, this project now attempts eliminating the additional high frequency harmonics from Q_{MAIN} 's gate switching trajectory.

One possible solution involves putting additional capacitance between Q_{MAIN} 's gate and ground, along with additional capacitance between the SEPIC's feedback node

and ground for stabilizing its voltage. Putting additional capacitance between Q_MAIN's gate and ground creates a low pass filter and can help shunt higher frequency harmonics to ground. Similarly, RC low pass filters dampen any oscillation that occurs on the LTC1871's GATE, FB and SENSE pins; however, adding a series resistor to those pins requires physically severing PCB copper traces already connected to them. Therefore, this project first implements purely capacitive low pass filters on the LTC1871's GATE, FB and SENSE pins and then only replaces them with RC low pass filters if necessary. An RC low pass filter, however, eliminates the leading harmonic spike on the current sense resistor voltage that results from the LTC1871 controller sensing Q_MAIN's drain current - figure 10.14 shows this phenomenon [28]. This project begins its signal harmonic suppression task by first placing a $0.01\mu\text{F}$ capacitor across the LTC1871 controller's FB pin.

Figure 10.15 the result from inserting a $0.01\mu\text{F}$ capacitor across the LTC1871's FB pin, using the same waveforms as from figure 10.14 above.

Figure 10.15: Finalized SEPIC with $0.01\mu\text{F}$ Feedback Capacitor Output Voltage, Q_{MAIN} Switching Trajectory, L_1 Current and Feedback and Sense Node Voltages from LTSpice Simulation

Although adding capacitance across the LTC1871's FB pin cleans the SEPIC's feedback voltage signal, its other waveforms are still nowhere close to their ideal shapes.

Therefore, increasing capacitance across the LTC1871's FB pin alone is not effective for obtaining proper switching waveform shapes and hence proper SEPIC operation. This project next adds a small capacitor across the LTC1871's GATE pin for smoothing out Q_{MAIN} 's gate switching waveform. For Q_{MAIN} 's gate, adding too little capacitance is ineffective for harmonic suppression, while adding too much results in unstable and inefficient converter operation. Large gate capacitances across Q_{MAIN} result in it expending more energy charging and discharging the gate, resulting in higher switching power dissipation losses and hence less efficient overall converter operation. Therefore

this project selects $0.01\mu\text{F}$ as a conservative Q_MAIN gate capacitor filter value, as it is neither too small nor too large.

Figure 10.16 shows the result from inserting a $0.01\mu\text{F}$ capacitor across the LTC1871's GATE pin (in addition to the $0.01\mu\text{F}$ capacitor across the LTC1871's FB pin), using the same waveforms as from figure 10.14 and 10.15.

Figure 10.16: Finalized SEPIC with $0.01\mu\text{F}$ Feedback Capacitor and $0.01\mu\text{F}$ Q_MAIN Gate Capacitor Output Voltage, Q_MAIN Switching Trajectory, L_1 Current and Feedback and Sense Node Voltages from LTSpice Simulation

Adding a $0.01\mu\text{F}$ capacitor across the LTC1871's GATE pin helps return Q_MAIN's gate switching waveform as well as its drain voltage waveform to their proper shapes. Likewise, the same occurs with D_MAIN1's and L_1 's current waveforms. Overall Q_MAIN power dissipation also decreases to 1.8W (from 2.2W in section 7.4) after

making this change. However, the LTC1871's SENSE pin voltage waveform still contains a leading harmonic spike. Regardless, this project performs another hardware test checking if there is any overall converter improvement with the FB and GATE pin changes.

Tables E-19 and E-20 show the results from these modifications, using the same test instrument setup as the results from tables E-17 and E-18 above. This project uses decade capacitance boxes for the additional capacitors across the LTC1871's FB and GATE pins, for prototyping purposes. SEPIC revisions in later sections of this report use discrete ceramic capacitors. Again, likewise with the results from tables E-17 and E-18, Table E-19 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-20 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-19: Modified Feedback and Gate Nodes SEPIC Load Tests with Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)
5	2.463	36.01	0.17	12.315	6.122	49.709	2.030
10	3.067*	36.02	0.24*	30.67*	8.645*	28.187*	3.261*
12	3.067*	36.02	0.62*	36.804*	22.332*	60.679*	3.913*
15	2.946*	36.02	0.72*	44.19*	25.934*	58.688*	5.092*
18	3.068*	36.02	0.9*	55.224*	32.418*	58.703*	5.867*
20	2.037	36.02	0.65	40.74	23.413	57.469	9.818
25	2.423	36.02	1	60.575	36.020	59.463	10.318
27	2.594	36.02	1.18	70.038	42.504	60.686	10.409
30	2.851	36.02	1.48	85.53	53.310	62.329	10.523
36	3.067*	36.02	1.79*	110.412*	64.476*	58.396*	11.738*
40	3.067*	36.02	1.79*	122.68*	64.476*	52.556*	13.042*
45	3.067*	36.02	1.79*	138.015*	64.476*	46.717*	14.672*
50	3.067*	36.02	1.79*	153.35*	64.476*	42.045*	16.303*
52	3.067*	36.02	1.79*	159.484*	64.476*	40.428*	16.955*
60	3.067*	36.02	1.79*	184.02*	64.476*	35.037*	19.563*
* GW-Instek Current Limited							

Table E-20: Modified Feedback and Gate Nodes SEPIC Load Tests with 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, TPS-4000 and GW Instek Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Training Resistance Level
8.08*	3.068*	36	0.251*	24.789*	9.036*	36.451*	2.634*	2
11.54*	3.068*	36	0.251*	35.405*	9.036*	25.522*	3.761*	3
13.49*	3.068*	36	0.251*	41.387*	9.036*	21.833*	4.397*	4
15.88*	3.067*	36	0.804*	48.704*	28.944*	59.428*	5.178*	5
18.27*	3.067*	36	0.804*	56.034*	28.944*	51.654*	5.957*	6
21	2.145	36	0.748	45.045	26.928	59.780	9.790	7
23.3	2.289	36	0.885	53.334	31.86	59.737	10.179	8
26.2	2.533	36	1.128	66.365	40.608	61.189	10.343	9
29.1	2.774	36	1.397	80.723	50.292	62.302	10.490	10
31.8	3.032	36	1.691	96.418	60.876	63.138	10.488	11
34.3*	3.069*	36	1.735*	105.267*	62.46*	59.335*	11.176*	12
35.1*	3.069*	36	1.735*	107.722*	62.46*	57.983*	11.437*	13
39.9*	3.069*	36	1.735*	122.453*	62.46*	51.007*	13.001*	14
44.8*	3.069*	36	1.735*	137.491*	62.46*	45.428*	14.598*	15
46.4*	3.069*	36	1.735*	142.402*	62.46*	43.862*	15.119*	16
47.6*	3.069*	36	1.735*	146.084*	62.46*	42.756*	15.510*	17
48.3*	3.069*	36	1.735*	148.233*	62.46*	42.136*	15.738*	18
58.2*	3.069*	36	1.735*	178.616*	62.46*	34.969*	18.964*	19
59.8*	3.069*	36	1.735*	183.526*	62.46*	34.033*	19.485*	20
* GW-Instek Current Limited								

Likewise with the results from tables E-17 and E-18, this test does not include input voltages above 60V because beyond that input level the converter begins shutting down operation, which the LT4356-1 input protection circuit causes. Asterisks indicate input voltages for which the GW Instek power supply reached or exceeded its current limit, therefore maximum load current and efficiency results for such input voltages are not completely accurate. A problem with load current capacity exists between 15V to 20V input – the SEPIC’s maximum load current driving capability drops after the 18V point. The next troubleshooting round in section 10.3.4 addresses this issue. However, this issue is not critical because at 18V and lower input voltages, the SEPIC does not maintain a

10 Ω input impedance while driving its maximum load current while simultaneously maintaining 36V output. Adding a 0.01 μ F capacitor to the LTC1871's FB pin and a 0.01 μ F capacitor to its GATE pin shows promising results and improves overall converter efficiency, although the GW Instek power supply's 3A current limit renders most of these results inaccurate. Therefore, this project requires a power supply capable of supplying this converter's specified 6.5A current limit. The next section describes a new instrument setup for removing the input supply current limit from results.

10.3.3 Constant Voltage Load Tests Using Non-Current Limited Input Supply

The instrument setup for this section's testing round uses a BK Precision XLN3640 power supply. This power supply can supply up to 40A but it also has a 36V output limit. Therefore this testing round only tests the SEPIC at input voltages up to 36V. This voltage limit suffices for testing the SEPIC when operating in boost mode, as section 3.3 describes the typical elliptical trainer exercise output voltage range as 15-30V. Subsequent sections in this report use a power supply capable of this SEPIC's required input voltage range as well as current limit. Figure 10.17 shows the instrument setup block diagram for this section's testing round.

Figure 10.17: Finalized SEPIC with $0.01\mu\text{F}$ Feedback Capacitor and $0.01\mu\text{F}$ Q_{MAIN} Gate Capacitor Test Setup Block Diagram

This power supply was in Cal Poly's electrochemical engineering laboratory (room 20-130), therefore this project performs this testing round using equipment in that laboratory. Furthermore, this testing round removes the inductor current wire loops from the PCB and replaces those wires with solder bridges because no current probe amplifiers exist in the 20-130 laboratory. This project requires current probe amplifiers in addition to the mentioned wire loops for measuring inductor current. This testing round uses the same Tektronix oscilloscope as the previous one, but a different electronic load. The 12V power supply used to simulate the elliptical trainer's onboard battery is also different – this time this project uses an Agilent E3630A power supply for simulating the elliptical trainer's battery. This testing round uses a BK Precision 8518 1.2kW electronic load, which more than suffices for this SEPIC's 288W maximum output load driving requirement. Tables E-21 and E-22 show the results from this instrument setup. Likewise

with the tests from the previous section, this project uses decade capacitance boxes for the additional capacitors across the LTC1871's FB and GATE pins, for prototyping purposes. Also, likewise with the results from the previous section, Table E-21 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-22 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-21: Modified Feedback and Gate Nodes SEPIC Load Tests with Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)
5	3.037	36	0.25	15.185	9	59.269	1.646
10	3.466	36	0.626	34.66	22.536	65.020	2.885
12	3.407	36	0.687	40.884	24.732	60.493	3.522
15	3.029	36	0.772	45.435	27.792	61.169	4.952
18	3.59	36	1.149	64.62	41.364	64.011	5.014
20	2.099	36	0.697	41.98	25.092	59.771	9.528
25	2.445	36	1.043	61.125	37.548	61.428	10.225
27	2.585	36	1.23	69.795	44.28	63.443	10.445
30	2.802	36	1.553	84.06	55.908	66.510	10.707
36	3.513	36	2.473	126.468	89.028	70.396	10.248

Table E-22: Modified Feedback and Gate Nodes SEPIC Load Tests with 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)	Training Resistance Level
8.08	3.445	36	0.46	27.835	16.56	59.492	2.345	2
11.54	3.067	36	0.595	35.393	21.42	60.520	3.763	3
13.49	2.736	36	0.62	36.908	22.32	60.474	4.931	4
15.88	3.183	36	0.866	50.546	31.176	61.678	4.989	5
18.27	2.683	36	0.825	49.018	29.7	60.589	6.810	6
21	2.163	36	0.758	45.423	27.288	60.075	9.709	7
23.3	2.316	36	0.903	53.962	32.508	60.241	10.060	8
26.2	2.508	36	1.158	65.709	41.688	63.443	10.447	9
29.1	2.736	36	1.445	79.617	52.02	65.337	10.636	10
31.8	2.985	36	1.76	94.923	63.36	66.749	10.653	11
34.3	3.291	36	2.152	112.881	77.472	68.631	10.422	12
35.1	3.402	36	2.297	119.41	82.692	69.250	10.317	13

From these non-current limited results, adding a $0.01\mu\text{F}$ capacitor across the LTC1871's FB pin and across its GATE pin improves the SEPIC's overall efficiency. Furthermore, these modifications also improve the SEPIC's output load driving ability, as the SEPIC now drives more current at its load while maintaining a 36V output. Overall converter efficiency increases with increasing input voltage and output load current, however, the load current capacity problem at input voltages between 15V and 20V still exists.

Furthermore, the SEPIC does not maintain a 10Ω input impedance until the 20V input point. This project then hypothesizes that a slightly larger capacitance across the LTC1871's GATE pin may correct the load current capacity problem. This project then increases the LTC1871's new GATE pin capacitance to $0.02\mu\text{F}$.

Figure 10.18 shows the result from inserting a $0.02\mu\text{F}$ capacitor across the LTC1871's GATE pin (in addition to the $0.01\mu\text{F}$ capacitor across the LTC1871's FB pin), using the same waveforms as from figure 10.14 and 10.15.

Figure 10.18: Finalized SEPIC with 0.01 μ F Feedback Capacitor and 0.02 μ F Q_{MAIN} Gate Capacitor Output Voltage, Q_{MAIN} Switching Trajectory, L_1 Current and Feedback and Sense Node Voltages from LTSpice Simulation

The results remain unchanged from figure 10.16, when this project used a 0.01 μ F capacitor across the LTC1871's GATE pin. Power dissipation across Q_{MAIN} also remains the same as before. However, simulation results do not always necessary translate to physical hardware test results, therefore this project now tests this capacitance change using the same instrument setup described earlier in this section. Tables E-23 and E-24 show these results. Likewise with the tests from earlier in this section, this project uses decade capacitance boxes for the additional capacitors across the LTC1871's FB and GATE pins, for prototyping purposes. Also, likewise with those tests, Table E-23 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-24 shows the same test results, except using input

voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-23: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
5	0.486	36	0.03	2.43	1.08	44.444	10.288
10	0.948	36	0.136	9.48	4.896	51.646	10.549
12	1.135	36	0.211	13.62	7.596	55.771	10.573
15	1.406	36	0.351	21.09	12.636	59.915	10.669
18	1.684	36	0.531	30.312	19.116	63.064	10.689
20	1.881	36	0.677	37.62	24.372	64.785	10.633
25	2.347	36	1.102	58.675	39.672	67.613	10.652
27	2.51	36	1.287	67.77	46.332	68.367	10.757
30	2.755	36	1.591	82.65	57.276	69.299	10.889
36	3.479	36	2.484	125.244	89.424	71.400	10.348

Table E-24: SEPIC Load Test with 0.02 μ F Gate Capacitor Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Training Resistance Level
8.08	0.764	36	0.08	6.173	2.88	46.654	10.576	2
11.54	1.088	36	0.191	12.555	6.876	54.765	10.607	3
13.49	1.263	36	0.274	17.037	9.864	57.895	10.681	4
15.88	1.478	36	0.396	23.47	14.256	60.740	10.744	5
18.27	1.703	36	0.547	31.113	19.692	63.290	10.728	6
21	1.972	36	0.752	41.412	27.072	65.372	10.649	7
23.3	2.189	36	0.944	51.003	33.984	66.630	10.644	8
26.2	2.436	36	1.206	63.823	43.416	68.025	10.755	9
29.1	2.669	36	1.488	77.667	53.568	68.971	10.903	10
31.8	2.92	36	1.801	92.856	64.836	69.824	10.890	11
34.3	3.243	36	2.188	111.234	78.768	70.812	10.577	12
35.1	3.362	36	2.33	118.006	83.88	71.081	10.440	13

Increasing the LTC1871's GATE pin capacitance to 0.02 μ F improves overall converter efficiency. It also decreases the SEPIC's maximum load current driving capability at input voltages below 25V, but allows the SEPIC to maintain a 10 Ω input impedance at all tested input voltage levels. These results would lead one believing that increasing the

LTC1871's GATE pin capacitance further would further increase SEPIC performance, but that is not true. Q_MAIN gate capacitances above $0.02\mu\text{F}$ result in deteriorated converter performance, and its overall performance at such higher gate capacitances is no different from the original finalized SEPIC design with no additional gate capacitance. Therefore this project does not include results for Q_MAIN gate capacitances above $0.02\mu\text{F}$. This project hypothesizes that Q_MAIN's switching timing is crucial for proper operation with the SEPIC's snubber, as the snubber component selection process from section 6.3 and figure 6.14 relies on Q_MAIN's gate charging and discharging time periods. Selecting the proper Q_MAIN gate capacitor alters Q_MAIN's charging and discharging time periods such that Q_MAIN effectively exchanges energy with the SEPIC's snubber, therefore Q_MAIN's gate capacitance cannot be too large or small. Furthermore, as mentioned in section 10.3.3, too much capacitance on Q_MAIN's gate increases the energy required for charging and discharging it, therefore this project does not desire larger capacitance across Q_MAIN's gate.

Now this project investigates Q_MAIN's switching trajectory waveforms with the additional $0.02\mu\text{F}$ capacitor connected across its gate. Figures 10.19 and 10.20 show Q_MAIN's new gate voltage waveforms (with respect to signal ground) under 5V and 36V input, while the SEPIC drives its maximum load current (while maintaining 36V output) under those input conditions.

Figure 10.19: Q_MAIN Switching Trajectory with $0.02\mu F$ Gate Capacitor at 5V Input and 0.03A Load

Figure 10.20: Q_MAIN Switching Trajectory with $0.02\mu F$ at 36V Input and 2.484A Load

The decade capacitor boxes used for adding capacitance to the LTC1871's FB and GATE pins introduce oscillations into Q_MAIN 's switching trajectory waveform. These oscillations originate from parasitic resistance and inductance in both the decade capacitor boxes as well as the long wire leads that this project uses for connecting them. Furthermore, there are still higher frequency harmonics present in the waveforms. Using discrete ceramic capacitors with low ESR ratings rather than decade capacitor boxes may

eliminate the parasitic resistance and inductance problem. This project's next test set uses discrete ceramic capacitors for the LTC1871's FB and GATE pins, for checking any differences between the test results from tables E-23 and E-24 as well as the results that using discrete capacitors show.

Tables E-25 and E-26 show the results from when this project uses discrete capacitors instead for the additional capacitors across the LTC1871's FB and GATE pins. Likewise with the tests from earlier in this section, Table E-25 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-26 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-25: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
5	1.037	36	0.07	5.185	2.52	48.602	4.822
10	2.778	36	0.465	27.78	16.74	60.259	3.600
12	2.868	36	0.585	34.416	21.06	61.192	4.184
15	2.646	36	0.692	39.69	24.912	62.766	5.669
18	1.605	36	0.507	28.89	18.252	63.178	11.215
20	1.785	36	0.634	35.7	22.824	63.933	11.204
25	2.237	36	1.012	55.925	36.432	65.144	11.176
27	2.417	36	1.184	65.259	42.624	65.315	11.171
30	2.686	36	1.471	80.58	52.956	65.719	11.169
36	3.239	36	2.164	116.604	77.904	66.811	11.115

Table E-26: SEPIC Load Test with 0.02 μ F Gate Capacitor Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Training Resistance Level
8.08	2.75	36	0.379	22.22	13.644	61.404	2.938	2
11.54	2.835	36	0.575	32.715	20.7	63.272	4.071	3
13.49	2.379	36	0.568	32.092	20.448	63.715	5.670	4
15.88	2.901	36	0.829	46.067	29.844	64.783	5.474	5
18.27	1.622	36	0.523	29.633	18.828	63.535	11.264	6
21	1.867	36	0.703	39.207	25.308	64.550	11.248	7
23.3	2.075	36	0.875	48.347	31.5	65.153	11.229	8
26.2	2.338	36	1.115	61.255	40.14	65.529	11.206	9
29.1	2.599	36	1.382	75.630	49.752	65.783	11.197	10
31.8	2.845	36	1.66	90.471	59.76	66.054	11.178	11
34.3	3.078	36	1.946	105.575	70.056	66.356	11.144	12
35.1	3.153	36	2.046	110.67	73.656	66.554	11.132	13

Using discrete ceramic capacitors rather than decade capacitor boxes for the LTC1871 FB and GATE pin capacitors decreases overall converter efficiency, decreased load driving ability as well as removes the SEPIC's 10 Ω input resistance regulation ability at all tested input voltages. These results, however, are better than the results from tables E-17 and E-18, when the SEPIC did not have the LTC1871 FB and GATE pin capacitors. This project later discovered that the discrete capacitors are not the reason for decreased

converter functionality. Before replacing the decade capacitor boxes across the LTC1871's FB and GATE pins with discrete capacitors, this project swapped out the IPP110N20N3 MOSFET used for Q_MAIN with the IXFK230N20T MOSFET used for Q_MAIN_ALT, for determining whether Q_MAIN_ALT would yield higher overall converter efficiency and functionality. That side-experiment proved disastrous, as the IXFK230N20T yielded poorer overall converter efficiency and failed to function correctly with the SEPIC's snubber. At 36V input its incorrect functionality with the snubber shorted out the BK XLN3640A power supply and the XLN3640A then supplied 30A (average) to the converter, destroying at least one unknown component on the SEPIC's switching node. During testing this project's author enclosed the SEPIC inside a cinderblock container for preventing any physical injury in case any components (particularly electrolytic capacitors) exploded. The author witnessed a spark and a loud pop noise somewhere near the SEPIC's switching node, but could not determine which component it was from because of difficulty looking inside the enclosure. The author does not know if that short circuit condition destroyed any other components. This project replaced the IXFK230N20T with the previously-used IPP110N20N3 MOSFET shortly afterward. This project discarded all test results using the IXFK230N20T MOSFET because overall converter efficiency was worse than with the IPP110N20N3 MOSFET. Therefore, this project's author speculates that such component destruction caused reduced overall SEPIC efficiency, input resistance maintenance and load current driving abilities. Later, this project author's determined that one of the snubber diodes, DSNUB1, was destroyed by the excessive DC current flow, after measuring its forward

and reverse voltages as 0V. The author replaced DSNUB1 with a working APT30S20BG Schottky diode afterward. After replacing DSNUB1, this project moves on and focuses on other converter efficiency and functionality improvement methods. Although this project made a mistake in replacing the SEPIC's main switching transistor for side-tests before testing with discrete capacitors on the LTC1871's FB and GATE pins, the side-test did verify that the IXFK230N20T does not function with the SEPIC's snubber, because the SEPIC's snubber relies heavily on proper main switching transistor timing, based on the snubber design equations from section 6.3 as well as figure 6.14.

The next converter efficiency and functionality improvement method that this project focuses on is decreasing the SEPIC's output voltage ripple. Figure 10.21 shows the SEPIC's output voltage ripple at 36V input, while driving its maximum load current that maintains a 36V output (2.164A). This SEPIC includes the 0.01 μ F LTC1871 FB pin and 0.02 μ F GATE pin capacitors.

Figure 10.21: SEPIC Output Voltage Ripple at 36V Input and 2.164A Load

This project measured the SEPIC's peak to peak output voltage ripple at 36V input and 2.164A load as 4.40V, which corresponds to a 12.22% ripple with respect to the SEPIC's 36V average output. This value is higher than the output ripple specification outlined in section 1.2. Therefore this project adds another large capacitor to the SEPIC's output node. The SEPIC's PCB layout in figure 9.6 has a placeholder footprint on the lower right corner for another radial electrolytic capacitor. Therefore, this project adds another 470µF capacitor identical to C_{OUT} for testing if the SEPIC's output ripple voltage decreases, as well as if the SEPIC's overall efficiency and load current driving ability improves. This project adds another 470µF Epcos B43504A9477M electrolytic capacitor to the SEPIC's output. This capacitor has 190mΩ ESR, therefore its overall ripple

reducing performance may not be as ideal as a single 940 μ F capacitor with a lower ESR value. Tables E-27 and E-28 show the results from when this project uses an additional 470 μ F electrolytic capacitor at the SEPIC's output, along with the current discrete ceramic capacitors for the LTC1871's FB and GATE pins. Likewise with the tests from earlier in this section, Table E-27 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-28 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-27: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Additional 470 μ F Output Capacitor, Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)
5	1.216	36	0.085	6.08	3.06	50.329	4.112
10	0.921	36	0.131	9.21	4.716	51.205	10.858
12	1.101	36	0.194	13.212	6.984	52.861	10.899
15	1.366	36	0.319	20.49	11.484	56.047	10.981
18	1.63	36	0.478	29.34	17.208	58.650	11.043
20	1.822	36	0.608	36.44	21.888	60.066	10.977
25	2.284	36	0.999	57.1	35.964	62.984	10.946
27	2.453	36	1.176	66.231	42.336	63.922	11.007
30	2.683	36	1.46	80.49	52.56	65.300	11.182
36	3.302	36	2.231	118.872	80.316	67.565	10.902

Table E-28: SEPIC Load Test with 0.02 μ F Gate Capacitor and Additional 470 μ F Output Capacitor Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load, Agilent E3630A and BK XLN3640 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Training Resistance Level
8.08	0.741	36	0.083	5.987	2.988	49.906	10.904	2
11.54	1.058	36	0.179	12.209	6.444	52.779	10.907	3
13.49	1.228	36	0.251	16.565	9.036	54.546	10.985	4
15.88	1.421	36	0.352	22.565	12.672	56.157	11.175	5
18.27	1.63	36	0.479	29.78	17.244	57.904	11.209	6
21	1.846	36	0.626	38.766	22.536	58.133	11.376	7
23.3	2.118	36	0.845	49.349	30.42	61.642	11.001	8
26.2	2.372	36	1.093	62.146	39.348	63.315	11.046	9
29.1	2.595	36	1.357	75.514	48.852	64.692	11.214	10
31.8	2.81	36	1.628	89.358	58.608	65.588	11.317	11
34.3	3.049	36	1.93	104.58	69.48	66.437	11.250	12
35.1	3.143	36	2.048	110.319	73.728	66.831	11.168	13

Adding a 470 μ F capacitor to the SEPIC's output slightly increases converter efficiency as well as load current driving ability at higher input voltages. Furthermore, at all tested SEPIC input voltages except for 5V the converter maintains close to a 10 Ω input resistance while driving the maximum load current that it tolerates while maintaining a 36V output. The additional capacitor's output voltage ripple reduction may have been responsible for that. Figure 10.22 shows the SEPIC's output voltage ripple at 36V input, while driving its maximum load current that maintains a 36V output (2.231A), with the additional 470 μ F output capacitor.

Figure 10.22: SEPIC Output Voltage Ripple at 36V Input and 2.231A Load With Additional 470 μ F Output Capacitor

Adding a 470 μ F capacitor to the SEPIC's output reduces its output ripple voltage at 36V input to 2V, which corresponds to a 5.56% ripple relative to the 36V output. This ripple value suffices from the specifications outlined in section 1.2. Therefore, this project performs all future hardware tests with the additional 470 μ F capacitor on the SEPIC's output. Figures 10.23 and 10.24 show Q_MAIN's switching trajectory at 5V input and 36V input while driving their respective maximum load currents that maintain 36V SEPIC output, when this project uses discrete capacitors on the LTC1871's FB and GATE pins.

Figure 10.23: Q_MAIN Switching Trajectory with $0.02\mu F$ Discrete Gate Capacitor and Additional $470\mu F$ Capacitor at 5V Input and 0.085A Load

Figure 10.24: Q_MAIN Switching Trajectory with $0.02\mu F$ Discrete Gate Capacitor and Additional $470\mu F$ Capacitor at 36V Input and 2.231A Load

Compared to figures 10.19 and 10.20 when this project used decade capacitor boxes for the additional LTC1871 FB and GATE pins, using discrete capacitors on those same pins yields less oscillation, however, high frequency harmonics are still present at 36V input (from figure 10.24). In the next section, this project tests whether those high frequency harmonics still greatly affect converter operation. After improving the SEPIC's output voltage ripple performance, this project performs constant voltage load tests at input voltages beyond 36V.

10.3.4 Constant Voltage Load Tests Using Non-Voltage and Current Limited Input Supply

The instrument setup for this section's testing round uses a BK Precision 9153 power supply. This power supply can supply up 60V and 9A. Because of this voltage limitation, this project does not perform any tests beyond 60V input voltage. This voltage limit suffices for testing the SEPIC, as it is very close to the 65V absolute maximum voltage specification outlined in section 1.2. Furthermore, no-load tests from section 10.2 shows that the SEPIC still outputs its nominal output voltage at 65V input. Figure 10.25 shows the instrument setup block diagram for this section's testing round.

Figure 10.25: Finalized SEPIC with 0.01 μ F Feedback Capacitor, Additional 470 μ F Output Capacitor and 0.02 μ F Q_{MAIN} Gate Capacitor Test Setup Block Diagram (Non-Voltage and Current Limited)

This BK Precision power supply was in Cal Poly's renewable energy laboratory (room 20-150), therefore this project performs this testing round using equipment in that laboratory. Likewise with the previous testing round, this testing round keeps the inductor

current wire loops removed from the PCB and replaces those wires with solder bridges because no current probe amplifiers exist in the 20-150 laboratory. Unlike the previous testing round, this testing round uses an Agilent MSO-X 2012A oscilloscope. This project also measures the 12V source's current during this testing round. Measuring the 12V source current allows complete converter efficiency calculations. This testing round also uses a BK Precision 8510 600W electronic load, which more than suffices for this SEPIC's 288W maximum output load driving requirement. Tables E-29 and E-30 show the results from this instrument setup. Likewise with the previous section, Tables E-29 and E-30 show the results from when this project uses an additional 470 μ F electrolytic capacitor at the SEPIC's output, along with the current discrete ceramic capacitors for the LTC1871's FB and GATE pins. Table E-29 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-30 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-29: SEPIC Load Tests with 0.02 μ F Gate Capacitor, Additional 470 μ F Output Capacitor, Constant Voltage Electronic Load and BK 9153 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	V _{BAT} (V)	I _{BAT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)
5	1.237	36	0.083	12	0.024	6.473	2.988	46.161	4.042
10	0.944	36	0.131	12	0.024	9.728	4.716	48.479	10.593
12	1.128	36	0.196	12	0.023	13.812	7.056	51.086	10.638
15	1.396	36	0.324	12	0.023	21.216	11.664	54.977	10.745
18	1.665	36	0.486	12	0.023	30.246	17.496	57.846	10.811
20	1.86	36	0.619	12	0.023	37.476	22.284	59.462	10.753
25	2.34	36	1.022	12	0.023	58.776	36.792	62.597	10.684
27	2.518	36	1.207	12	0.023	68.262	43.452	63.655	10.723
30	2.754	36	1.5	12	0.023	82.896	54	65.142	10.893
36	3.37	36	2.28	12	0.023	121.596	82.08	67.502	10.682
40	3.838	36	2.936	12	0.023	153.796	105.696	68.725	10.422
45	4.101	36	3.566	12	0.023	184.821	128.376	69.460	10.973
50	3.039	36	2.832	12	0.023	152.226	101.952	66.974	16.453
52	2.3	36	2.19	12	0.023	119.876	78.84	65.768	22.609
60	0.113	36	0.089	12	0.023	7.056	3.204	45.408	530.973

Table E-30: SEPIC Load Test with 0.02 μ F Gate Capacitor and Additional 470 μ F Output Capacitor, Using 160 Strides/Min. Theoretical Input Voltages from [1], Constant Voltage Electronic Load and BK 9153 Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	V_{BAT} (V)	I_{BAT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)	Training Resistance Level
8.08	0.762	36	0.082	12	0.024	6.445	2.952	45.803	10.604	2
11.54	1.086	36	0.18	12	0.023	12.808	6.48	50.592	10.626	3
13.49	1.264	36	0.257	12	0.023	17.327	9.252	53.395	10.672	4
15.88	1.471	36	0.367	12	0.023	23.635	13.212	55.899	10.795	5
18.27	1.69	36	0.503	12	0.023	31.152	18.108	58.127	10.811	6
21	1.96	36	0.693	12	0.023	41.436	24.948	60.209	10.714	7
23.3	2.185	36	0.877	12	0.023	51.187	31.572	61.680	10.664	8
26.2	2.444	36	1.132	12	0.023	64.309	40.752	63.369	10.720	9
29.1	2.674	36	1.408	12	0.023	78.089	50.688	64.910	10.883	10
31.8	2.894	36	1.69	12	0.023	92.305	60.84	65.912	10.988	11
34.3	3.14	36	2.006	12	0.023	107.978	72.216	66.880	10.924	12
35.1	3.238	36	2.126	12	0.023	113.930	76.536	67.178	10.840	13
39.9	3.82	36	2.918	12	0.023	152.694	105.048	68.796	10.445	14
44.8	4.082	36	3.533	12	0.023	183.150	127.188	69.445	10.975	15
46.4	4.146	36	3.72	12	0.023	192.650	133.92	69.515	11.192	16
47.6	3.351	36	3.036	12	0.023	159.784	109.296	68.403	14.205	17
48.3	3.245	36	2.949	12	0.023	157.010	106.164	67.616	14.884	18
58.2	0.113	36	0.088	12	0.023	6.853	3.168	46.231	515.044	19
59.8	0.113	36	0.09	12	0.023	7.033	3.24	46.066	529.204	20

In tables E-29 and E-30, V_{BAT} corresponds to the 12V source's voltage and I_{BAT} corresponds to the 12V source's current. In this section and subsequent hardware testing sections without the elliptical trainer, this project factors in those two parameters into the converter's input power for the final converter efficiency calculation. From tables E-29 and E-30, the SEPIC still experiences the same problems as from section 10.3.2. The SEPIC cannot drive any sufficient load current (greater than 1A) beyond 45V input. After adding capacitance to the LTC1871's FB and GATE pins and still achieving less than desirable functionality results, this project hypothesizes that the current sensing pin on the LTC1871 controller (the SENSE pin) requires additional filtering.

10.3.5 Second Troubleshooting Phase

Figure 10.26 shows the LTC1871's SENSE pin switching waveform at 36V input while the SEPIC drives its maximum load current while maintaining 36V output (2.28A), while still using discrete capacitors for the LTC1871's FB and GATE pins.

Figure 10.26: LTC1871 Current Sense Pin Switching Waveform at 36V Input and 2.28A Load

From figure 10.26, large voltage spikes exist on the LTC1871's current sense pin, from sensing Q_MAIN's current. These voltage spikes are undesirable and this project hypothesizes that it contributes to the SEPIC's undesirable operation at higher input voltages. One possible solution for this problem involves adding an RC low pass filter at the LTC1871's SENSE pin input, as an RC low pass filter removes the higher frequency harmonic signals [28]. Figure 10.27 shows this modification to the SEPIC's schematic.

Figure 10.27: LTC1871 Current Sense Pin RC Low Pass Filter

The red arrow in figure 10.27 denotes the SENSE pin filter location. Equation (10.1) determines the RC low pass filter's attenuation (cut-off) frequency based on values selected for its resistor ($RSNS_FILT$) and capacitor (C_{SNS2}). Because the filter consists of only one resistor and one capacitor (this project combines the existing 47pF capacitor C_{SNS1} on the LTC1871's SENSE pin in parallel with the new filter capacitor, thus adding the two capacitances together), the attenuation after the cut-off frequency only occurs at 20 dB per decade. This suffices as this project does not desire complex filters for any controller pins, as complex filters increase overall converter cost, complexity, and also require more PCB trace severing.

$$f_c = \frac{1}{2\pi RSNS_FILT(C_{SNS1} + C_{SNS2})} = \frac{1}{2\pi(10\Omega)(47pF + 0.03\mu F)} = 531 \text{ kHz} \quad (10.1)$$

Therefore, the chosen component values for the RC low pass filter attenuate harmonics higher than 531 kHz at 20 dB per decade. Figure 10.28 shows the LTC1871's current sense pin's waveform from LTSpice simulation results after adding this RC low pass filter to the LTC1871's SENSE pin.

Figure 10.28: LTC1871 Current Sense Pin Waveform With Low Pass RC Filter In Steady-State Operation

Adding this low pass RC filter to the LTC1871's SENSE pin eliminates the leading harmonic spike on the sense resistor's voltage. This current sense signal waveform now appears closer to the ideal Q_MAIN drain current waveform shown in figure 5.9. Thus, the chosen resistor and capacitor values suffice for this filter. Inserting the filter resistor (RSNS_FILT) requires severing an existing PCB trace between R_{SNS} and the LTC1871's SENSE pin. This project's author then severed the trace and implemented the RC low

pass filter on the LTC1871's current sense pin. Tables E-31 and E-32 show the results from adding this RC low pass filter to the LTC1871's current sense pin, while keeping all other SEPIC components intact. Table E-31 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-32 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-31: SEPIC Load Tests with LTC1871 Current Sense Pin Filter and BK 9153 Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	V_{BAT} (V)	I_{BAT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)
5	0.5	37.34	0.013	12	0.024	2.788	0.485	17.411	10.000
10	1.099	37.34	0.148	12	0.024	11.278	5.526	49.001	9.099
12	1.196	37.34	0.202	12	0.023	14.628	7.543	51.563	10.033
15	1.477	37.32	0.333	12	0.023	22.431	12.428	55.404	10.156
18	1.76	37.31	0.498	12	0.023	31.956	18.580	58.144	10.227
20	1.979	37.3	0.638	12	0.023	39.856	23.797	59.708	10.106
25	2.496	37.26	1.048	12	0.023	62.676	39.048	62.302	10.016
27	2.721	37.24	1.248	12	0.023	73.743	46.476	63.024	9.923
30	2.958	37.2	1.528	12	0.023	89.016	56.842	63.855	10.142
36	3.646	36	2.4	12	0.023	131.532	86.400	65.687	9.874
40	3.96	36	2.92	12	0.023	158.676	105.120	66.248	10.101
45	4.24	36	3.517	12	0.023	191.076	126.612	66.263	10.613
50	4.996	36	4.73	12	0.023	250.076	170.280	68.091	10.008
52	5.112	36	5.022	12	0.023	266.1	180.792	67.941	10.172
60	5.253	36	6.093	12	0.023	315.456	219.348	69.534	11.422

Table E-32: SEPIC Load Test with LTC1871 Current Sense Pin Filter, Using 160 Strides/Min. Theoretical Input Voltages from [1] and BK 9153 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	V _{BAT} (V)	I _{BAT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Training Resistance Level
8.08	0.754	37.37	0.066	12	0.024	6.380	2.466	38.657	10.716	2
11.54	1.111	37.35	0.168	12	0.023	13.097	6.275	47.910	10.387	3
13.49	1.314	37.34	0.248	12	0.023	18.002	9.260	51.441	10.266	4
15.88	1.636	37.33	0.388	12	0.023	26.256	14.484	55.165	9.707	5
18.27	1.814	37.32	0.518	12	0.023	33.418	19.332	57.849	10.072	6
21	2.108	37.3	0.718	12	0.023	44.544	26.781	60.123	9.962	7
23.3	2.327	37.29	0.898	12	0.023	54.495	33.486	61.448	10.013	8
26.2	2.644	37.26	1.168	12	0.023	69.549	43.520	62.574	9.909	9
29.1	2.865	37.23	1.428	12	0.023	83.648	53.164	63.558	10.157	10
31.8	3.115	37.14	1.718	12	0.023	99.333	63.807	64.235	10.209	11
34.3	3.4	36.93	2.058	12	0.023	116.896	76.002	65.017	10.088	12
35.1	3.479	36	2.242	12	0.023	122.389	80.712	65.947	10.089	13
39.9	3.92	36	2.908	12	0.023	156.684	104.688	66.815	10.179	14
44.8	4.196	36	3.51	12	0.023	188.257	126.360	67.121	10.677	15
46.4	4.377	36	3.803	12	0.023	203.369	136.908	67.320	10.601	16
47.6	4.819	36	4.365	12	0.023	229.660	157.140	68.423	9.878	17
48.3	5.103	36	4.715	12	0.023	246.751	169.740	68.790	9.465	18
58.2	5.265	36	5.924	12	0.023	306.699	213.264	69.535	11.054	19
59.8	5.27	36	6.092	12	0.023	315.422	219.312	69.530	11.347	20

From tables E-31 and E-32, adding a RC low pass filter to the LTC1871's current sense pin dramatically improves the SEPIC's functionality. The SEPIC now can drive sufficient load currents at input voltages greater than 45V. At 60V input the SEPIC drives a 220W load (6.092A output current), while taking in 315W from the input power supply.

Therefore, adding filtering to the LTC1871's FB, GATE and SENSE pins is necessary for improving the SEPIC's overall load driving ability. For this set of tests, this project used constant current mode on the electronic load for input voltages less than 36V because at those input voltages, the SEPIC draws the absolute maximum current limit (6.5A) when using constant 36V output mode on the electronic load. This means that the

SEPIC can drive higher load currents at those input voltages, but this project only selects current limits that allow the SEPIC to maintain a 10Ω input resistance. Figure 10.29 shows the SEPIC's maximum load current driving ability at all tested input voltages while still maintaining a 36V output. Figure 10.30 shows the SEPIC's efficiency at all tested input voltages.

Figure 10.29: SEPIC Maximum Load Current Driving Ability vs. Input Voltage with LTC1871 Current Sense Pin Filtering

Figure 10.30: SEPIC Efficiency vs. Input Voltage with LTC1871 Current Sense Pin Filtering

The SEPIC drives a 6.092A maximum load at 60V input, and its efficiency peaks at nearly 70%. Maximum load current driving ability and efficiency increase with input voltage. Therefore, adding filtering to the LTC1871's critical control pins (the FB, GATE and SENSE pins) is necessary for proper converter operation. However, its efficiency is still below the 75% specification outlined in section 1.2. This project hypothesizes that magnetic losses (such as core losses) in the SEPIC's main switching inductors L_1 and L_2 are responsible for the low efficiency. Magnetic losses in turn cause switching and conduction losses in Q_MAIN, leading to increased Q_MAIN operating stress. The next section describes a method for improving the converter's overall efficiency.

10.3.6 SEPIC Efficiency Improvement

Magnetic losses are unavoidable for this project's SEPIC because this project selected commercially available inductors for L_1 and L_2 rather than use custom inductors. Therefore, the efficiency improvement method described in this section focuses on minimizing switching and conduction losses across Q_MAIN. The power supply's input transient voltage step to the SEPIC affects Q_MAIN's performance. Linear Technology's LT4356-1 datasheet describes a method for improving transistor switching and conduction performance when fast input voltage steps occur at the converter's input. Figure 10.31 shows this method, which involves adding damping circuitry to Q_MAIN [55].

Figure 10.31: Q_MAIN Performance Enhancement Circuit [55]

This modification involves adding a series resistor to the already existing $0.02\mu\text{F}$ capacitor at the LTC1871's GATE pin, along with a parallel diode with that series resistor. This project omits the additional gate resistor (R3 in figure 10.31) because in LTSpice simulations adding that resistor resulted in increased power dissipation across Q_MAIN. Because the SEPIC's input voltage step may reach faster than 5V per ms from

the Precor elliptical trainer's output, the $0.02\mu\text{F}$ capacitor at the LTC1871's GATE pin is necessary for preventing Q_MAIN's self-enhancement. The additional series resistor (R1 in figure 10.31) helps improve Q_MAIN's turn-off time and the parallel diode helps Q_MAIN maintain fast switching trajectories during turn-on. Furthermore, the series RC combination acts as a low pass filter for suppressing high frequency harmonics. Figure 10.32 shows this modification to the SEPIC's schematic.

Figure 10.32: LTC1871 GATE Pin Damping Filter

The red arrow in figure 10.27 denotes the GATE pin filter location. Earlier in this chapter, this project used decade capacitor boxes for CGATE_MAIN and C_{FB}. Those decade capacitor boxes yielded better overall converter performance than using discrete

capacitors, and that was because of additional series resistance that those decade capacitor boxes had. This project's author measured the ESR of those decade capacitor boxes as roughly 45Ω . Therefore this project selects 45Ω for the series gate resistor (RGATE_MAIN in figure 10.32) for the LTC1871's GATE Pin damping filter, because it closely matches the decade capacitor box's ESR that helped improve converter functionality in earlier tests. This project uses the same LTC1871 GATE pin capacitor value as before ($0.02\mu\text{F}$) for the damping filter capacitor (CGATE_MAIN in figure 10.32) because that was the optimal Q_MAIN gate capacitance value determined from previous sections. The LTSpice simulation results with this new damping filter are roughly the same as from the last test run, therefore this project omits those simulation results. Unlike the LTC1871 current sense pin RC low pass filter, this damping filter does not require severing any existing PCB traces leading to Q_MAIN's gate as well as the LTC1871 controller's GATE pin.

Tables E-33 and E-34 show the hardware results from adding this damping filter to the LTC1871's GATE pin, while keeping all other SEPIC components intact. Table E-33 shows the SEPIC's output load driving abilities while using the electronic load in constant voltage mode (36V) and Table E-34 shows the same test results, except using input voltages (with their corresponding elliptical training resistance levels) that match the 160 strides per minute exercise test results from the previous EHFEM group in [1].

Table E-33: SEPIC Load Tests with LTC1871 Current Sense Pin Filter and GATE Pin Damping Filter and BK 9153 Power Supplies

V_{IN} (V)	I_{IN} (A)	V_{OUT} (V)	I_{OUT} (A)	V_{BAT} (V)	I_{BAT} (A)	P_{IN} (W)	P_{OUT} (W)	η (%)	R_{IN} (Ω)
5	0.389	37.39	0.019	12	0.024	2.233	0.710	31.814	12.853
10	1.002	37.37	0.163	12	0.024	10.308	6.091	59.093	9.980
12	1.227	37.36	0.253	12	0.023	15.000	9.452	63.014	9.780
15	1.538	37.34	0.418	12	0.023	23.346	15.608	66.856	9.753
18	1.83	37.33	0.618	12	0.023	33.216	23.070	69.454	9.836
20	2.057	37.31	0.778	12	0.023	41.416	29.027	70.087	9.723
25	2.548	37.27	1.258	12	0.023	63.976	46.886	73.286	9.812
27	2.69	37.25	1.448	12	0.023	72.906	53.938	73.983	10.037
30	3.013	37.22	1.838	12	0.023	90.666	68.410	75.453	9.957
36	3.652	37.14	2.748	12	0.023	131.748	102.061	77.467	9.858
40	4.05	37.07	3.418	12	0.023	162.276	126.705	78.080	9.877
45	4.516	37	4.318	12	0.023	203.496	159.766	78.511	9.965
50	5.075	36.89	5.418	12	0.023	254.026	199.870	78.681	9.852
52	5.194	36.86	5.768	12	0.023	270.364	212.608	78.638	10.012
60	5.863	36.7	7.498	12	0.023	352.056	275.177	78.163	10.234

Table E-34: SEPIC Load Test with LTC1871 Current Sense Pin Filter and GATE Pin Damping Filter, Using 160 Strides/Min. Theoretical Input Voltages from [1] and BK 9153 Power Supplies

V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	V _{BAT} (V)	I _{BAT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Training Resistance Level
8.08	0.792	37.38	0.088	12	0.024	6.687	3.289	49.189	10.202	2
11.54	1.193	37.36	0.228	12	0.023	14.043	8.518	60.656	9.673	3
13.49	1.401	37.36	0.328	12	0.023	19.175	12.254	63.905	9.629	4
15.88	1.629	37.34	0.468	12	0.023	26.145	17.475	66.840	9.748	5
18.27	1.849	37.33	0.628	12	0.023	34.057	23.443	68.835	9.881	6
21	2.163	37.31	0.868	12	0.023	45.699	32.385	70.866	9.709	7
23.3	2.363	37.29	1.068	12	0.023	55.334	39.826	71.973	9.860	8
26.2	2.682	37.26	1.398	12	0.023	70.544	52.089	73.839	9.769	9
29.1	2.919	37.23	1.718	12	0.023	85.219	63.961	75.055	9.969	10
31.8	3.272	37.19	2.138	12	0.023	104.326	79.512	76.215	9.719	11
34.3	3.519	37.16	2.503	12	0.023	120.978	93.011	76.883	9.747	12
35.1	3.577	37.15	2.618	12	0.023	125.829	97.259	77.295	9.813	13
39.9	4.039	37.08	3.398	12	0.023	161.432	125.998	78.050	9.879	14
44.8	4.519	37	4.298	12	0.023	202.727	159.026	78.443	9.914	15
46.4	4.68	36.97	4.618	12	0.023	217.428	170.727	78.521	9.915	16
47.6	4.781	36.95	4.848	12	0.023	227.852	179.134	78.619	9.956	17
48.3	4.883	36.93	5.028	12	0.023	236.125	185.684	78.638	9.891	18
58.2	5.878	36.72	7.298	12	0.023	342.376	267.983	78.272	9.901	19
59.8	5.877	36.71	7.498	12	0.023	351.721	275.252	78.259	10.175	20

Adding the damping filter from figure 10.31 to Q_MAIN's gate dramatically improves SEPIC performance, as the SEPIC now drives higher load currents at all input voltages. Despite added resistance on the LTC1871's GATE pin, I_{BAT} remains constant relative to previous tests. At 60V input the SEPIC drives a 275W load (7.498A output current), while taking in roughly 350W from the input power supply. For this set of tests, this project used constant current mode on the electronic load for all input voltages because for all input voltages, the SEPIC now draws the absolute maximum current limit (6.5A) when using constant 36V output mode on the electronic load. This means that the SEPIC can drive higher load currents at all input voltages relative to previous tests, but this

project only selects current limits that allow the SEPIC to maintain a 10Ω input resistance. Figure 10.33 shows the current SEPIC's maximum load current driving ability at all tested input voltages while still maintaining a 36V output. Figure 10.34 shows the current SEPIC's efficiency at all tested input voltages.

Figure 10.33: SEPIC Maximum Load Current Driving Ability vs. Input Voltage with LTC1871 Current Sense Pin Filtering and GATE Pin Damping Filter

Figure 10.34: SEPIC Efficiency vs. Input Voltage with LTC1871 Current Sense Pin Filtering and GATE Pin Damping Filter

The SEPIC now drives a 7.498A maximum load at 60V input, and its efficiency peaks at nearly 79%, which is an almost 10% improvement from the previous design.

Theoretically, this project's SEPIC can drive up the 8A maximum specified load current outlined in section 1.2, but because this converter only uses 7.5A output fuses, this project was unable to test with output load currents at the 8A level. Therefore, implementing this SEPIC with a Precor elliptical trainer for Cal Poly Recreational Center operation would necessitate a 10A blade fuse, which not only helps the SEPIC supply its specified maximum 8A load current but 10A is also the Enphase M175-24-240 micro-inverter's absolute maximum input short circuit current rating. Likewise with the previous test results, maximum load current driving ability and efficiency increase with input voltage. Overall, adding filtering to the LTC1871's critical control pins (the FB, GATE and SENSE pins) is necessary for proper converter operation. The SEPIC's

efficiency now surpasses the 75% requirement outlined in section 1.2. Thus, adding a damping circuit to the LTC1871's GATE pin was the crucial key for reducing the SEPIC's switching and conduction losses, as well as decreasing Q_MAIN's operating stresses. From tables E-33 and E-34, overall maximum input power decreased at all input voltages, relative to the results from E-31 and E-32, indicating that the losses incurred across L₁, L₂ and Q_MAIN decreased as well. Overall SEPIC efficiency is still poor at lower input voltages; however this is because the SEPIC and its PWM controller operate in CCM for all input voltages, rather than DCM at lower input voltages, resulting in greater magnetic losses in L₁ and L₂ at lighter output load currents. For the typical Precor elliptical workout output's 15-30V SEPIC input range, the efficiency values suffice. At around 30V input, the converter reaches its specified 75% efficiency. 30V is the peak input for a typical workout session on the Precor elliptical trainer. (10.2) calculates this project's SEPIC's line regulation. For calculating line regulation, (10.2) uses the SEPIC's output voltages when the SEPIC operates at 8V input and 60V input using the maximum load current levels from tables E-33 and E-34.

$$Line\ Reg. = \frac{V_{OUT_{8V\ input}} - V_{OUT_{60V\ input}}}{V_{OUT_{nominal}}} \times 100\% = \frac{37.39V - 36.7V}{36V} \times 100\% = 1.92\% \quad (10.2)$$

The SEPIC experiences an overall 1.92% line regulation when its input changes from 8V to 60V. This value is lower than the 5% requirement outlined in section 1.2, therefore it suffices. Figures 10.35 and 10.36 show the SEPIC's current sense pin and Q_MAIN switching trajectory waveforms with all the converter enhancement modifications mentioned in this chapter. In both figures the SEPIC runs at 36V input along while driving a 2.748A output load.

Figure 10.35: LTC1871 Current Sense Pin Switching Waveform with All SEPIC Performance Enhancement Modifications at 36V Input and 2.748A Load

Figure 10.36: Q_MAIN Gate Switching Trajectory with All SEPIC Performance Enhancement Modifications at 36V Input and 2.748A Load

Although there are still harmonics present in both waveforms, such harmonics are much lower in magnitude than from previous SEPIC designs and the waveforms look closer to the ideal waveforms in figure 5.9. This project could make further improvements at eliminating those harmonics by tweaking the LTC1871's FB, GATE and SENSE pin filter component values, however, lack of time prevents those improvements from taking place. Figures 10.37 to 10.39 show the SEPIC's output voltage ripple in its three main DC-DC conversion modes – boost, buck-boost and boost. Figure 10.37 uses 30V input while driving a 1.838A load for boost mode, figure 10.38 uses 36V input while driving a

2.748A load for buck-boost mode, and figure 10.39 uses 60V input while driving a 7.498A load for buck mode.

Figure 10.37: SEPIC Output Voltage Ripple at 30V Input and 1.838A Load (Boost Mode)

Figure 10.38: SEPIC Output Voltage Ripple at 36V Input and 2.748A Load (Buck-Boost Mode)

Figure 10.39: SEPIC Output Voltage Ripple at 60V Input and 7.498A Load (Buck Mode)

From figures 10.37 to 10.39, the SEPIC's peak to peak output voltage ripple worsens with increasing input voltage and output load current. With 30V input, the SEPIC's peak to peak output voltage ripple is 1.8V. It increases to 2V at 36V input and in reaches a 2.6V maximum at 60V input. At worst this corresponds to a 7.08% ripple at 60V input relative to the 36.7V output voltage, which is under the maximum 10% output peak to peak ripple specification outlined in section 1.2. Therefore, the 940 μ F output capacitance from the two output electrolytic capacitors suffice, although future improvements to this SEPIC design should use a single electrolytic capacitor with low ESR. Also, although during this test the project removed the inductor wire loops from the SEPIC's PCB, Q_MAIN's switching waveform at 36V input from figure 10.36 show no major harmonic

transients. Therefore inductors L_1 and L_2 's current waveforms should not contain any dead time (which those switching harmonics caused) and thus L_1 and L_2 should run in CCM at 31V input and 2A load, as specified in section 1.2. Therefore, this project safely assumes that both SEPIC inductors run in CCM at the specified input voltage and output load levels. The current modifications made to the current SEPIC design already satisfy the specifications requirements outlined in sections 1.1 to 1.3, therefore no further modifications are necessary because of project time constraints. Figure 10.40 shows this project's SEPIC's final schematic.

Figure 10.40: Final SEPIC Schematic

Table E-35 shows this project's SEPIC's final BOM and total cost.

Table E-35: Final SEPIC BOM

PCBs	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Custom 2-layer 5.8" x 5.275"	1	84.42	84.42	PCB	2-layer	N/A	N/A
Total Cost (\$)			84.42				
Controller I.C.s:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
LTC1871HMS (10-MSOP)	1	4.83	4.83	U1	10-MSOP	N/A	N/A
LT4356-1HMS (10-MSOP)	2	3.70	7.40	U2	10-MSOP	N/A	N/A
Total Cost (\$)			12.23				
Transistors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Infineon IPP110N20N3	1	8.06	8.06	Q_MAIN	TO-220-3	2.45W	4.01W
IXYS IXFK230N20T	1	9.50	9.50	Q_OVP	TO-264	1.625kW	1.625kW
Total Cost (\$)			17.56				
Diodes:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
MBR20100CT (Schottky)	7	0.88	6.16	D_MAIN1, D_MAIN2, D_MAIN3, D_MAIN4, DSNUB2, DSNUB3, DSNUB4	TO-220-3	31.11A, 110V	27.72A, 56.46V
Microsemi APT30S20BG (Schottky)	1	3.67	3.67	DSNUB1	TO-247	39.43A, 15.95V	27.89A, 9.58V
1N5378B (Zener)	1	0.57	0.57	TSD	Through-Hole	100V	100V
SML-LX1206SRC-TR (LED)	1	0.43	0.43	FLT_LED	1206	1.2V min. forward drop	1.2V min. forward drop
1N4148-TR (Silicon)	1	0.06	0.06	DGATE_MAIN	Through-Hole	N/A	N/A
Total Cost (\$)			10.89				
Heatsinks:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
531202B02500G (TO-220)	1	1.38	1.38	Q_MAIN heatsink	TO-220	N/A	N/A
MV-102-55E (TO-247 and TO-264)	1	7.39	7.39	Q_OVP heatsink	TO-264	N/A	N/A
WA-T247-101E (TO-247)	1	2.07	2.07	DSNUB1 heatsink	TO-247	N/A	N/A
574502B00000G (TO-220)	7	0.27	1.89	D_MAIN1, D_MAIN2, D_MAIN3, D_MAIN4, DSNUB2, DSNUB3, DSNUB4	TO-220	N/A	N/A
Total Cost (\$)			12.73				

Table E-35 (Continued)

Fuses:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Radio Shack 270-270-1234 30A Blade Fuse Holder	2	2.99	5.98	Fuse Holder for Input and Output Fuse	N/A	65V, 25A	5V, 25A
Cooper Bussmann 270-1085 30A Blade Fuse	1	0.73	0.73	Input Fuse	Blade Fuse	65V, 25A	5V, 25A
Cooper Bussmann 270-1081 10A Blade Fuse	1	0.73	0.73	Output Fuse	Blade Fuse	36V, 8A	5V, 0.5A
Bourns SF-0402F200-2 2A Fuse	1	0.68	0.68	Elliptical Battery Fuse	0402	12V, 2A	12V, 2A
Total Cost (\$)			8.12				
Resistors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
274k, 1/8W, 1%, ERJ-6ENF2743V	1	0.07	0.07	RRUN1_MAIN	0805	0.521mW	0.522mW
255k, 1/8W, 1%, ERJ-6ENF2553V	1	0.07	0.07	RRUN1_OVP	0805	16.289mW	0.094mW
220k, 1/8W, 1%, ERJ-6ENF2203V	1	0.07	0.07	RFREQ	0805	0.002mW	0.002mW
133k, 1/8W, 1%, ERJ-6ENF1333V	1	0.07	0.07	RRUN2_MAIN	0805	0.115mW	0.122mW
115k, 1/8W, 1%, ERJ-6ENF1153V	1	0.07	0.07	RFB1	0805	11.12mW	12.28mW
33.2k, 1/8W, 1%, ERJ-6ENF3322V	1	0.07	0.07	RTH	0805	28.32mW	0.028mW
4.99k, 1/8W, 1%, ERJ-6ENF4991V	1	0.07	0.07	RRUN2_OVP	0805	0.319mW	0.002mW
3.92k, 1/8W, 1%, RMCFO805FT3K92	1	0.04	0.04	RFB2	0805	0.387mW	0.427mW
3.6k, 1/8W, 1%, ERJ-6ENF3601V	1	0.07	0.07	RFLT	0805	32.4mW	32.4mW
887, 1/8W, 1%, ERJ-6ENF8870V	1	0.07	0.07	RVCC1	0805	< 125mW	< 125mW
680, 1/8W, 5%, ERJ-6GEYJ681V	1	0.04	0.04	RVCC2	0805	< 125mW	< 125mW
47, 1/8W, 5%, CF18JT47R0	1	0.09	0.09	RGATE_MAIN	Through-Hole	< 125mW	< 125mW
10, 1/8W, 1%, ERJ-6ENF10R0V	2	0.07	0.14	RGATE_OVP, RSNS_FILT	0805	13.05mW	0W
0.005, 5W, 1%, OARSXPR005FLF	1	1.42	1.42	RSNS	SMT (Custom)	410mW	1.132W
0.002, 5W, 1%, WSLP39212L000FEB	1	3.07	3.07	RSNS_OVP	SMT (Custom)	64mW	1.000W
Total Cost (\$)			5.43				

Table E-35 (Continued)

Inductors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
Vishay IHV28BZ60 (60 μ H, I_{SAT} =28A)	2	20.30	40.60	L1, L2	Through-Hole	30.68A (max.), 65V	22.37A (max.), 52.22V
Vishay IHLP4040DZERR36M01 (360nH, I_{SAT} =60A)	1	2.52	2.52	LSNUB	SMT (Custom)	29.3A (max.), 106.25V	17.85A (max.), 55.35V
Total Cost (\$)			43.12				

Table E-35 (Continued)

Capacitors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
1000μ, Electrolytic, 63V, B41695A8108Q7	1	5.15	5.15	COUT	Through-Hole	36V, 5.43A (RMS)	36V, 2.368A (RMS)
1000μ, Electrolytic, 100V, UPW2A102MHD	1	3.03	3.03	CIN	Through-Hole	65V, 1.886pA (RMS)	5V, 227.67fA (RMS)
47μ, Electrolytic, 100V, AFK476M2AH32T-F	1	2.35	2.35	CCLAMP	SMT (Custom)	65.68V, 119.05mA (RMS)	5V, 42.067mA (RMS)
33μ, Electrolytic, 35V, EEE-HA1V330WP	1	0.63	0.63	CBAT	SMT (Custom)	12V, 206.56fA (RMS)	12V, 442.09fA (RMS)
15μ, Ceramic, 250V, KHD251E156M99C0B00	1	25.78	25.78	CINT	Through-Hole	77.05V, 3.67A (RMS)	21.22V, 1.404A (RMS)
10μ, Ceramic, 50V, UMK325C7106MM-T	3	0.96	2.88	COUT2, COUT3, COUT4	1210	36V, 2.758A (RMS)	36V, 2.016A (RMS)
10μ, Ceramic, 25V, TMK316B7106KL-TD	2	0.58	1.16	CBAT2, CBAT3	1206	12V, 963.63fA (RMS)	12V, 1.211pA (RMS)
4.7μ, Ceramic, 100V, C4532X7S2A475M	5	1.75	8.75	CINT2, CINT3, CINT4, CINT5, CINT6	1812	77.05V, 1.22A (RMS)	21.22V, 1.22A (RMS)
4.7μ, Ceramic, 16V, EMK212B7475KG-T	1	0.38	0.38	CVCC	0805	5.23V, 153.1mA (RMS)	5.23V, 170.84mA (RMS)
0.56μ, Ceramic, 25V, C0805C564K3RACTU	1	0.84	0.84	CSNUB1	0805	11.33V, 3.629A (RMS)	11.00V, 1.577A (RMS)
0.1μ, Ceramic, 100V, CGA3E3X7S2A104K	7	0.30	2.10	CIN2, CIN3, CCLAMP2, CCLAMP3, CCLAMP4, CBAT4, CBAT5	0603	65V, 1.000mA (RMS)	5V, 0.285mA (RMS)
0.1μ, Ceramic, 100V, HMK212B7104KG-T	4	0.30	1.20	CINT7, CINT8, CINT9, CINT10	0805	77.05V, 47.184mA (RMS)	21.22V, 17.59mA (RMS)
0.1μ, Ceramic, 50V, C1608X7R1H104M	4	0.16	0.64	COUT_CPH1, COUT_CPH2, COUT_CPH3, COUT_CPH4	0603	36V, 63.00mA (RMS)	36V, 46.00mA (RMS)
0.033μ, Ceramic, 50V, RPER71H333K2K1A03A	1	0.10	0.10	CSNS2	Through-Hole	1.5V, 0.028mA (RMS)	1.307V, 0.020mA (RMS)
0.02μ, Ceramic, 50V, RDEF51H103Z0K1C03B	1	0.22	0.22	CGATE_MAIN	Through-Hole	5V, 50mA (RMS)	5V, 50mA (RMS)
10000p, Ceramic, 16V, C1005X7R1C103K	1	0.10	0.10	CFB2	0402	1.5V, 0.028mA (RMS)	1.307V, 0.020mA (RMS)
10000p, Ceramic, 250V, C2012X7R2E103K	1	0.33	0.33	CSNUB2	0805	103.44V, 1.619A (RMS)	56.06V, 756.43mA (RMS)
6800p, Ceramic, 100V, C1608X8R2A682K	1	0.36	0.36	CGATE_OVP	0603	77.47V, 4.796mA (RMS)	11.331V, 0.006mA (RMS)
6800p, Ceramic, 25V, C1005X7R1E682K	1	0.09	0.09	CTH2	0402	1.232V, 0.110mA (RMS)	1.232V, 0.030mA (RMS)
56p, Ceramic, 10V, C0402C560J8GACTU	1	0.35	0.35	CTH1	0402	1.232V, 0.390mA (RMS)	1.232V, 0.009mA (RMS)

Table E-35 (Continued)

Capacitors:	Quantity:	Unit Price (\$):	Total Price:	Components:	Type:	65V, 8A Required Rating	5V, 0.5A
47p, Ceramic, 10V, 0603ZA470DAT2A	5	0.30	1.50	CFB1, CSNS1, CTMR, CRUN_OVP, CRUN_MAIN	0603	1.5V, 0.028mA (RMS)	1.307V, 0.020mA (RMS)
Total Cost (\$)			57.94				
Component Count	88						
Total Converter Cost (\$)	252.44						

Because single 940 μ F electrolytic capacitors were unavailable from Digi-Key, this project changes C_{OUT} to a 1000 μ F capacitor. Using a single 1000 μ F capacitor for C_{OUT} decreases the total converter cost. This final SEPIC costs \$252.44, including components, heatsinks and a PCB. Overall, the additional parts do not add much to the converter's overall cost and are necessary for proper operation and meeting all guidelines in sections 1.1 to 1.3. Although this SEPIC costs more than the Vicor Maxi 28V DC-DC converter used by the EHFEM group in [1], this SEPIC functions over a wider input voltage range, which satisfies this project's purpose. For the final portions of this report, this project tests this SEPIC design with the full Precor elliptical trainer generation system for functionality, investigates efficiency losses and describes further improvements for this SEPIC design.

10.4 Precor Elliptical Trainer Full System Hardware Testing

The final hardware testing round tests this project's final SEPIC design with the Precor elliptical trainer as the SEPIC's input voltage source. This project first tests the SEPIC with a constant 10 Ω output load before using the Enphase micro-inverter as a load. However, this project later could not conduct any tests using the Enphase micro-inverter as a load because initializing the micro-inverter caused an open load condition at the elliptical trainer's output, causing it to output voltage levels greater than 100V and subsequently destroy the SEPIC's input pass transistor (Q_OVP) and main switching transistor (Q_MAIN). Figures 10.41 and 10.42 show this project's SEPIC's test setup with the Precor elliptical trainer.

Figure 10.41: SEPIC Testing with Precor Elliptical Trainer, Angled View

Figure 10.42: SEPIC Testing with Precor Elliptical Trainer, Instrument and Grid Panel View

The first full system test uses a BK Precision 8510 600W electronic load for simulating the constant 10Ω resistor load connected to the SEPIC's output, because for this test this project connects the SEPIC's main input capacitor C_{IN} in series with the elliptical trainer's onboard 10Ω resistor coils for limiting C_{IN} 's RMS current to safe levels (1.52A maximum). Figure 10.40 shows this series resistor connection with C_{IN} . A BK Precision 9153 DC power supply simulates the elliptical trainer's onboard 12V battery because this project's author could not obtain any test leads long enough for connecting the elliptical trainer's battery to the SEPIC. Replacing the elliptical trainer's battery with a 12V DC power supply does not affect the SEPIC's performance. This project also sets up the

testing bench for connecting the Enphase micro-inverter to the 240V (RMS) grid. Also, for safety purposes, this project's author covered the SEPIC with a cardboard box visible in figures 10.41 and 10.42. Table E-36 shows this test's results.

Table E-36: SEPIC Test Results with Precor Elliptical Trainer Input and Constant 10 Ω Load

Training Resistance Level	V _{IN} (V)	I _{IN} (A)	V _{OUT} (V)	I _{OUT} (A)	P _{IN} (W)	P _{OUT} (W)	η (%)	R _{IN} (Ω)	Error (V _{IN})	Error (I _{IN})	Error (V _{OUT})	Error (I _{OUT})
2	2.5	2.319	18.04	0.166	5.80	2.99	51.65	1.08	1.5	1.5	2.5	0.8
3	4	2.995	23.4	0.307	11.98	7.18	59.96	1.34	1.2	1.2	3.4	0.045
4	5.6	5	28.3	0.694	28	19.64	70.14	1.12	0.5	1	1	0.05
5	11	4.2	35	0.967	46.2	33.85	73.26	2.62	1	0.2	0.8	0.05
6	13.7	4.5	37.33	1.285	61.65	47.97	77.81	3.04	1	0.2	0.5	0.05
7	16.7	4.734	37.33	1.72	79.06	64.21	81.22	3.53	2	1	0	0.5
8	18.8	5.2	37.44	2.23	97.76	83.49	85.40	3.62	2	1.5	0	0.5
9	23.3	5.493	37.44	2.5	127.99	93.6	73.13	4.24	2	1.5	0	0.1

Table E-36 shows the Precor elliptical trainer's selected training resistance level as well as the SEPIC's input voltage from the elliptical trainer's output (V_{IN}), input current (I_{IN}), SEPIC output voltage (V_{OUT}), output current (I_{OUT}), elliptical trainer output power (P_{IN}), SEPIC output power (P_{OUT}), efficiency (η), SEPIC input resistance (R_{IN}), as well as error values for V_{IN} , I_{IN} , V_{OUT} and I_{OUT} . The SEPIC's input voltage, input current, output voltage and output current fluctuated as the participant exercised on the Precor elliptical trainer, indicating that the elliptical trainer's unstable, rectified DC output voltage caused those fluctuating values in the SEPIC. However, at training resistance levels 8 and 9 the SEPIC's input voltage and input current were more stable than at lower elliptical machine training resistance levels, resulting in more stable output current levels. The last four columns in table E-36 indicate error values, which indicate the observed deviations from the most frequently occurring values (which this project uses for the V_{IN} , I_{IN} , V_{OUT} and I_{OUT} values in table E-36) for V_{IN} , I_{IN} , V_{OUT} and I_{OUT} . These error values tend to begin decreasing at higher training resistance levels. For safety reasons this project only tests the Precor elliptical trainer up to resistance level 9. Also, the SEPIC begins regulating 36V output around training resistance level 5 from the elliptical trainer. Resistance levels 2 and 3 do not produce the required minimum input threshold voltage (5V) for the SEPIC regulating 36V output. Figures 10.43 and 10.44 show the SEPIC's output load current driving ability as well as efficiency from the results in table E-36.

Figure 10.43: SEPIC Load Current Driving Ability vs. Elliptical Training Resistance Level with Precor Elliptical Trainer Input

Figure 10.44: SEPIC Efficiency vs. Elliptical Training Resistance Level with Precor Elliptical Trainer Input

From figure 10.43, this project's SEPIC outputs 2.5A to the 10 Ω electronic load at elliptical training resistance level 9. From figure 10.44, the SEPIC reaches 85% peak efficiency at elliptical training resistance level 8 and 73% efficiency at training level 9. The 85% efficiency figure may be inaccurate because of fluctuating input voltages and currents from the elliptical trainer. Although loading the SEPIC with a constant resistive load and using the elliptical trainer as its input voltage source doesn't maintain the 10 Ω input resistance required for keeping the elliptical trainer's eddy current braking system intact, the participant during this test did not report any deviation in proper elliptical trainer user experience as he changed the elliptical trainer's training resistance level – in other words, higher resistance levels were more difficult for him to exercise at and lower resistance levels were easier. However, this test shows that this project's SEPIC functions with the Precor elliptical trainer, provided that its output load resistance remains constant.

Unfortunately, this project could not test the entire EHFEM system with the Enphase micro-inverter. One characteristic that this project's author did not know about the Enphase micro-inverter is that its energy monitoring unit (EMU) has a roughly 20 minute warm-up period for detecting if the micro-inverter is connected to the grid. Afterwards, the Enphase micro-inverter requires an additional 5 minute initialization period before it functions with a DC input source. During this entire time period, the SEPIC had an open load (i.e. no load), causing the Precor elliptical trainer to sense its output as having an open load. One other Precor elliptical trainer aspect that this project did not characterize was how it behaved with no load (i.e. by removing its 10 Ω resistor coils). As the participant exercised on the elliptical trainer during the Enphase micro-

inverter's warm-up period, the elliptical trainer outputted voltages exceeding 100V. The participant then slowed down the workout pace such that the elliptical trainer outputted 60V, but that voltage soon increased back to 100V. The SEPIC's LT4356-1 input overvoltage and overcurrent protection circuit has a 100V input limit, as well as the input zener diode. The output voltage exceeding 100V from the elliptical trainer then destroyed the SEPIC's input zener diode, as well as its Q_OVP and Q_MAIN transistors. Those voltages also physically destroyed the LT4356-1 IC. This project's author theorizes that the LT4356-1 input circuit did not respond fast enough to the changes between the 60V and 100V output voltages from the elliptical trainer and allowed excess current flowing through the SEPIC. That excess current first destroyed Q_MAIN, causing a short across it, and that in turn caused excessive drain-source voltage across and drain current through Q_OVP, resulting in excessive power dissipation and thermal runaway across Q_OVP, destroying it in a chain reaction. Lastly, the zener diode could not tolerate the excessive inrush current and hence was destroyed as well. Therefore, this project could not test the SEPIC's functionality with the Enphase micro-inverter because excessive output levels from the Precor elliptical trainer destroyed crucial SEPIC components. However, this test proved that the Precor elliptical trainer cannot function properly with an open load and that it should only output power once the micro-inverter finishes initializing.

However, as speculated from [10], this project's SEPIC would behave in the same manner as the designed DC-DC converter from [10] when loaded with the Enphase micro-inverter. The author in [10] speculates that the Enphase micro-inverter implements a hill-climbing maximum power point tracking (MPPT) algorithm for maximum power

transfer. As observed in [10], the Enphase micro-inverter begins power tracking at its input voltage level (output voltage from the DC-DC converter) and then slowly increased its input current (output current from the connected DC-DC converter) provided that it could still receive power from the connected DC-DC converter. When the Enphase micro-inverter increased its input current, its input resistance decreased, decreasing the DC-DC converter's load resistance. Furthermore, when the Enphase micro-inverter increased its input current beyond a level that the DC-DC converter could tolerate while regulating 36V at its output, the DC-DC converter began de-regulating and hence the DC-DC converter's output voltage dropped below 36V. This behavior is certainly not desirable for this project's EHFEM system application and furthermore, it does not maintain 10Ω input resistance across the DC-DC converter's input. This behavior functions properly with a photovoltaic panel, as Enphase designed its M175-24-240 micro-inverter for connecting to a photovoltaic panel. However, this project designed its SEPIC for increasing its output load current with input voltage, which is the opposite behavior of a photovoltaic panel. Overall, from the results in [10], this project's SEPIC would produce the same behavior as the DC-DC converter in [10] when loaded with the Enphase micro-inverter – the Enphase micro-inverter would draw as much current as possible from the SEPIC until either its input or output fuse blew. Therefore, this project's SEPIC would not function as this project desires when loaded with the Enphase micro-inverter. Getting this project's SEPIC to function properly with the Enphase micro-inverter, however, requires an additional control mechanism for the Enphase micro-inverter such that it limits the micro-inverter's input current based on the SEPIC's input

voltage (for maintaining 10Ω resistance at the SEPIC's input). Such control mechanisms require additional research and are best suited for a separate project, but for the most part, this project's SEPIC met its design requirements, specifications and constraints outlined in sections 1.1 to 1.3 when tested with a DC power supply as its input. Furthermore, this project's SEPIC functions with the Precor elliptical trainer while loaded with a constant resistance.

CHAPTER 11: CONCLUSION, FUTURE IMPROVEMENTS AND FUTURE EHFEM PROJECTS

11.1 Conclusion

Overall, this project's SEPIC satisfied the design requirements, specifications and constraints outlined in sections 1.1 to 1.3. At 60V input this project's SEPIC outputs 275W while taking a 352W input. This corresponds to 78.3% converter efficiency at 60V input. The SEPIC's peak efficiency level is 78.7%, and it reaches this level at 50V input. Based on these results, this project's SEPIC can also tolerate 65V input and can drive an 8A load. However this project did not test the SEPIC under those conditions because of test equipment electrical limitations. This project's SEPIC also runs in CCM at 31V input and 2A load as specified in section 1.2. Furthermore, its line regulation when its input changes from 8-60V (and while driving a load under that input voltage range) is 1.89%, which is well below the 5% maximum level specified in section 1.2. Although this project's SEPIC did not meet its \$120 price limit for achieving a 10-year EHFEM system payback period, as mentioned in section 2.3.1, implementing this project's EHFEM system in a frequently-used gym such as Cal Poly's Recreation Center can also raise electricity conservation awareness and reduce overall energy consumption in the city of San Luis Obispo. Furthermore, it can also draw attention from potential donors and sponsors that may frequent Cal Poly's recreation Center, leading to reduced overall system lifecycle costs. Proper SEPIC functionality with this project's selected Enphase M175-24-240 micro-inverter, however, requires additional control mechanisms for the inverter. However, this project did prove that the SEPIC functions with the Precor EFX

546i elliptical trainer as its input as well as with a constant output load. Therefore, this project's SEPIC would function properly with an inverter that maintains a constant load or a variable load such that it maintains 10Ω input resistance. The latter case, however, is more desirable for the EHFEM system and as mentioned already it requires an additional control mechanism for the Enphase micro-inverter. This chapter later describes such possible mechanisms. However, this project proved that a PWM-switching SEPIC topology provides a functional DC-DC converter for a dynamic DC voltage generation system such as the Precor elliptical trainer as well as inverter interfacing, given that the inverter takes in 36V and limits its input current based on the Precor elliptical trainer's output voltage. Therefore, based on its test results, this project proves that a PWM-switching SEPIC topology provides a functional DC-DC converter design for DC power generation and inverter interfacing from a dynamic input voltage generator because of its wide input voltage range, high power driving capability and inherent voltage step-up and step-down functions. The next sections describe possible improvements for this project's SEPIC design, as well as future projects for the EHFEM project as a whole.

11.2 Future SEPIC Improvements

Despite the fact that this project's SEPIC satisfied all design requirements, specifications and constraints, numerous modifications can improve its functionality and efficiency performance, as well as reduce its cost. Such improvements include adding output current limiting circuitry, eliminating unnecessary PCB footprints and reducing the size of PCB copper planes in the SEPIC's power path, as well as adding modifications for conserving elliptical trainer battery use, and using better input

protection methods than the LT4356-1 protection circuit used in this project.

Furthermore, this project's SEPIC design also creates important EHFEM projects for future senior project or master thesis students. As mentioned earlier, one such important project is a control mechanism for the currently used Enphase M175-24-240 micro-inverter for helping maintain a constant 10Ω input resistance to this project's SEPIC.

This project's SEPIC could also use a single custom coupled inductor for housing L_1 and L_2 rather than two discrete inductors, potentially decreasing converter cost as well as creating another project for future EHFEM project students. Lastly, another potential project includes designing a grid-tie inverter that properly controls its input current for proper functionality with this project's SEPIC.

11.2.1 SEPIC Output Current Limiting

The first improvement that this project's SEPIC necessitates is output current limiting. Output current limiting is a standard feature in most offline DC power supplies. This feature prevents power supply components from exceeding their electrical limits as well as provides electrical safety to its end-user. Although this project's SEPIC's output fuses already provide overcurrent protection, the EHFEM system's owner must replace them every time an overcurrent fault occurs at the SEPIC's output. This causes inconvenience as well as brings added operation costs (for replacing the fuses) for the EHFEM system's owner. Furthermore, if the fuses do not react quickly enough to an overcurrent fault, then crucial SEPIC components can be destroyed, adding further component replacement costs for the EHFEM system's owner. Therefore, output current limiting greatly benefits this project's SEPIC in eliminating those mentioned problems. A

relatively simple current limiting circuit that future EHFEM project students can expand on is a crowbar circuit, or simply known as a crowbar. A crowbar detects an overcurrent or overvoltage fault at the SEPIC's output (typically through a current sense resistor) and shorts the SEPIC's output to a low voltage, typically via a silicon-controlled rectifier (SCR) or thyristor [72]. SCRs typically can tolerate very high current levels while maintaining low forward voltages, resulting in very low power dissipation during an overcurrent fault. Using an active crowbar (rather than a passive one) returns the SEPIC's output to normal operation once the fault disappears. Figure 11.1 shows an active overvoltage crowbar (which can be used at the SEPIC's input as an alternative to the LT4356-1) and figure 11.2 shows an overcurrent crowbar [72, 73].

Figure 11.1: Active Overvoltage Crowbar [72]

Figure 11.2: Overcurrent Crowbar [73]

Using a crowbar for current limiting requires an additional current sense resistor.

However, the main flaw with using a crowbar as an output current limiter for this project's SEPIC is that a crowbar drops the SEPIC's output voltage to a level below the Enphase micro-inverter's input voltage range during an overcurrent fault. Ideally this project desires keeping the SEPIC's output voltage at 36V for optimal EHFEM system operation, even during an overcurrent fault. However, future projects can make modifications to the default overcurrent crowbar for limiting SEPIC output current while maintaining 36V output voltage. In Spring Quarter 2011, EHFEM team member Greg Hollister attempted implementing a DC-DC converter output current limiter that maintained 36V output voltage for his DC-DC converter [53]. Greg Hollister's implementation uses a multiplier circuit that limits the DC-DC converter's output current proportionally to the squared value of the converter's input voltage. This implementation

would thus let a DC-DC converter maintain a constant 10Ω input resistance. Figure 11.3 shows his current limiter circuit [53].

Figure 11.3: DC-DC Converter Output Current Limiter Using Multiplier [53]

This current limiting implementation connects to and controls a DC-DC converter's input, feedback, current sense and output nodes. It has a very small PCB footprint and dissipates very little power. However, it ultimately did not function as desired because the Enphase micro-inverter kept increasing its input current level, ultimately de-regulating the DC-DC converter's output voltage. Pass transistor current limiting topologies may also function but typically require more PCB space and dissipate very large amounts of power, requiring very effective and potentially expensive cooling methods. Therefore, while such current limiting circuits (as well as crowbars) may be feasible for this project's SEPIC, they require additional modifications for functional designs that overcome the Enphase micro-inverter's hill-climbing MPPT algorithm. The

next section describes another further improvement for this project's SEPIC – reducing its PCB size.

11.2.2 SEPIC PCB Size Reduction

Reducing the SEPIC's PCB size can reduce its overall cost, as typically PCB manufacturers determine PCB cost by its surface area. Final production-level PCBs of this project's SEPIC can eliminate unnecessary footprints such as ones for Q_MAIN_ALT, or the additional footprint for adding another large electrolytic capacitor at the SEPIC's output. Such footprints were only necessary for prototyping and hardware testing and a final production-level PCB would not require them. Furthermore, this project's SEPIC's power path PCB copper planes can also be reduced in size such that they tolerate just the average required current levels and not the peak current levels. This project originally sized those planes for peak current tolerance levels from design calculations and computer simulations. Reducing the size of those copper planes reduces PCB size as well as plane capacitance. The next section describes a method for helping conserve elliptical trainer battery use when the elliptical trainer idles.

11.2.3 Elliptical Trainer Battery Conservation

One major flaw with this project's SEPIC is that the Precor elliptical trainer's onboard 12V battery still supplies power to both the SEPIC's LTC1871 PWM controller and LT4356-1 surge stopper IC even when the elliptical trainer idles with no user exercising on it. This drains the battery's charge and the battery supplies power to both ICs until it completely discharges. Discharging the elliptical trainer's battery results in extra energy required from the next elliptical trainer user (from exercise) for charging the

battery to its nominal 12V level. A method that helps eliminate this problem involves connecting the LTC1871 and LT4356-1's RUN and EN pins instead to the elliptical trainer's output (the SEPIC's input) rather than the battery terminals. Using this method, the battery only transfers power to the LTC1871 and LT4356-1 ICs if the elliptical trainer outputs its minimum threshold voltage – in the case of this project, that threshold is 5V. Because each mentioned pin also has maximum voltage limits, the existing resistor voltage dividers at the mentioned pins may also require modification. Figure 11.4 shows such an example, from the designed DC-DC converter in [10].

Figure 11.4: Elliptical Trainer Battery Minimum Voltage Threshold Detection Circuit [10]

Implementing this modification requires modifying the SEPIC's PCB layout, however. Designing a new PCB layout for this project's SEPIC while using the same SEPIC components as this project suffices as a potential senior project. Implementing this modification, however, results in large potential energy savings for this project's EHFEM

system. The next section describes possible alternatives for the SEPIC's LT4356-1 input protection circuit.

11.2.4 Improved Input Protection Schemes

As mentioned in section 10.4, the Precor elliptical trainer outputs voltages exceeding 100V when it detects an open load at its output. This problem occurs during the Enphase micro-inverter's initialization phase. The LT4356-1 input protection circuit in this project's SEPIC does not adequately safeguard against that previously unforeseen problem and adds complexity as well as additional required PCB space for the converter, mainly because of its pass transistor (Q_OVP). Furthermore, any catastrophic event that destroys Q_MAIN destroys Q_OVP as well, because of the series connection path to ground between both transistors. This design flaw results in additional component replacement costs for the EHFEM system's owner when such an event occurs. This project's SEPIC's current PCB has an alternate connection point for the elliptical trainer's output, located at VIN_ALT in figures 9.6 and 9.9. This connection point bypasses the LT4356-1 input protection circuit. However, while bypassing the LT4356-1 input protection circuit eliminates the transistor destruction problem, the SEPIC still requires inrush current limiting at start up. An additional soft-start circuit for the LTC1871 controller would eliminate that problem, but does not address the mentioned overvoltage problem. Therefore, this project's SEPIC necessitates more efficient input protection schemes that do not utilize pass transistors. Metal oxide varistors (MOVs) and transient voltage suppression (TVS) diodes at the SEPIC's input can help alleviate overvoltage and overcurrent problems, however, they must tolerate high voltage and

current levels during a fault. Future EHFEM projects can research this topic in greater detail for improving this project's SEPIC design. Section 11.3 discusses future projects that can further improve this project's SEPIC design, previously designed DC-DC converters for the EHFEM project as well the whole EHFEM system.

11.3 Future EHFEM Projects

This section discusses future projects that can further improve this project's SEPIC design, previously designed DC-DC converters for the EHFEM project as well the whole EHFEM system. These projects include SEPIC inductor design, control schemes for the Enphase M175-24-240 micro-inverter and grid-tie inverter design.

11.3.1 SEPIC Inductor Design

Along with PCB size reduction, using a single magnetically coupled inductor for housing L_1 and L_2 in this project's SEPIC may potentially reduce the SEPIC's cost. Furthermore, using a coupled inductor can also reduce this project's SEPIC's physical PCB size. However, using a coupled inductor with this project's SEPIC requires a custom inductor design because as mentioned earlier in this report, this project's author could not find a suitable commercially available coupled inductor or switching transformer that satisfied L_1 and L_2 's saturation current rating requirements. A custom coupled inductor requires using commercially available magnetic cores, which are typically expensive. However, using a custom, coupled inductor that produces the required inductance at 100 kHz switching frequency may result in greatly improved efficiency, offsetting the magnetic core costs while maintaining a 10 year system payback period. One possible source of efficiency loss in this project's SEPIC may have resulted from this project's

author's failure in recognizing that the $60\mu\text{H}$ inductance and $85\text{m}\Omega$ ESR values stated in Vishay's IHV28BZ60 datasheet was the nominal L_1 and L_2 inductance value at 1 kHz frequency (and not this project's SEPIC's 100 kHz switching frequency) [36]. After noticing that caveat, this project's author then measured L_1 's inductance and ESR at 1 kHz and 100 kHz frequencies using a GW Instek LCR meter. This project's author did not measure L_2 's inductance because L_2 is the exact same physical inductor as L_1 . Figures 11.5 and 11.6 show L_1 's inductance and ESR at 1 kHz frequency, while figures 11.7 and 11.8 show L_1 's inductance and ESR at 100 kHz frequency.

Figure 11.5: L_1 Measured Inductance at 1 kHz

Figure 11.6: L_1 Measured ESR at 1 kHz

Figure 11.7: L_1 Measured Inductance at 100 kHz

Figure 11.8: L_1 Measured ESR at 100 kHz

Figures 11.6 to 11.8 show that as the LCR meter's test frequency increases changes from 1 kHz to 100 kHz, L_1 's inductance drops from 60 μ H to 18 μ H, and its ESR increases to 4.5 Ω . This inductance drop and ESR increase at 100 kHz switching frequency in both L_1 and L_2 may have caused the majority of the efficiency losses in this project's SEPIC. The 18 μ H inductance value is lower than the SEPIC's required critical inductance for both L_1 and L_2 , and may have caused either DCM operation at lower input voltages or inductor saturation, resulting in greater stresses across Q_{MAIN} and those stresses across Q_{MAIN} in turn may have contributed to the SEPIC's efficiency loss. Vishay Siliconix manufactures similar inductors to the IHV28BZ60 used in this project's SEPIC, with higher inductance values and saturation current ratings. However, those inductors are more expensive than the IHV28BZ60. Therefore, a future EHFEM project could first re-simulate this project's SEPIC with the mentioned inductance and ESR values for L_1 and L_2 and note how much inductance drop and ESR increase at 100 kHz affects overall SEPIC performance. Afterwards, that project could then design a custom coupled

inductor that maintains 60 μ H and low ESR (close to the Vishay IHV28BZ60's specified 85m Ω ESR) at 100 kHz, for maintaining compatibility with this project's snubber. That project is suitable for students who have completed a magnetic component design course.

11.3.2 Inverter Control Mechanism

So far, each DC-DC converter from each EHFEM project (including this project's SEPIC) functioned properly as DC-DC converters. However, the Enphase micro-inverter's behavior is undesirable for EHFEM system operation. As mentioned earlier in this report, Enphase designed the M175-24-240 micro-inverter for operation with a photovoltaic panel input. This project's SEPIC behaves in the opposite manner from a photovoltaic panel; maintaining a 10 Ω input resistance requires its output current increasing with input voltage rather than outputting as much current as possible at any given input voltage. Therefore, a proper control mechanism for the Enphase micro-inverter is necessary for proper EHFEM system functionality and deployment. A proper control mechanism for the Enphase micro-inverter allows any chosen DC-DC converter to maintain a constant 10 Ω input resistance for keeping the Precor elliptical trainer's eddy current braking system intact, while regulating the DC-DC converter with the proper 36V output voltage. Maintaining a constant 10 Ω resistance at the DC-DC converter's input requires a controller limiting the Enphase micro-inverter's input current based on input voltage and power levels at the DC-DC converter. Figure 11.9 shows a potentially feasible control scheme, taken from [10].

Figure 11.9: Possible Control Scheme for Enphase M175-24-240 Micro-Inverter [10]

This control scheme only requires an external controller for varying the DC-DC converter's duty cycle as well as the inverter's phase shift angle relative to that of the electrical grid, based on the elliptical trainer's output power. Modifying the DC-DC converter's duty cycle and inverter's phase shift angle based on the elliptical trainer's output power should help limit the DC-DC converter's output current and help maintain 10Ω resistance at the DC-DC converter's input. This control scheme may or may not be costly for implementation. In any case, possible control mechanisms for the Enphase micro-inverter require further research and are beyond this project's purpose and scope.

11.3.3 Inverter Design

An alternative project to implementing an input current controller for this EHFEM system's existing Enphase M175-24-240 micro-inverter involves designing a custom grid-tie inverter that optimally takes in 36V DC voltage (within a small error range) and at least 8A current. However, the custom inverter would still require a mechanism for limiting its input current based on the DC-DC converter's input voltage. In other words, the custom inverter would need to behave in a similar fashion to a

variable current electronic load. This project may also produce an inverter design that costs more than designing an external controller for both a selected DC-DC converter and the Enphase micro-inverter as well as the Enphase micro-inverter itself, but further research in this area may turn up feasible designs that function with DC-DC converters requiring constant input resistance. In any case, proper inverter control mechanisms are the highest priority for future projects, as inverter control mechanisms are crucial for both EHFEM system functionality and deployment.

APPENDIX A: ĆUK CONVERTER DESIGN

Appendices A and B document initially selected DC-DC converter topologies and designs for this project that either did not function in computer simulations with the project requirements and specifications outlined in sections 1.1 to 1.3 or had exorbitantly high component costs that such designs would not produce a reasonable payback time period for the EHFEM system. In spring quarter 2010 this project began by selecting a non-inverting Ćuk DC-DC converter topology for its DC-DC converter design because of its voltage boost and buck capabilities and inherently good input and output current characteristics (i.e. it produces non-pulsating input and output current) because of inductors connected directly to its input and output nodes. However, finding a PWM controller that suited a non-inverting Ćuk DC-DC converter proved difficult because such a design required two switches and using two separate PWM controllers for switch synchronization added such difficulties. This project then explored variations of the buck-boost DC-DC converter for its design, culminating with an interleaved, multi-phase, non-inverting buck-boost converter design. However, that interleaved, multi-phase, non-inverting buck-boost converter design had a very high component cost (much greater than that of the SEPIC design used in this project) and its overall efficiency level could not justify its high cost. Simulating those mentioned designs that were unfeasible for this project led to selecting the SEPIC design used for this project. The following appendix subsections document the initial converter topology selection and design for this project, beginning with selecting an appropriate PWM controller for the initially selected Ćuk DC-DC converter design.

A1.1 PWM Controller Decision Matrix

Once this project's author drafted the initial specifications for this project's DC-DC converter, this project's author then made a decision on which PWM controller IC to use based on the input extremes set in the specifications. To select the PWM controller IC, this project's author made a decision matrix that weighs each of the following PWM controller ICs' characteristics:

- Input voltage range (the greater the range, the better, however it must support the 5-65V or 60V range required for this project converter; the 60V maximum suffices without the maximum voltage headroom specification)
- Current or voltage mode sensing (this project prefers current-mode sensing because of its ability for sensing both voltage and current to ensure that they both don't reach values out of the desired electrical ranges specified in sections 1.1 to 1.3)
- Turn-on threshold voltage
- Reference voltage
- Power dissipation (the lower the better)
- Duty cycle (must support a high enough duty cycle for stepping up an minimum input voltage of 8V to the nominal 36V that the Enphase micro-inverter requires)
- Switching frequency (higher range is better)
- Package size / number of pins (smaller package size is better)

This project then set absolute requirements for the PWM controller's supported duty cycle levels and switching frequencies. This project instantly disqualified any controller IC that failed one of these requirements:

- 81.82% duty cycle required for regulating 8V input to 36V output
- Switching frequency capability must be high to minimize ripple (≥ 50 kHz)

This project calculated the 81.82% duty cycle requirement based on the non-inverting Ćuk converter and SEPIC's voltage transfer function in continuous conduction mode operation, which (A1.1) describes:

$$\frac{V_{OUT}}{V_{IN}} = \frac{D}{1-D} \quad (A1.1)$$

D denotes the controller's switching duty cycle. This project then researched possible PWM controllers to use with the DC-DC converter:

- Leadtrend LD7576J/K
- ON Semiconductor NCP1219
- Fairchild SG6741
- Fairchild FAN6300A
- National LM5022
- National LM3429
- Maxim MAX5003
- National LM5025A
- National LM5020
- National LM3421
- National LM3423

This project then assigned a weight to each characteristic for determining a final rating (based on a 0-100% scale) for deciding which PWM controller IC to use:

- Input voltage range: 15%
- Sensing mode: 5%
- Turn-on threshold voltage: 20%
- Reference voltage: 5%
- Power dissipation: 10%
- Duty cycle: 20%
- Switching frequency: 10%
- Package size: 15%

The quantitative total adds to 100%. Tables A.A-1 through A.A-3 below show the decision matrix in tabular form. This project disqualifies and marks in pink any controller scoring 0% in any category (if it does not meet this project's absolute requirements). This project marks in yellow the two highest scoring controller ICs.

Table A.A-1: PWM Controller Decision Matrix – Characteristics of Each Controller IC

PWM Controller	Input Voltage Range (V)	Sense Mode	Min. Turn-On Voltage (V)	Reference Voltage (V)	Continuous Power Dissipation	Duty Cycle Range (%)	Switching Frequency Range	Package	# of Pins
Leadtrend LD7576J/K	-0.3 - 600	I	15	3	400-650 mW	0 - 75	94-106 kHz	DIP8 or SOP8	8
ON NCP1219	-0.3 - 500	I	11.2	2	920 mW	85 (Max.)	65 kHz or 100 kHz	SOIC7	7
Fairchild SG6741	0 - 500	I	15.5	2.3	400 mW	80 (Max.)	47-109 kHz	SOP8	8
Fairchild FAN6300A	0 - 500	I	15	2.5	400-800 mW	85 (Max.)	100 kHz (Max.)	DIP8 or SOP8	8
National LM5022	6 - 60	I	6.6	1.25	Internally Limited	0 - 95	170-1115 kHz	MSOP-10	10
National LM3429	4.5 - 75	I	6.3	1.235	Internally Limited	Programmed	2 MHz (Max.)	TSSOP14	14
MAXIM MAX5003	11 - 110	V	25	3	667-762 mW	75 (Max.)	200-1200 kHz	QSOP-16	16
National LM5025A	-0.3 - 105	V	7.3	5	Not Specified	0 - 80	180-650 kHz	TSSOP16	16
National LM5020	-0.3 - 100	I	7.4	1.25	Internally Limited	0 - 80	175-700 kHz	MSOP-10	10
National LM3421	4.5 - 75	I	6.3	1.235	Internally Limited	0 - 95	Programmable	TSSOP16	16
National LM3423	4.5 - 75	I	6.3	1.235	Internally Limited	0 - 95	Programmable	TSSOP20	20

Table A.A-2: PWM Controller Decision Matrix – Preference Ratings for Each Controller IC

Preference Ratings:								
PWM Controller	Input Voltage Range (V)	Sense Mode	Min. Turn-On Voltage (V)	Reference Voltage (V)	Continuous Power Dissipation	Duty Cycle Range (%)	Switching Frequency Range	Package
Leadtrend LD7576J/K	95%	95%	10%	90%	65%	0%	70%	95%
ON NCP1219	95%	95%	25%	90%	50%	95%	70%	95%
Fairchild SG6741	95%	95%	5%	90%	75%	70%	75%	95%
Fairchild FAN6300A	95%	95%	10%	90%	60%	95%	70%	95%
National LM5022	70%	95%	95%	95%	95%	95%	95%	85%
National LM3429	90%	95%	95%	95%	95%	95%	95%	50%
MAXIM MAX5003	40%	95%	0%	90%	55%	0%	95%	30%
National LM5025A	95%	70%	80%	85%	0%	70%	95%	30%
National LM5020	95%	95%	80%	95%	95%	70%	95%	90%
National LM3421	90%	95%	95%	95%	95%	95%	95%	30%
National LM3423	90%	95%	95%	95%	95%	95%	95%	20%

Table A.A-3: PWM Controller Decision Matrix – Weighted Final Preference Ratings for Each Controller IC

PWM Controller	Input Voltage Range (V)	Sense Mode	Min. Turn-On Voltage (V)	Reference Voltage (V)	Continuous Power Dissipation	Duty Cycle Range (%)	Switching Frequency Range	Package	Total Rating
ON NCP1219	14.25%	4.75%	5.00%	4.50%	5.00%	19.00%	7.00%	14.25%	73.75%
Fairchild SG6741	14.25%	4.75%	1.00%	4.50%	7.50%	14.00%	7.50%	14.25%	67.75%
Fairchild FAN6300A	14.25%	4.75%	2.00%	4.50%	6.00%	19.00%	7.00%	14.25%	71.75%
National LM5022	10.50%	4.75%	19.00%	4.75%	9.50%	19.00%	9.50%	12.75%	89.75%
National LM3429	13.50%	4.75%	19.00%	4.75%	9.50%	19.00%	9.50%	7.50%	87.50%
National LM5020	14.25%	4.75%	16.00%	4.75%	9.50%	14.00%	9.50%	13.50%	86.25%
National LM3421	13.50%	4.75%	19.00%	4.75%	9.50%	19.00%	9.50%	4.50%	84.50%
National LM3423	13.50%	4.75%	19.00%	4.75%	9.50%	19.00%	9.50%	3.00%	83.00%

From Table A.A-3, the two best PWM controller chips based on this project's characteristic selection matrix are the LM5022 and LM3429, both from National Semiconductor. This project disqualified Leadtrent L7576J/K and Maxim MAX5003 because they did not support the minimum duty cycle required for the converter, and this project disqualified the National LM5025A IC for National not specifying its power consumption. From this controller decision matrix this project then constructed in Intusoft ICAP a non-inverting Ćuk converter using the National LM5022 PWM controller IC.

A1.2 Non-Inverting Ćuk Converter Design using National LM5022 PWM Controller

Figure A-1 shows this project's initial non-inverting Ćuk converter schematic using the National LM5022 PWM controller.

Figure A-1: Non-Inverting Cuk Converter Using National LM5022 PWM Controller

Obtaining a non-inverted voltage (relative to the converter's input) at the converter's output requires an additional switch and diode for the converter. Each switch (power MOSFETs) requires an individual PWM controller IC, hence this implementation requires two National LM5022 ICs. The National LM5022 has a nominal 1.25V feedback voltage, and its feedback network influences the converter's output voltage. At the converter's output this project selected a feedback resistor network that would obtain the 36V that the converter requires (resistance values of 10k Ω and 260k Ω for the bottom and top feedback resistors). Resistor R_T also sets the PWM controller's switching frequency; this project set it to 34k Ω for obtaining a 500 kHz switching frequency, which is high

enough for minimizing any undesirable ripple at the converter's output. This project also selected a MOSFET and diode (International Rectifier's IRFM460 N-channel power MOSFET and the MBR20100 diode) that tolerated the wide input voltage range of the DC-DC converter. This project then simulated this converter using Intusoft ICAP (as the only simulation model for the LM5022 that National Semiconductor offered was for Intusoft ICAP/ISpice) using a 50ms simulation time. Figure A-2 shows the converter's performance with the PWM controller's 60V maximum input voltage with no load:

Figure A-2: Non-Inverting Cuk Converter Performance Characteristics at 60V Input, No Load, Using National LM5022 PWM Controller

VFB denotes the voltage across first PWM controller's feedback pin, VFB2 denotes the voltage across the second PWM controller's feedback pin, VSW denotes the converter's input MOSFET's drain voltage, VSW2 denotes the converter's output MOSFET's drain

voltage, I_{R1} denotes the current through the input inductor, and V_{OUT} denotes the converter's output voltage.

At 60V input and no load, not only does the converter not output the correct voltage, but both PWM controller feedback voltages also rise sharply above the 1.25V nominal value. The drain voltages across each MOSFET also rise to dangerous levels. In a real-world situation, this converter would be instantly destroyed. This project hypothesizes that the reason for this abnormal converter behavior may be the way that this project configured the two PWM controller ICs. This project next examines how the converter behaves with 60V input and an 8A full load current (by Ohm's Law, the required load resistance at the converter's output is 4.5Ω with 36V output). Figure A-3 shows the converter's performance in this situation:

Figure A-3: Non-Inverting Cuk Converter Performance Characteristics at 60V Input, Full Load, Using National LM5022 PWM Controller

Both PWM controllers' feedback voltages drop significantly below the nominal 1.25V and the PWM controllers fail to regulate. This is the opposite effect of what this project observed earlier with no load. Thus, the non-inverting Cuk converter design may not work properly with this particular PWM controller. The same results occur with 8V input.

From these preliminary simulation test results, the National LM5022 is not a feasible PWM controller for use with this project's DC-DC converter. Despite its wide input voltage range, this design has very poor line and load regulation, and these two characteristics are very crucial for the DC-DC converter functioning with the rest of the

elliptical system because of the elliptical machine's wide input voltage range and the Enphase micro-inverter's wide load range for DC-to-AC power conversion. Thus, after this first phase design process, this project concludes that a non-inverting, PWM switching Ćuk DC-DC converter is not feasible for functioning with the EHFEM system in this project. Appendix B describes a more potentially suitable DC-DC converter design for this project's EHFEM system using an interleaved, multi-stage non-inverting buck-boost topology.

APPENDIX B: INTERLEAVED BUCK-BOOST CONVERTER DESIGN

B1.1 Interleaved, Multi-Stage Buck-Boost Converter Design Using LT3845 and LTC3862 PWM Controllers

After unsuccessfully designing a DC-DC converter based on this project's original non-inverting, PWM switching Ćuk topology selection, this project designed a new converter based on an interleaved, multi-stage buck-boost topology. However, this project still retained its original DC-DC converter operation requirements and specifications from sections 1.1 to 1.3. Likewise the Ćuk converter topology, a basic buck-boost DC-DC converter outputs a DC voltage either greater or less than its input voltage, though with its polarity inverted with respect to its input voltage. Because of the buck-boost converter's inherently low power density, achieving greater stability under larger loads ($> 100\text{W}$) requires a multi-stage, multi-phase brute-force approach: this project designs a buck-boost converter with its input stage as a multi-phase, non-isolated PWM switching buck converter that steps down the input voltage. Its output stage is a multi-phase, non-isolated PWM switching boost converter that takes that stepped-down voltage and steps it up to the 36V that the Enphase micro-inverter requires at its input. This approach also results in a non-inverted output voltage polarity with respect to the converter's input voltage. This project initially began with a single phase design for each stage; however its simulation results show that using a single phase for each stage does not supply enough current needed for adequately driving the converter under high power loads. Hence, this project later moved to a multi-phase approach after receiving dismal results from its single-phase converter design. By adding more phases to each stage, the

converter increases its current-carrying capacity for driving high power loads at high switching frequencies. Finding a suitable PWM controller for implementing this design was perhaps the most difficult part of this design phase. The majority of the PWM controllers that this project's author researched could not tolerate the switching current (30A maximum) required for driving a 288W load for this design. However, this project's author eventually selected suitable PWM controllers for its input buck and output boost stages. For the buck stage design this project selected an LT3845A PWM controller from Linear Technology. For the boost stage design this project selected an LTC3862, also from Linear Technology. This project lists the LT3845A features below (based on the PWM controller selection matrix this project specified in appendix A1.1):

- 4V to 65V input voltage range
- Current-mode sensing
- 4V turn-on threshold
- 1.231V reference (feedback) voltage
- Internally limited power dissipation
- 0-90% duty cycle operation
- 100 kHz to 500 kHz switching frequency operation
- TSSOP-16 package

The 4V to 65V input range is suitable for this project's EHFEM system application, as this project's DC-DC converter specifications require a 5-65V input range. This project lists the LTC3862's controller's features below:

- -0.3V to 40V input voltage range
- Current-mode sensing
- 3.3V turn-on threshold
- 1.223V reference (feedback) voltage
- Internally limited power dissipation
- 96% maximum duty cycle operation
- 50 kHz to 650 kHz switching frequency operation
- TSSOP-24 or 24-lead QFN packages

For the output boost converter stage, the LTC3862's 40V input ceiling more than suffices. This project designs the buck converter stage such that it steps down the input to a much lower voltage than 40V. The next section describes this buck-boost converter's design process.

B1.2 Single-Phase Input and Output Stage Implementation

Using the two selected PWM controller ICs, this project initially designed a multi-stage, single-phase buck boost converter. Figure B-1 shows its schematic.

Figure B-1: Single-Phase, Multi-Stage Buck-Boost Converter Design

This converter is essentially a buck converter and boost converter connected in cascade. For the purpose of all testing in this design phase, this project uses 12V as the voltage that the input buck converter stage steps the input voltage down to. This means that this converter cannot use the full input voltage range down to 8V; however, using 12V as the buck stage's output provides this project with ideas for modifications in its next design phases. Section 3.2 indicates that any power generated from the Precor EFX 546i elliptical trainer is insignificant for any generated output voltage below 12V; however this project aims its final design at using its original specified input voltage range from section 1.2. The input stage in the schematic in figure B-1 has a 12V output voltage, rated at 6.25A (per phase). The input stage runs at a 300 kHz switching frequency while the output stage runs at a 500 kHz switching frequency. Selecting components required for the input buck converter stage requires the following parameters:

V_{IN} = Buck stage input voltage

V_{OUT} = Buck stage output voltage = 12V

V_{OR} = Maximum amount of desired output voltage ripple = 10%

D = Duty cycle

f_s = Buck stage switching frequency = 300 kHz

I_{OUT} = Buck stage average output current = 6.25A (maximum)

C_O = Output filter capacitor

L_C = Minimum inductor size required to maintain constant conduction mode (CCM)

V_{SW} = Switching MOSFET drain – source voltage rating

I_D = Switching MOSFET average drain current rating

V_R = Diode maximum reverse voltage rating

I_F = Diode maximum average forward current rating

A buck converter's voltage transfer function determines its duty cycle. Equation (B1.1)

shows the buck converter's transfer function.

$$D = \frac{V_{OUT}}{V_{IN}} \quad (B1.1)$$

V_{IN} ranges from 12 to 65V. However, a 12V input requires 100% duty cycle operation, and the LT3845A can only handle 90% maximum duty cycle. Thus the minimum input voltage required for regulating the buck converter stage is 13.3V. For a 13.3-65V input range, this translates to a 5.42% to 90% duty cycle range for the input buck converter stage. Equation (B1.2) determines the buck stage's output filter capacitor (C_O) size.

$$C_O = 65 \times 10^{-6} \left(\frac{0.2I_{OUT}}{V_{OR}} \right) = 65 \times 10^{-6} \left(\frac{0.2(6.25A)}{0.10} \right) = 813\mu F \quad (B1.2)$$

V_{OR} denotes the converter's output ripple voltage. Because the boost converter output stage regulates the voltage outputted by the buck converter input stage, it doesn't require a completely clean input voltage, just as long as it is high enough (at least 5V minimum) in order for the converter to output 36V. This project arbitrarily selected 100 μ F as a baseline for the purpose of testing. Equation (B1.3) determines this buck stage's required inductor size, using duty cycle and input voltage parameters. For the duty cycle parameter, this project assumes the 90% maximum condition, and for the input voltage parameter, this project assumes the 65V maximum condition.

$$L_C = \frac{(V_{IN}-V_{OUT})D}{2I_{OUT}f_s} = \frac{(65-12)0.9}{2(6.25)(300 \text{ kHz})} = 12.7\mu H \quad (B1.3)$$

f_s denotes the input buck stage's switching frequency. This project selects 40 μ H for allowing for extra headroom in case the output stage requires more current than expected on the physical converter design. Equations (B1.4) and (B1.5) determine the buck stage's switching MOSFET drain-source voltage and drain current ratings.

$$V_{DS} = V_{IN}(\text{maximum}) = 65V \quad (B1.4)$$

$$I_D = I_{OUT} = 6.25A \quad (B1.5)$$

For testing this project used a MOSFET with a 60V maximum V_{DS} rating (as 60V is the absolute maximum that any typical Precor EFX 546i elliptical trainer user generates according to [1]). Later designs used MOSFETs with higher V_{DS} ratings to account for the extra 5V headroom. Equations (B1.6) and (B1.7) determine the buck stage's diode ratings.

$$V_R = V_{IN}(\text{maximum}) = 65V \quad (\text{B1.6})$$

$$I_F = I_{OUT} = 6.25A \quad (\text{B1.7})$$

This project's author later realized that the selected diode for the input buck stage had only a 1A average forward current rating. Later designs the MBR20100CT Schottky diode, which tolerates up to 10A forward average current.

Likewise with the input buck converter stage, selecting the components required for the output boost converter stage requires the following parameters:

V_{IN} = Boost stage input voltage

V_{OUT} = Boost stage output voltage = 36V

V_{OR} = Maximum amount of desired output voltage ripple = 10%

D = Duty cycle

f_s = Boost stage switching frequency = 500 kHz

I_{OUT} = Boost stage average output current = 8A (maximum)

P_{OUT} = Boost stage output power = 288W (maximum)

C_O = Output filter capacitor

L_C

= Minimum inductor size required to maintain constant conduction mode (CCM)

V_{SW} = Switching MOSFET drain – source voltage rating

I_D = Switching MOSFET average drain current rating

V_R = Diode maximum reverse voltage rating

I_F = Diode maximum average forward current rating

A boost converter's voltage transfer function determines its duty cycle. Equation (B1.8) determines the boost stage's duty cycle.

$$D = 1 - \frac{V_{IN}}{V_{OUT}} \quad (\text{B1.8})$$

V_{IN} ideally is 12V. However, the filter capacitor this project selected for the buck input stage can cause V_{IN} shifting as low as 5V. Even then, that 5V minimum input voltage translates to 86% duty cycle operation, and the LTC3862 PWM controller can handle up to 96% duty cycle operation. For a 5-12V input range to the boost converter stage, the boost stage's duty cycle ranges from 67% to 86%. Equation (B1.9) determines the boost stage's output filter capacitor size, assuming the maximum 86% duty cycle operation.

$$C_O = \frac{DV_{OUT}}{\left(\frac{V_{OUT}}{I_{OUT}}\right)f_s(V_{OUT})(V_{OR})} = \frac{(0.86)(36)}{\left(\frac{36}{8}\right)(500 \text{ kHz})(36)(0.10)} = 3.82\mu F \quad (B1.9)$$

In reality, the minimum capacitance value is usually not enough for regulating the converter at the desired voltage under the maximum load condition. For this output stage, this project requires a clean output voltage for ensuring that the Enphase micro-inverter receives its desired optimal input voltage. Therefore this project boosted the output filter capacitor value from its calculated minimum to a reasonable 100 μ F. Equation (B1.10) determines the boost stage's inductor size, assuming 67% duty cycle operation for obtaining the largest possible minimum value.

$$L_C = \frac{D(1-D)^2\left(\frac{V_{OUT}}{I_{OUT}}\right)}{2f_s} = \frac{0.67(1-0.67)^2\left(\frac{36}{8}\right)}{2(500 \text{ kHz})} = 0.33\mu H \quad (B1.10)$$

This project selects 33 μ H (which is 100 times the minimum calculated value) for extra headroom in case the physical converter's output stage requires more current than expected. Equations (B1.11) and (B1.12) determine the boost stage's switching MOSFET drain-source voltage and drain current ratings.

$$V_{DS} = V_{OUT}(\text{maximum}) = 36V \quad (B1.11)$$

$$I_D = \frac{2P_{OUT}}{V_{IN}(\text{minimum})} = \frac{2(288)}{5} = 115.2A \quad (B1.12)$$

The output stage requires power MOSFETs that can tolerate >120A. For simulation this project uses a MOSFET with a 60V maximum V_{DS} rating. Equations (B1.13) and (B1.14) determine the boost stage's diode ratings.

$$V_R = V_{IN}(maximum) = 12V \quad (B1.13)$$

$$I_F = I_{OUT} = 8A \quad (B1.14)$$

This project's author realized the mistake during testing that the selected diode used for the input buck stage only had a 1A average forward current rating. Later designs use the MBR20100CT and MBRB2545CT Schottky diodes, which can tolerate up to 10A and 25A forward average current.

This design is now ready for simulation. Figure B-2 shows a 3ms simulation operation of the designed single-phase buck-boost design using 60V input and a 1A load. The green waveform denotes boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-2: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 1A Load

The converter appears to regulate the required 36V output without too much problem.

This project then boosts the load to 2A and simulates the converter design. Figure B-3 shows these results. The green waveform denotes the boost stage's output voltage $V(out)$ and blue denotes the buck stage's output voltage $V(n002)$.

Figure B-3: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 2A Load

These results show a noticeable load regulation problem for this converter, as the output voltage begins dropping to 25V (which is below the minimum input operating voltage required for the Enphase micro-inverter). Figure B-4 shows what happens when the simulation increases the load current to 4A.

Figure B-4: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 4A Load

The load regulation problem worsens. Figure B-5 shows converter operation with a 5A load.

Figure B-5: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 5A Load

Figure B-6 shows converter operation with the maximum 8A load.

Figure B-6: Single-Phase Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load

From these simulation results, the converter has an obvious load regulation problem. The input buck stage may not be providing enough average current under higher load conditions to the output boost stage for the boost stage to properly regulate the required 36V. Furthermore this load regulation problem can also be attributed to the switching frequencies on both stages not being exactly the same as each other (causing each PWM controller to operate out of sync) as well as using components that are not rated for this converter's ideal maximum values. Using a multi-phase design for the converter's input buck stage can mitigate the input current supply problem. The next section describes turning this single-phase cascaded buck-boost converter design into a multi-phase implementation.

B1.3 Two-Phase Input Stage and Single-Phase Output Stage Implementation

This project attempts improving the design from appendix B1.2 by adding another phase to both the input buck stage, while maintaining the same components as with the previous design. Adding another phase to the input buck stage doubles its current-carrying capacity, now to 12.5A. A multi-phase design further requires an additional LTC6908-1 external oscillator (also available from Linear Technology) tied to the SYNC pin of each PWM controller for every two phases for properly synchronizing each controller. Because LTSpice's LT3845A model does not contain a SYNC pin, the LTC6908 is not shown for any of the design schematics. However, it is necessary for the final physical converter design. The input stage still outputs 12V to the boost-converter output stage. Furthermore, the component ratings remain the same for each phase added, hence this project uses the same components for each multi-phase design iteration. Figure B-7 shows the first multi-phase design realization, a two-phase buck, single-phase boost converter.

Figure B-7: Two-Phase Buck, Single-Phase Boost, Multi-Stage Buck-Boost Converter Design

This project then tests this design using the maximum nominal stress condition with 60V input and an 8A load. Figure B-8 shows the converter's operation for a 30ms simulation time window. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-8: Two-Phase Input, Single-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load

Strangely enough, the boost stage's maximum output voltage only reaches 20V, which is not enough for the Enphase micro-inverter to properly regulate and deliver power. Figure B-9 below shows what happens if the simulation decreases the load to 6A, using the same 30ms time simulation window.

Figure B-9: Two-Phase Input, Single-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 6A Load

The result looks better, as the converter outputs 27V – enough for the Enphase micro-inverter to properly deliver power. However, it is still far from the ideal 36V that allows for maximum efficiency from the Enphase micro-inverter. Figure B-10 shows the converter's operation with a 4.75A load over the same 30ms simulation period (This project selected 4.75A as an intermediate point between 4 and 6A, with 4A being the midpoint relative to the 8A full load condition).

Figure B-10: Two-Phase Input, Single-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 4.75A Load

Now the converter reaches 33V output. However, this project desires that the converter properly regulates 36V output under the maximum stress condition (60V input, 8A load). This two-phase input design, however, shows much better results than with the previous purely single-phase design. The next section describes a further improved iteration of this design using four phases for the input buck stage.

B1.4 Four-Phase Input Stage and Single-Phase Output Stage Implementation

This project now determines how a 4-phase input design would theoretically perform (theoretical current carrying capacity of the input stage then becomes 20A). Again, this project uses the same components as with the previous designs. This project adds two more phases to the input (while maintaining the same number of phases at the output), resulting in a converter with a 4-phase buck input and single-phase boost output. Figure B-11 shows this design.

Figure B-11: Four-Phase Buck, Single-Phase Boost, Multi-Stage Buck-Boost Converter Design

This project then tests this design using the nominal maximum stress condition with 60V input and an 8A load. Figure B-12 shows the converter's operation from start-up to steady-state. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-12: Four-Phase input, Single-Phase Output Buck-Boost converter Individual Stage Output Voltages at 60V Input, 8A Load

The results are nearly identical to that of the previous 2-phase-input design. It is possible that the output stage reaches a current limit because of the increased number of phases at the input stage in this design and the output stage cannot handle the current fed in by the input stage in order to properly regulate its output. This project further enhances this design by adding an additional phase for the output boost converter for greater current carrying capacity for this stage. This results in a converter with a four-phase buck input and two-phase boost output. Figure B-13 shows this design.

Figure B-13: Four-Phase Buck, Two-Phase Boost, Multi-stage Buck-Boost Converter Design

This project then tests this design using the nominal maximum stress condition with 60V input and an 8A load. Figure B-14 shows the converter's operation from start-up to steady-state. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-14: Four-Phase Input, Two-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load

The results are greatly improved from previous designs, although in steady state the voltage swings between 33 and 37V, which is quite a huge ripple (approximately 11%, though this is under the maximum stress condition with a 60V input and 8A load), but this converter's output voltage reaches the optimal level, and stays within the voltage range that the Enphase micro-inverter desires. The efficiency for this time window is roughly 72.6% (Using LTSpice's math functions this project measured a 25.6V average output voltage with 8A average output current, along with 60V average input voltage with 4.7A average input current). The only major problem with this design is the large

number of components it uses (as well as the large physical space that it would consume). Because of the large number of components that this design utilizes, this design may end up too costly and take up too much space, considering the large number of inductors and MOSFETs in the system (which comprise most of the PCB's real estate), despite meeting almost all of this project's original required specifications. Perhaps this project can use a more simplified design instead, with 2 phases at both the input and output stages. The next section describes a such a 2x2 converter design.

B1.5 Two-Phase Input and Output Stage Implementation

Although reducing the number of phases on the input stage theoretically reduces the input stage's maximum current-carrying capacity, this project now analyzes how maintaining a large current-carrying capacity on the output stage affects converter performance. Figure B-15 shows the simplified, two-phase buck input and two-phase boost output design.

Figure B-15: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

This project now tests this design using the nominal maximum stress condition with 60V input and an 8A load. Figure B-16 shows the converter's operation from start-up to steady-state. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-16: Two-Phase Input, Two-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load

The ripple at the output is very high and undesirable, and overall performance is worse than that of this project's previous two-phase-input, single-phase-output design.

This project now addresses one crucial design flaw mentioned earlier – the asynchronous switching frequencies for both the input and output stages. This project then revises the two-phase design described in this section, which finally corrects the asynchronous frequency problem by increasing the switching frequency of the input stage to 500 kHz, as well as changing the inductor and filter capacitor values at each stage (in order to account for this switching frequency increase). Increasing the switching

frequency to 500 kHz on each stage only required changing the switching frequency resistor on the LT3845A PWM controller to 29.2k Ω . Now this project tests this revised two-phase design using the nominal maximum stress condition with 60V input and an 8A load. Figure B-17 shows the revised converter's operation from start-up to steady-state. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-17: Two-Phase Input, Two-Phase Output Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load, Using 500 kHz Switching Frequencies for Each Stage

This result is greatly improved from previous designs, and now the output reaches the 36V optimal voltage for the Enphase micro-inverter. Output ripple voltage is approximately 0.75V, which is about 2.08% with respect to the 36V output voltage. The output ripple shown in figure B-17 is more desirable than that in this project's original four-phase-input, two-phase-output design from appendix B1.4 because of the higher and synchronous switching frequency for both stages. Average output voltage is 36.236V and

average output current is 8A. Average input voltage is 60V and average input current is 6.14A. This translates to 78.6% efficiency. However, there is still noticeable oscillation at the input stage, though again, this project does not require a completely clean buck input stage output voltage. This converter design is feasible for this project's purpose, but the next section attempts shrinking the design further in terms of component count and physical space.

B1.6 Single-Phase Input and Output Stage Implementation at 500 kHz Switching Frequency

Figure B-18 shows this project's pure single-phase converter design, but using a 500 kHz switching frequency for both the input and output stages.

Figure B-18: Single-Phase, Multi-Stage Buck-Boost Converter Design with 500 kHz Switching Frequency for Each Stage

This project now tests this design using the nominal maximum stress condition with 60V input and an 8A load. Figure B-19 shows the converter's operation from start-up to steady-state. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-19: Single-Phase, Multi-Stage Buck-Boost Converter Individual Stage Output Voltages at 60V Input, 8A Load, Using 500 kHz Switching Frequencies for Each Stage

Even when using the same switching frequencies for both the input and output stages, the output stage still cannot draw the current required that the input stage requires under maximum load. The final output voltage hovers between 4.5V to 7V, with heavy ripple. Therefore, using a purely single-phase design for this converter is completely unfeasible. The next section describes using a single-phase input stage and two-phase output stage variation of this design.

B1.7 Single-Phase Input Stage and Two-Phase Output Stage Implementation at 500 kHz Switching Frequency

This project now determines if a single-phase-input, two-phase-output design can produce close to its desired electrical values before falling back to the pure two-phase design. Figure B-20 shows this design, again using the same components as with previous designs and a 500 kHz switching frequency for each stage.

Figure B-20: Single-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

This project now tests this design using the nominal maximum stress condition with 60V input and an 8A load. Figure B-21 shows the converter's operation from start-up to steady-state. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-21: Single-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Output Voltage with 60V Input and 8A Load

The results are very similar to the purely single-phase design. The input stage isn't pushing enough current to the output stage for regulating the proper output power level. Thus for this project uses the dual-phase configuration from figure B-15. In the next section this project attempts using a lower switching frequency for the dual-phase converter from figure B-15 for determining whether such a change improves overall converter operation and efficiency.

B1.8 Two-Phase Input and Output Stage Implementation Analysis at 500 kHz and 350 kHz Switching Frequencies

This project now examines the individual inductor currents of the dual-phase converter configuration and make sure that the converter itself is not operating in discontinuous conduction mode (DCM). Figure B-22 shows inductor L_1 's current (from the input stage, denoted by the red waveform) under maximum input voltage and load. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-22: Inductor Current of L_1 (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

L_1 's current reaches 0A, but not for a significant time interval. Thus the converter operates in BCM. While this condition isn't ideal, it does still allow the converter to properly regulate its output, though it operates dangerously close to DCM. The inductor required for the physical converter requires 30A current-carrying capacity. Figure B-23

shows inductor L_4 's (located in the buck stage's second phase, denoted by the red waveform) current under maximum input and load. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-23: Inductor Current of L_4 (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

The result is identical to that of L_1 's. These results show that each phase operates synchronously with the other. Now this project examines the inductor currents at the output boost stage. This project first examines L_2 , which is the output stage's primary (high-side) inductor. Figure B-24 shows L_2 's current under maximum input and load (denoted by the red waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-24: Inductor Current of L_2 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

No problems occur with L_2 , as it operates in CCM (L_2 's current never reaches 0A for any time period). Next this project examines L_3 , which is the output stage's secondary (low-side) inductor. Figure B-25 shows L_3 's current under maximum input and load (denoted by the red waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-25: Inductor Current of L_3 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

L_3 's overall current is much lower than L_2 's, hence the more noticeable ripple with respect to the current scale that this simulation uses. Like L_2 , its current never reaches 0A, assuring that converter still operates in CCM. Next this project examines L_7 , which is the output stage's primary (high-side) inductor located in the output stage's second phase. Figure B-26 shows L_7 's current under maximum input and load (denoted by the red waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-26: Inductor Current of L_3 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

L_7 's current is identical to that of L_2 's, and does not stay at 0A for any significant time period. This project next examines L_8 , which is the output stage's secondary (low-side) inductor located in the output stage's second phase. Figure B-27 shows L_8 's current under maximum input and load (denoted by the red waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-27: Inductor Current of L_8 (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

L_8 's current is identical to that of L_3 's; because of the simulation's current scale there is more noticeable ripple than with the current through the high-side inductors. Like all of the other inductors, its current never reaches 0A, assuring that the converter still operates in CCM. Furthermore, because the currents through each high-side and low-side inductor are identical to their respective counterparts, each phase in the output boost stage operates synchronously.

This project next examines the drain currents through each switching MOSFET and ensures that they are within a reasonable operation level. This project first examines Q_1 's (the primary, or high-side MOSFET in the buck stage's first phase) drain current. Figure B-28 shows Q_1 's drain current under maximum input and load (denoted by the red waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-28: Q_1 Drain Current (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Q_1 's drain current reaches a 27.5A peak, and the average drain current for this simulation time window is 3.13A. This result is not unreasonable, however, this project must ensure that its converter MOSFETs withstand at least 30A for their drain currents. Overall average drain current for Q_1 , however, is lower than the 6.25A (per phase) drain current current rating that this project established when determining the buck stage's component ratings earlier. This project next examines Q_2 's drain current. Q_2 is the secondary (low-side) switching MOSFET located in the input buck stage's first phase. Figure B-29 shows Q_2 's drain current (in red) under maximum input and load (denoted by the red waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-29: Q_2 Drain Current (Input Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Q_2 's drain current ranges from 6A to -20.5A (total peak amplitude is 26.5A). Overall average drain current for this simulation time window is -11A. This result is still within reasonable operation levels, as MOSFETs that can handle average currents up to 20A exist on the commercial market. From the inductor current results, Q_3 's drain current characteristics are the same that of Q_1 's (as both are buck stage high-side MOSFETs, just located in different phases), and Q_6 's drain current characteristics are the same as that of Q_2 's (both of these are buck stage low-side MOSFETs located in different phases), because of the converter's synchronous nature for each individual phase. Thus this project does not require any drain current plots for Q_3 and Q_6 , as that data is redundant. However, this project now examines the drain currents for the MOSFETs in the output boost stage. This project begins its examination with with Q_4 , which is the high side MOSFET located in the output boost stage's primary phase. Figure B-30 shows Q_4 's

drain current (denoted by the red waveform) at maximum input voltage and load. The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-30: Q_4 Drain Current (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Q_4 's drain current reaches a 14A peak, and the average drain current for this simulation time window is 8.86A. The original converter design equations required for the MOSFETs in the output boost stage handling at least 112.5A, however, this simulation result is far from the original specified minimum drain current rating. The MOSFET switches at the desired 500 kHz frequency, which explains the proximity of each individual rise and fall in its drain current. This project now examines Q_5 , the low-side MOSFET located in the output boost stage's primary phase. Figure B-31 shows Q_5 's drain current (denoted by the red waveform) at maximum input voltage and load. The

green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-31: Q_5 Drain Current (Output Stage) of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Q_5 's drain current ranges from -1.8A to 2.2A (total peak amplitude is 4A). Overall average drain current for this simulation time window is 1.54A. As expected, the low-side MOSFET does not conduct as much current as its high-side counterpart. Again, because of the converter's synchronous nature for each phase, Q_{11} 's (the high-side MOSFET on the output boost stage's secondary phase) drain current characteristics is identical to that of Q_4 's, and Q_{12} 's (the low-side MOSFET on the output boost stage's secondary phase) drain current characteristics is identical to that of Q_5 's. Thus, showing any plots for Q_{11} and Q_{12} is redundant.

The next operation characteristic that this project investigates in this converter design is the power dissipation across each switching MOSFET. MOSFET switching

power losses are the primary source of inefficiency in most DC-DC converter topologies. Improving overall converter design to maximize overall converter efficiency requires this project investigating any major power losses imposed across the switching MOSFETs in its converter. Furthermore, investigating power losses imposed across switching MOSFETs allows for noticing any lethal power transients that may occur during MOSFET switching transitions. A real MOSFET does not switch instantaneously. For example, when a MOSFET turns on and begins conducting current, there is usually a short delay between when its drain-source voltage decays to zero and when current begins to flow through its drain. The drain current usually begins flowing long before the drain-source voltage decays to 0V. By Ohm's Law, the power imposed across the MOSFET during that switching transition would be a non-zero value, and an especially large value in this converter's design because of its high switching currents, as well as high input voltages. This project begins its investigation with Q_1 and Q_2 . Figure B-32 shows Q_1 's power dissipation (denoted by the red waveform) and Q_2 's power dissipation (denoted by the teal waveform) during start-up and steady-state under maximum input voltage and output load.

Figure B-32: Q_1 and Q_2 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

During steady-state, Q_1 's average power dissipation is 11.9W and Q_2 's average power dissipation is 1.49W. However, the transients across Q_1 reach 1.4kW. This shows that Q_1 does not switch quickly enough. Unless Q_1 tolerates such a large power level within a short time period, the high power transients would destroy it. Additionally, such high power transients may cause arcing across other components within the MOSFET's proximity. This project later devises a few solutions for mitigating this switching power transient problem. Again, because of this converter's synchronous nature for each individual phase, Q_3 's power dissipation characteristics are identical to that of Q_1 's and Q_6 's power dissipation characteristics are identical to that of Q_2 's. Therefore the total dissipated power across the MOSFETs at the input stage is 26.78W. This project now examines MOSFET power dissipation at the output stage. Figure B-33 shows Q_4 's power dissipation (denoted by the red waveform) during start-up and steady-state. The green

waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-33: Q_4 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Q_4 's average power dissipation in steady state is 3.23W; however, its power dissipation transients reach 500W. While it's not as lethal as Q_1 's power transients, this project nonetheless addresses this problem as not all MOSFETs can tolerate such high power levels (i.e. simultaneous high drain-source voltage and high current) for even a time period in the microsecond range. This project now examines the low-side MOSFET in the output stage. Figure B-34 shows Q_5 's power dissipation (denoted by the teal waveform). The green waveform denotes the boost stage's output voltage and blue denotes the buck stage's output voltage.

Figure B-34: Q_5 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Average power dissipation across Q_5 is 16.79W; much higher than that of its high-side counterpart. Q_{11} 's power dissipation characteristics are identical to that of Q_4 's and Q_{12} 's power dissipation characteristics are identical to that of Q_5 's. Therefore the total wasted power on the MOSFETs at the output stage is 40.04W. After summing together the power dissipation losses at both the input and output stages, the total power dissipated power on the MOSFETs in the converter is 66.82W, which is mostly responsible for the converter's inefficiency.

This project then made changes to the converter's components for mitigating the power dissipation problem (as well as the lethal power transient problem). This project changed all transistors in the input stage to Siliconix Si4980DYs. According to LTSpice, the Si4980DY's maximum drain-source voltage rating is 80V, its drain-source on-resistance is 0.095Ω and it has a 15nC gate charge. This project selected this transistor because if a strong athlete generates more than 60V at the input from the elliptical, this

project's design's previous transistor configuration would not handle such high drain-source voltages (previously, the transistors that this project used at the input stage only had a 60V drain-source voltage rating). Selecting the Si4980DY accounts for the original 5V input headroom problem. This project also changes the output boost stage transistors to Si4850EYs. According to LTSpice, the Si4850EY has a 60V drain-source voltage rating, a 0.031Ω drain-source on resistance and 18nC gate charge. This alone may not affect transistor power consumption, so as another improvement attempt this project finally addresses the diode forward current rating problem mentioned earlier in this appendix. This project then changes all diodes at the input stage to MBR20100CTs, which as mentioned earlier, has a 100V reverse breakdown voltage of 100V (plenty enough for both the input and output stages, like the previously-used MBRS1100s) and can tolerate 10A average forward current. Furthermore, this project changes the converter's output stage diodes to MBRB2545CTs, which have a 45V reverse breakdown voltage (much lower than that of the MBR20100CTs, however, this converter's output stage diodes require reverse voltage ratings for as high as V_{OUT} – which is 36V) and tolerates 25A average forward current. The schematic in figure B-35 reflects these applied changes.

Figure B-35: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost converter Design With Improved Diodes and MOSFETs

This project now examines how these applied changes affect the converter's output characteristics as well as transistor power consumption characteristics. The plot in figure B-36 displays the improved converter's output voltage (denoted by the green waveform) and Q_1 's power dissipation (denoted by the red waveform) in steady-state operation, under maximum input and load. The blue waveform denotes the buck stage's output voltage.

Figure B-36: Output Voltage and Q_1 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs

Q_1 's average power consumption during steady-state is 5.5W. Using diodes that can handle a higher average forward current helps sink the additional current that previously flowed through the transistors. Thus there is converter improvement in terms of average power dissipation (it is half of what it was before), but power transients across Q_1 still occur in the 1.5kW to 1.6kW range. V_{OUT} swings between 35.5 to 37V, denoting a 4.2% ripple relative to the converter's ideal 36V output. This ripple level is tolerable for this project's application. Average output voltage for this time window is 36.089V, average

output current is 8A, average input voltage is 60V and average input current is -5.8402A.

This results in an overall 82.4% system efficiency, which exceeds the 75% requirement specified in section 1.2.

The plot in figure B-37 shows a zoomed-in view of Q_1 's switching power transients.

Figure B-37: Zoomed-in View of Q_1 Power Dissipation Transients of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

The plot in figure B-38 shows the highest power transient width.

Figure B-38: Width of Largest Q_1 Power Dissipation Transient Under Maximum Input and Load

This transient's width is approximately (4.8004442 ms - 4.8004369 ms), or 7.3ns. The MOSFETs that this project selected tolerate 1.5 to 1.6kW in such a short time period as this, however, such high power transients still poses a danger to other components within the MOSFETs' physical proximity.

This project now investigates Q_1 's drain-source voltage and drain current under maximum input voltage and load. Figure B-39 shows Q_1 's drain source-voltage (denoted by the teal waveform) and drain current (denoted by the blue waveform) under steady-state operation.

Figure B-39: Q_1 Drain-Source Voltage and Drain Current Under Maximum Input and Load

Q_1 's average drain-source voltage during steady-state is 47.772V. Q_1 's average drain-source current during steady-state is 2.92A. These average values; however, do not provide an accurate average power dissipation representation across this transistor.

Instead, this project then investigates what occurs when Q_1 's drain current reaches its highest level (roughly 28.9A), that occurs during Q_1 's turn-on transition. Figure B-40 shows a zoomed in plot of Q_1 's turn-on transition (with the teal waveform denoting drain-source voltage and blue waveform denoting drain current).

Figure B-40: Q_1 Drain-Source Voltage and Drain Current at Turn-On Transition Under Maximum Input and Load

At Q_1 's drain current's highest point, Q_1 's drain-source voltage equals approximately 13.54V. This translates to a 391.3W power dissipation at that instant, when the MOSFET is under a transition state. The maximum power dissipation period (i.e. the largest power transient), however, occurs when the drain current and drain-source voltage plots intersect. This occurs when Q_1 's drain-source voltage is 59.02V and Q_1 's drain current is 25.29A. This translates to 1.493kW power dissipation at that very instant, which explains the ~1.5kW power transients visible from earlier figures. From the plot in figure B-40, this project also confirms that Q_1 does not switch instantaneously; there is approximately a 7.3ns delay before Q_1 's drain-source voltage fully decays to 0V when Q_1 turns on, which also explains the ~1.5kW power transient lasting approximately 7.3ns.

The plot in figure B-41 shows Q_1 's gate at the same instant (the green waveform denotes gate voltage and the red waveform denotes gate-source voltage, the teal waveform denotes drain-source voltage and the blue waveform denotes drain current).

Figure B-41: Q_1 Drain-Source Voltage, Gate Voltage, Gate-Source Voltage and Drain Current at Turn-On Transition Under Maximum Input and Load

Q_1 's gate-source voltage shows the MOSFET's turn-on transition in greater detail. From this plot, this project concludes that when operating at this maximum stress level, turning on transistor Q_1 results in a large and potentially lethal power spike. The transistor's junction capacitance as well as drain-source resistance might be culprits for this delay, as those two items form a series RC circuit, in which both of those items affect the time constant for charging the capacitance portion. This project then devises two possible solutions for the switching power transient problem: either speed up transistor drain-source voltage decay at MOSFET turn-on (as well as slow down its rise during turn-off), or slow down transistor drain current rise at turn-on (as well as speed up its decay during

turn-off). However, there is a compromise involved for these two methods, because slowing down or speeding up the transistor's switching transition by too large of a margin results in worsening the power dissipation problem.

Figure B-42 shows Q_2 's power dissipation (denoted by the blue waveform) in steady state, using the new diodes and MOSFETs for the converter.

Figure B-42: Q_2 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs

Q_2 's average power dissipation in steady state is 4.039W, which is larger than before.

This increased power dissipation may have resulted from the new diode this project selected for Q_1 . Q_3 and Q_6 's power dissipation characteristics are similar to those of Q_1 and Q_2 . The total power dissipated across the transistors at the input stage is 19.078W.

Thus, using higher voltage and current-rated diodes and MOSFET reduces the total power dissipated across the MOSFETs at the input stage by roughly 7W.

This project now investigates MOSFET power dissipation at the boost output stage. Figure B-43 shows Q_4 's power dissipation (denoted by the blue waveform) in steady state, using the new diodes and MOSFETs for the converter.

Figure B-43: Q_4 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs

Strangely enough, Q_4 's average power dissipation increases to 5.922W. This project then determines if that power dissipation increase compensates for significantly reducing Q_5 's power dissipation. Figure B-44 shows Q_5 's power dissipation (denoted by the green waveform) in steady state, using the new diodes and MOSFETs for the converter.

Figure B-44: Q_5 Power Dissipation of Dual-Phase Buck-Boost Converter Under Maximum Input and Load, Using Improved Diodes and MOSFETs

Using higher-rated voltage and current diodes and MOSFETs reduces Q_5 's average power dissipation to 5.8W (from 16.79W with the previous converter design). This is an overall converter performance improvement, even if it slightly increases Q_4 's power dissipation. Q_{11} and Q_{12} 's power dissipation characteristics are similar to those of Q_4 and Q_5 . Thus, the total power dissipated across the transistors at the output stage is 23.444W. The total power that each transistor consumes in the entire converter is 42.522W, a 24W overall reduction from the previous converter design.

This project now determines if all the MOSFETs in the converter switch at the proper frequency (500 kHz). Figure B-45 shows a zoomed-in plot of Q_1 's gate switching trajectory.

Figure B-45: Q_1 Gate Voltage of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

This project measures the switching period of the middle rise-fall-rise transition cycle as 2.04μs, which corresponds with the 500kHz switching frequency selected for the input stage. This project now examines one of the output stage MOSFETs's switching trajectory. Figure B-46 shows a zoomed-in plot of Q_4 's gate switching rate.

Figure B-46: Q_4 Gate Voltage of Dual-Phase Buck-Boost Converter Under Maximum Input and Load

Again, this project the measures switching period of the middle rise-fall-rise transition cycle, and it equals $1.98\mu\text{s}$. This value corresponds with the 500kHz switching frequency selected for the output stage. From figures B-45 and B-46 this project concludes that its converter switching frequency resistors are the correct values and that the converter synchronizes both its input and output stages at the same switching frequency.

This project now address the lethal switching power transient problem. As mentioned earlier, mitigating the switching delay problem requires either speeding up the drain-source voltage decay at MOSFET turn-on (as well as slow down its rise during turn-off), or slow down the drain current rise at turn-on (as well as speed up its decay during turn-off). This project can use transistors with faster switching speeds, however switching delays always occur regardless of transistor switching speed and cause large power dissipation transients.

Engineers at National Semiconductor listed numerous MOSFET selection design notes on their website, which aim at reducing switching power losses (as well as reducing lethal power transients, which are partially responsible for those switching power losses).

Those design notes are as follows: [74]

- 1. Switching losses increase for larger gate and drain capacitance and these capacitances are inversely proportional to the on-resistance. FETs with the lowest on-resistance inevitably have the higher capacitance hindering HS (high side – primary FET) switching speed.*
- 2. Reducing the switching clock frequency reduces switching losses; that is, at lower frequencies the losses during on/off transitions become a diminishing proportion of the total on-time of the FET causing conduction losses to increasingly dominate.*
- 3. For higher input voltages relative to the output voltage the duty cycle of the HS FET decreases causing the switching losses to increasingly dominate.*
- 4. In order to further reduce conduction losses, multiple, parallel, LS (low side – synchronous or secondary FET) FETs are often employed. The number of parallel FETs is determined ultimately by cost, the gate driver's ability to drive them, and the point of diminishing returns.*

The engineer should be aware that in most POL applications, especially for input voltages higher than 12V, the switching losses will likely dominate all other losses. Under these circumstances the lowest overall losses in the HS FET are not necessarily achieved by using a device with the lowest on-resistance. The FET must be selected to minimize the sum of all the losses. The FET's on-resistance must be optimized at a higher value to achieve reduced capacitance and so reduce the switching losses. The major MOSFET vendors now provide "reduced charge, fast switching" MOSFETs which are optimized in this way for high-side buck applications.

If optimizing the FETs does not enable high enough efficiency in a system, the switching frequency can be reduced to decrease the switching losses and improve the efficiency. This, however, can result in a physically larger system.

The foregoing discussions have made clear that to achieve maximum efficiency in a high input voltage buck converter, the high side MOSFET must be carefully selected to minimize the sum of the switching and conduction losses.

From National Semiconductor's suggestions, this project can further reduce switching power losses by using MOSFETs with a higher on-state drain-source resistance (which tend to have lower junction capacitance, allowing for faster switching transitions). Furthermore, this project can also decrease the switching frequency, which in turn decreases switching losses, leaving only MOSFET conduction losses. However, employing multiple MOSFETs in parallel can further reduce conduction losses, though this project refrains from implementing such a design because this project desires to keep its converter design simple. Furthermore, the converter's PWM controllers may have synchronization problems while each driving more than two MOSFETs.

This project then applies National's MOSFET selection suggestions to its converter design. This process involves changing the converter's input stage and output stage MOSFETs once again. For the new input stage MOSFETs this project selects International Rectifier's IRF510. According to LTSpice the IRF510 has a 100V maximum drain-source voltage rating and a 0.54Ω on-state drain-source resistance. For the output MOSFETs this project selects Fairchild's HUFA76413. According to LTSpice the HUFA76413 has a 60V maximum drain-source voltage rating and a 0.056Ω on-state drain-source resistance. This project carefully selected these MOSFETs based on their on-state drain-source resistance value. If the on-state drain-source resistance of the output MOSFETs is too high, the converter may not regulate at the optimal 36V that this project requires because of drain current reductions involved with increasing such resistance. Furthermore, this project also reduces the switching frequency to 350 kHz on both stages

as a further attempt for mitigating the switching losses. Figure B-47 shows the mentioned changes to the converter design.

Figure B-47: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 350 kHz Switching Frequency and IRF510 and HUF76413 MOSFETs

This project then examines this new converter's performance. The plot in figure B-48 shows the new converter's output voltage (denoted by the green waveform), buck stage output voltage (denoted by the blue waveform), and power dissipated across Q_1 (denoted by the red waveform).

Figure B-48: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 350 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs

This result is still undesirable (there are still very lethal power transients across Q_1), and the switching power transients across Q_1 appear to be larger than those from previous designs. From these inherent power dissipation problems, this project deems this dual-phase buck-boost converter design unfeasible and in the next section it moves to a four-phase input variation of this design.

B1.9 Four-Phase Input Stage and Two-Phase Output Stage Implementation

This project now returns to the concept of adding more phases to each stage for increasing current-carrying capacity, and determines if doing so is beneficial for reducing

switching power losses and switching power transient magnitudes. For this new 4-phase converter design, this project reduced the switching frequency to 300 kHz as a method of decreasing switching losses. Figure B-49 shows this new four-phase-input buck-boost converter.

Figure B-49: Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 300 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs

This project then tests this design's performance. Again, this project uses the maximum stress condition using 60V input voltage with an 8A load. Figure B-50 shows this converter's output voltage (denoted by the green waveform), buck stage output voltage (denoted by the blue waveform) and Q_1 power dissipation (denoted by the red waveform) in start-up and steady-state operation modes.

Figure B-50: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 300 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs

Figure B-51 shows the same plot, but in converter steady-state operation.

Figure B-51: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 300 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs

For steady-state operation, the average power dissipated across Q_1 is 4.7982W. Average output voltage is 35.268V, average output current is 8A, average input voltage is 60V, and average input current is -5.7403A. This corresponds to an overall 81.9% system efficiency. The maximum power transient is 1.17kW. Likewise with the two-phase converter design from the previous section, lethal switching power transients still exist.

This project then further reduces switching frequency on both stages to 100 kHz, as an attempt at reducing the switching power transient magnitudes. The only modifications required for this switching frequency change are changing the buck input stage switching frequency resistors to 191k Ω and the boost output stage switching frequency resistors to 129.9k Ω . This project then measures power dissipated across Q_1 using this project's maximum stress condition. Figure B-52 shows in steady-state operation the output voltage (denoted by the green waveform), buck stage output voltage

(denoted by the blue waveform) and Q_1 power dissipation (denoted by the red waveform) for the new converter 4-phase design running at 100 kHz switching frequency.

Figure B-52: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and IRF510 and HUFA76413 MOSFETs

During steady-state operation, the average power dissipated across Q_1 is 3.9228W.

Average output voltage is 35.321V, average output current is 8A, average input voltage is 60V, and average input current is -5.412A. This corresponds to an overall 87% system efficiency, which suffices. However, from figure B-52, the maximum power transient is 1.055kW. This project then examines Q_1 's switching trajectory waveforms. Figure B-53 is a zoomed in switching trajectory plot showing Q_1 's drain-source voltage (in magenta) and drain current (in teal), as well as Q_1 's maximum switching power transient (in red) at the maximum power transient's instant.

Figure B-53: Q_1 Drain-Source Voltage, Gate Voltage, Power Dissipation and Drain Current at Turn-On Transition Under Maximum Input and Load

Q_1 's drain-source voltage still does not decay quickly enough as Q_1 turns on (when current begins flowing through its drain). The transient lasts approximately $0.923\mu\text{s}$. This project then examines other solutions for mitigating this power transient problem, as using MOSFETs with higher on-state drain-source resistance and decreasing the switching frequency alone doesn't eliminate it. A soft-switching technique may not function properly with this multi-phase design, and increasing gate-drain capacitance as well as gate resistance on Q_1 only make matters worse, as doing so results in the transistor spending more time in the switching region.

A possible viable solution for the switching power loss and switching power transient problems is implementing a snubber. A snubber is a discrete circuit designed for altering a MOSFET's switching trajectory and in the end reduce its switching power loss. By reducing the switching power loss, a snubber in turn reduces the overall power loss in

a DC-DC converter, thus improving converter efficiency. There are two types of snubbers: turn-on and turn-off snubbers. As their names suggest, they are designed to alter switching trajectory during MOSFET turn-on and turn-off transistions, respectively. A snubber may in fact be the solution that this project desires for reducing its converter power loss problems, as depending on their component values, they slow down current rise or speed up drain-source voltage decay at MOSFET turn-on, or speed up current decay or slow down drain-source voltage rise at turn-off.

Figure B-54 shows a basic turn-off snubber, and it consists of a resistor, capacitor and two diodes.

Figure B-54: Basic Turn-Off Snubber Circuit

Diode D_1 allows for a drain current flow path when the MOSFET turns off – in which case the MOSFET's drain current flows into the snubber circuit.

There are some design issues with using a snubber, however, according to [13]:

- A smaller snubber capacitor allows for a quicker drain-source voltage rise before its drain current decays to zero, while a larger snubber capacitor slows down drain-source voltage rise and decay (the latter is undesirable during MOSFET turn-on).
- Energy absorbed by the MOSFET is inversely proportional to the snubber capacitor size.

- The most desirable situation is having the drain current decay to zero before the snubber capacitor voltage rises.
- Increasing the snubber capacitor size results in greater power losses across the snubber circuit itself.

There are numerous compromises that this project must make when designing a snubber.

Designing a snubber requires carefully selecting component sizes because the fast-switching nature of this converter's MOSFETs results in snubber components sensitivity (especially if this project selects the incorrect component sizes) [13].

Figure B-55 shows a basic turn-on snubber, and consists of an inductor, a resistor and two diodes.

Figure B-55: Basic Turn-On Snubber Circuit

The inductor is a turn-on snubber's most crucial component. Because inductors oppose changes in current, connecting an inductor in series with a MOSFET allows for drain current rise slowdown during the MOSFET's turn-on transition.

In either case, adding a resistive snubber results in some power loss across the snubber (and hence, overall efficiency loss in the converter itself). However, as long as

this project sizes the snubber components appropriately the losses from the snubber are almost negligible, while at the same time provides the potential for greatly reducing overall power loss across the converter's switching MOSFETs.

This project then returns to the pure two-phase converter design from appendix B1.3 and determines if using both a turn-on and turn-off snubber for the buck input stage MOSFETs improves converter performance. Likewise with the most recent 4-phase design, this project also reduces the switching frequency of both stages to 100 kHz.

This project then designs a turn-off snubber for the buck stage for optimal operation (such that the MOSFET's drain-source voltage reaches its final value at the same time the drain current reaches zero, while at the same time minimizing power loss across the snubber resistor).

First, we this project defines the turn-off snubber parameters:

V_{IN} = Input Voltage

V_{OUT} = Output Voltage

I_{SW} = Maximum Switch (MOSFET) Drain – Source Current

f_s = Switching Frequency

D = Duty Cycle

t_f = I_{SW} fall time (from 90% to 10% of its maximum value)

C = Snubber Capacitor Value

C_{opt} = Optimal Snubber Capacitor Value

C_{dis} = Capacitor Discharge Time

R = Snubber Resistor Value

Equations (B1.15) and (B1.16) determines the buck stage's duty cycle under this project's maximum stress level.

$$V_{OUT} = DV_{IN} \quad (B1.15)$$

$$V_{OUT} = 12V, V_{IN} = 60V \quad (B1.16)$$

Therefore:

$$D = \frac{V_{OUT}}{V_{IN}} = \frac{12}{60} = 0.2, \text{ or } 20\% \quad (\text{B1.17})$$

For the converter operating under the maximum stress level:

$$V_{IN} = 60V, I_{SW} = 27A, f_s = 100kHz, D = 0.2, t_f = 5ns \text{ (taken from earlier plots)} \quad (\text{B1.18})$$

Equation (B1.19) calculates the turn-off snubber capacitor value:

$$C = \frac{I_{SW}t_f}{2V_S} = \frac{27(5ns)}{2(60)} = 1.125 \text{ nF} \quad (\text{B1.19})$$

Equation (B1.20) calculates the turn-off snubber resistor value assuming 5 time constants for the capacitor discharge time:

$$R = \frac{\left(\frac{D}{f_s}\right)}{C_{dis}C} = \frac{\left(\frac{0.2}{100kHz}\right)}{(5)(1.125nF)} = 355.556\Omega \quad (\text{B1.20})$$

This project rounds this resistor value up to 356Ω . Equation (B1.21) calculates the turn-off snubber's optimum capacitor value associated with the switching power losses:

$$C_{opt} = \frac{I_{SW}t_f}{\sqrt{12}V_S} = \frac{(27)(5ns)}{\sqrt{12}(60)} = 0.6495nF \quad (\text{B1.21})$$

This project rounds this capacitor value up to $0.65nF$.

This project now designs its converter turn-on snubber. As mentioned before, connecting an inductor in series with the transistor slows down the transistor's drain-source current during the turn-on transition. Because of this project's converter's MOSFETs' fast switching abilities, it is crucial that the turn-on snubber inductor isn't too large (keeping it in the nH range is optimal). This project also adds a parallel resistor to the inductor. Designing the turn-on snubber first requires knowing the time constant of an RL circuit, as (B1.22) denotes.

$$\tau = \frac{L}{R} \quad (\text{B1.22})$$

L denotes the inductance and R is the resistance. The inductor current in the snubber also must decay close to zero (but not exactly to zero) for the maximum snubber effectiveness during the next time cycle that the transistor turns on. However, as mentioned earlier the inductor value cannot be too large, otherwise the MOSFET's drain-source current rises and decays too slowly when the drain-source voltage increases, resulting in larger switching power losses. Equation (B1.23) begins the inductor selection process, and it assumes that the snubber requires 3 time constants for inductor discharge.

$$t_{off-state} > \frac{3L}{R} \quad (B1.23)$$

$t_{(off-state)}$ is the time period that the drain current is zero. This project determined $t_{off-state}$ as $1.2\mu s$ from earlier plots. This project selects $40nH$ (a reasonably small value, but not too small) for its converter's snubber inductor. From that inductance value, (B1.24) calculates the turn-on snubber resistor value.

$$R > \frac{3L}{t_{off-state}} = \frac{3(40nH)}{1.2\mu s} = 0.1\Omega \quad (B1.24)$$

Because the resistor value must be greater than 0.1Ω , this project selects 0.125Ω for the turn-on snubber resistor.

Now this project adds both its designed turn-on and turn-off snubbers to its converter. Figure B-56 shows the snubber additions to this project's earlier pure two-phase converter design.

Figure B-56: Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

This project uses Schottky diodes for the snubber diodes for minimizing snubber response time during MOSFET turn-on and turn-off transitions. This project also increased the input buck stage's current sense resistor to $6.8\text{m}\Omega$ (from $1\text{m}\Omega$ prior) and decreased the buck stage's output capacitor to $40\mu\text{F}$ (from $100\mu\text{F}$ prior) for further decreasing the maximum current allowed through the input buck stage (and hence decreasing overall drain current through the MOSFETs in the buck stage). This project does not require a stable buck stage's output voltage – just as long as it is high enough for the converter's output boost stage to regulate the proper 36V output.

This project now simulates the refreshed two-phase converter and checks the power dissipation across Q_1 . Figure B-57 shows the new two-phase converter's output voltage (in green), buck stage output voltage (in blue), and Q_1 power dissipation (in red).

Figure B-57: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz switching frequency and Turn-On and Turn-Off Snubbers

The buck stage's output voltage is noticeably noisier after adding the snubber (affected mostly by the output capacitor value reduction – as the output capacitor acts as a low pass filter). However, the buck stage's output voltage doesn't need to be completely clean, as the output boost stage regulates that buck voltage to a stable value. The maximum power transient further reduces in magnitude; it is now 340W. Other smaller transients still occur, however. The majority of the transients are closer to 120W. The converter itself still regulates at 36V overall average output. While these power transients aren't too lethal (it is certainly much improved from before), this project can still further improve on this design. Figure B-58 is a zoomed in plot of figure B-57 showing the converter in steady-state operation.

Figure B-58: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Average power dissipated across Q_1 for this time window is 12.66W, which is actually worse than the previous converter designs that did not employ a snubber. Figure B-59 is a plot that includes the converter's input current (denoted by the teal waveform).

Figure B-59: Output Voltage, Input Current and Q_1 Power Dissipation of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Overall, the maximum input current decreased with respect to the previous design. For this simulation time window, average output voltage is 36.053V, average output current is 8A, average input voltage is 60V, average input current is -5.94A. This corresponds to an 80.9% overall system efficiency. Figure B-60 is a plot showing Q_1 's switching trajectory during the highest measured power transient. The teal waveform denotes Q_1 's drain current while the magenta waveform denotes drain to source voltage. The overall maximum drain current through Q_1 during this entire simulation period is 15.6A.

Figure B-60: Q_1 Power Dissipation, Drain-Source Voltage and Drain Current of Two-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Making the power transient levels more acceptable requires severely reducing the current through each transistor. This project accomplishes this task by adding more phases to either the buck or boost stage. While browsing through the LTC3862 datasheet, this project's author noticed that one of the application notes had a simplified method for creating a two-phase boost converter (for our boost stage) while using only one PWM controller IC [75]. This project then adopted that design from the LTC3862 datasheet and modified it for withstanding a 12V input (from the converter's input buck stage), 36V output and an 8A load, while using the same transistors and diodes from the previous converter design. Figure B-61 shows this simplified two-phase output boost stage realization.

Figure B-61: Simplified Two-Phase Output Boost Converter Stage

This simplified design was certainly a breakthrough, as now this project can add more phases to the converter's input stage without adding too much complexity. This means that a 4-phase input, 2-phase output buck-boost converter may now be more physically feasible DC-DC converter design. Adding more phases to the input buck stage reduces the current through each transistor at each phase, thereby reducing the power transients occurring across those transistors. Now this project revisits its old 4-phase design. Figure B-62 shows this newly revised 4-phase input, 2-phase output design, with turn-on and turn-off snubbers still in place on the buck stage and switching frequency maintained at 100 kHz for both stages. For this design this project changed the current sense resistors on the boost stage to $6\text{m}\Omega$, further reducing the maximum current through each output stage MOSFET.

Figure B-62: Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

This project now examines the power dissipation across Q_1 . Figure B-63 is a plot showing the newly-redesigned converter's performance under our maximum stress condition (converter output voltage is in green, buck stage output voltage is in blue, and power dissipated across Q_1 is in red).

Figure B-63: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

The buck stage's output voltage is still noisy. However, for this converter design, that noise is tolerable as long as the voltage is high enough for the boost stage to regulate. The major improvement here is that the majority of Q_1 's switching power transients reduce to the 80W range, and is certainly a significant step forward from the 1.5kW transients from this project's original converter designs. Converter output voltage swings between 34.8 and 37.4V, which corresponds to a 7.2% ripple with respect to a 36V average output voltage, which suffices based on the specifications from section 1.2. Figure B-64 shows a zoomed-in plot of figure B-63 in steady-state converter operation:

Figure B-64: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

For this overall converter the maximum power transient occurs at 125.4W (during start-up), but once the converter reaches steady-state the majority of the power transients lie near 80W, +/- 5W. These transient values are acceptable for the selected transistors used in this design. Average power dissipation across Q_1 for this simulation time window (3 to 15ms) is 3.24W. The converter's average output voltage is 35.994V, average output current is 8A, average input voltage is 60V, and average input current -5.477A. This corresponds to an overall 87.6% system efficiency, which more than suffices against the 75% efficiency requirement outlined in section 1.2. Figure B-65 shows Q_1 's switching trajectory with drain current in teal and drain-source voltage in magenta. The red waveform denotes Q_1 power dissipation, the green waveform denotes converter output voltage and the blue waveform denotes the buck stage's output voltage.

Figure B-65: Q_1 Power Dissipation, Drain-Source Voltage and Drain Current of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Figure B-66 shows a zoomed-in plot of Q_1 's switching trajectory during the highest power transient occurrence in steady-state.

Figure B-66: Q_1 Switching Trajectory at Maximum Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design in Steady-State Operation with 100 kHz Switching Frequency and Turn-On and Turn-On Snubbers

The same transistor turn-on issue from previous designs still exists. However, for now, this power transient level is acceptable for the selected transistors in this converter.

For the next 4-phase converter revision this project reduces the turn-on snubber resistor to $5\text{m}\Omega$ and increases the snubber inductor to 75nH , for determining any possible benefits with slightly increasing the inductor size and reducing the resistor size. Figure B-67 shows the converter's performance in steady-state operation with these changes (with output voltage in green, buck stage output voltage in blue, and Q_1 power dissipation in red).

Figure B-67: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Overall, for this refreshed converter, Q_1 's average power dissipation in steady-state is 3.24W, which remains unchanged from the previous design. However, the majority of Q_1 's switching power transients now lie below 80W, and only one small 118.5W transient occurs during start-up. For any higher or lower inductor values, this result becomes worse. Thus it appears that 75nH is the optimal turn-on snubber inductor size under maximum converter stress conditions. Even with a small 3.24W dissipation across Q_1 , however, all converter MOSFETs require adequate heatsinking. This project investigates how much power dissipates in the snubbers. Figure B-68 below shows the power dissipated across the turn-off snubber (in teal), and across the turn-on snubber (in magenta) in steady-state. The green waveform denotes the converter's output voltage.

Figure B-68: Turn-On and Turn-Off Snubber Resistor Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

Average power dissipation across turn-on snubber resistor in steady-state is 4.24mW, and dissipation across each turn-off snubber resistor is 41.6mW. For all four phases, this corresponds to a 183.36mW (average) total power dissipation across the entire snubber implementation, which is still less than a 0.25W, and is a very reasonable sacrifice for reducing switching power transients across the converter's input buck stage MOSFETs. Figure B-69 shows the overall input current (from the voltage source, in which case is the Precor EFX 546i elliptical trainer) for the converter in steady state (in blue). The green waveform denotes the converter's output voltage.

Figure B-69: Input Current of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

The converter's average input voltage in steady-state is 60V, average input current is - 5.5A, average output voltage is 35.893V and average output current is 8A. This translates to an overall 87% system efficiency, which is the same efficiency level as that of the most recent design's. Therefore this project maximized the converter's snubber's utility for the buck input stage.

The next major step involves correcting power transient issues with the boost stage transistors. Figure B-70 shows the power dissipation across M_1 , the primary-phase boost stage transistor (in red) with the converter operating in steady-state. The green waveform denotes converter output voltage and the blue waveform denotes the buck stage's output voltage.

Figure B-70: Output Voltage, Boost Stage Output Voltage and M_1 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Despite the large power transients, the overall average power dissipation across M_1 (for this time window from 3 to 15ms) is only 3.45W. However, this project does not desire power transients exceeding 400W across its converter components. A similar situation occurs with M_2 (the secondary phase transistor on our boost stage) in the plot in figure B-71.

Figure B-71: Output Voltage, Boost Stage Output Voltage and M_2 Power Dissipation of Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Figure B-72 shows the output boost stage's operation characteristics in steady state, with a 60V input and 8A load. Converter output voltage is in light green, buck stage output voltage is in dark blue, inductor L_3 's current is in red, inductor L_2 's current is in teal, M_1 's drain current is in magenta, M_2 's drain current is in gray, diode D_3 's current is in dark green and diode D_4 's current is in dark blue.

Figure B-72: Output Voltage, Boost Stage Output Voltage, Inductor Currents, Diode Currents and M_1 and M_2 Drain Currents Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

The two transistors for the converter's boost stage operate out of phase with each other.

While both boost stage inductors operate in continuous conduction mode, their current values increase as their corresponding series-connected transistors turn on. L_2 's current increases when M_1 turns on, and L_3 's current increases when M_2 turns on, and each inductor's current decreases when their corresponding transistors turn off. Each diode also behaves in a similar manner. D_3 turns on when M_1 turns off, and D_4 turns on when M_2 turns off. The diodes provide a path to ground for the transistors' drain current when their corresponding transistors switch off.

The two boost stage transistors also must switch at the proper frequency.

Transistor M_1 's measured switching period (using the drain current plot from figure B-72) occurs from 3.14645ms to 3.1565ms, which corresponds to a 10 μ s period, which in turn corresponds to a 100 kHz switching frequency. Likewise, M_2 's switching period

occurs from 3.15148ms to 3.16153ms, which corresponds to a 10ms period and 100 kHz switching frequency. Thus, both boost stage transistors switch at the proper frequency.

A snubber is also necessary for the output boost stage. Figure B-73 shows M_1 's and M_2 's drain-source voltage turn-off trajectories, in red and teal. The green waveform denotes converter output voltage and the blue waveform denotes the buck stage's output voltage.

Figure B-73: Output Voltage, Boost Stage Output Voltage and M_1 and M_2 Drain-Source Voltages for Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

M_1 's measured drain-source turn-off time is 14.7ns, and M_2 's measured drain-source turn-off time is 14.3ns.

This project now designs the converter's boost output stage's snubber using the exact same methodology as with that of the converter's input buck stage, however, that method is slightly more difficult because the boost output stage already has inductors connected in series with its transistors, and furthermore, the diodes in the output boost

stage already provide a path for current flow to the load when the transistors switch off. Likewise with the input buck stage, however, this project can design the snubber for the converter's output boost stage transistors using purely passive components, which do not add a significant amount of PCB real estate to the overall converter design. Working around the two mentioned issues with the snubber design for the output boost stage requires adding a series inductor, capacitor and diode (connected in parallel with each other) for the boost stage's turn-on snubber. This project also implements a turn-off snubber using the same method as with that of the converter's input buck stage.

Designing a turn-off snubber for the converter's output boost stage requires the following parameters:

V_{IN} = Input Voltage

V_{OUT} = Output Voltage

I_{SW} = Maximum Switch (MOSFET) Drain – Source Current

f_s = Switching Frequency

D = Duty Cycle

t_f = I_{SW} fall time (from 90% to 10% of its maximum value)

t_r = I_{SW} rise time (from 10% to 90% of its maximum value)

C = Snubber Capacitor Value

C_{opt} = Optimal Snubber Capacitor Value

C_{dis} = Capacitor Discharge Time

R = Snubber Resistor Value

L = Main Converter Inductor value

L_s = Snubber Inductor Value

$I_{L(0)}$ = Main Converter Inductor Minimum Current Value

Equations (B1.25) to (B1.27) determine the boost stage's duty cycle under this project's maximum converter stress level.

$$V_{OUT} = \frac{V_{IN}}{1-D} \quad (B1.25)$$

$$V_{OUT} = 36V, V_{IN} = 12V \text{ (average)} \quad (B1.26)$$

Therefore:

$$D = 1 - \frac{V_{OUT}}{V_{IN}} = 1 - \frac{12}{36} = 0.667, \text{ or } 67\% \quad (\text{B1.27})$$

For the converter operating under the maximum stress level:

$$V_{IN} = 12V, I_{SW} = 12A, f_s = 100kHz, D = 0.67, t_f = 14.3ns \text{ (taken from earlier plots)} \quad (\text{B1.28})$$

Equation (B1.29) calculates the turn-off snubber capacitor value.

$$C = \frac{I_{SW}t_f}{2V_S} = \frac{12(14.3ns)}{2(12)} = 7.15 \text{ nF} \quad (\text{B1.29})$$

Equation (B1.30) calculates the turn-off snubber resistor value assuming 5 time constants for the capacitor discharge time,

$$R = \frac{\left(\frac{D}{f_s}\right)}{C_{dis}C} = \frac{\left(\frac{0.67}{100kHz}\right)}{(5)(7.15nF)} = 187.41\Omega \quad (\text{B1.30})$$

This project rounds this resistor value down to 185Ω. Equation (B1.31) calculates the turn-off snubber's optimum capacitor value associated with the switching power losses.

$$C_{opt} = \frac{I_{SW}t_f}{\sqrt{12}V_S} = \frac{(12)(14.3ns)}{\sqrt{12}(12)} = 4.1281nF \quad (\text{B1.31})$$

This project rounds this capacitor value up to 4.3nF.

Now this project designs the turn-on snubber for the boost stage. Again, likewise with the buck stage, the turn-on snubber inductor cannot be too large (keeping it in the nH range is optimal). This project selects the snubber inductor as 90nH, which is reasonably small but not so small that it loses its effectiveness in slowing down M_1 's drain current rise.

The converter then underwent several further modifications for safe operation under all of the previously simulated electrical characteristics and conditions. The

modifications include changing the main inductors to 40 μ H and changing one of the output capacitors to 500 μ F. Another modification consists of changing the boost stage transistors from Fairchild HUFA76413DKs to Infineon IPB600N25N3s. The IPB600N25N3 has a 250V maximum drain-source voltage rating, 18-25A maximum continuous drain current rating and 136W power dissipation at 25°C rating. This project selected the IPB600N25N3 as its new boost stage transistors because the previously selected HUFA76413DK's 2.5W continuous power dissipation and 4.8A continuous drain current rating. The IPB600N25N3 has a much lower drain-source on-resistance (21m Ω compared to the HUFA76413DK's 55m Ω), thus it may be less efficient. Figure B-74 shows the converter's new boost stage with the snubbers installed, along with the mentioned modifications.

Figure B-74: Output Stage for Four-Phase Buck, Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with 100 kHz Switching Frequency and Turn-On and Turn-Off Snubbers

Figure B-75 shows the converter's output voltage (in green), buck stage output voltage (in blue), and power dissipation across M_1 (in red) at maximum load stress, with the converter's input voltage set to 60V and load set to 8A.

Figure B-75: Output Voltage, Buck Stage Output Voltage and M_1 Power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

Figure B-76 shows the converter's operation in steady state. The green waveform denotes converter output voltage, the blue waveform denotes buck stage output voltage and the red waveform denotes power dissipation across M_1 .

Figure B-76: Steady-State Output Voltage, Buck Stage Output Voltage and M_1 power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

The average power dissipation across M_1 in steady-state is 4.562W. M_2 behaves in a similar manner. Thus both boost stage transistors dissipate 9.124W in steady-state. The highest power transient occurs at 140W, however, that occurs at converter start-up (before the converter begins regulating). A few high power transients in the 90-100W range still occur in steady-state. However, those transients are far and few in between. This characteristic is acceptable for now. Figure B-77 shows steady-state converter waveforms including the converter's input current (denoted by the teal waveform).

Figure B-77: Steady-State Input Current, Output Voltage, Buck Stage Output Voltage and M_1 Power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

During steady state, the converter's average input voltage is 60V, average input current is -5.5402A, average output voltage is 36.185V, and average output current is 8A. This translates to an overall 87.1% system efficiency, which is roughly the same overall efficiency level as with previous implementations of this converter; however, the power transients across each transistor are drastically lower in this implementation because of the additional snubber circuits.

Figure B-78 is a zoomed-in plot showing M_1 's drain current switching trajectory (in teal) during the occurrence of a 100W transient dissipated across it (in red):

Figure B-78: Output Voltage, Buck Stage Output Voltage, M_1 Drain Current and M_1 Power Dissipation of Modified 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

The power transients during transistor turn-off are not as lethal as during turn-on. Thus the converter's turn-on snubber requires further optimization. Additionally, these transients occur only once the current through inductor L_8 (which is the switching inductor for M_1) reaches its minimum (i.e. when L_8 releases its stored energy). Appendix B1.10 describes the turn-on snubber optimization process.

B1.10 Component Downsizing and Optimization Process

Minimizing the converter's physical size and cost as much as possible as well as guaranteeing physical component availability required the following further changes for this project's converter:

- All primary switching transistors in the input buck stage (Q_1 , Q_3 , Q_7 and Q_9) changed to STMicroelectronics STW11NM80 (800V V_{DS} capability, 0.35Ω R_{DS-ON}) for handling the 27A pulses (and 80W power transients) during turn-on states.

- All input buck stage snubber diodes changed to 1N914 silicon diodes (75V reverse voltage capability, 0.2A average forward current, 4A maximum pulsed current).
- All output boost stage snubber diodes changed to MBR140 Schottky diodes (40V reverse voltage capability, 1A average forward current, 40A maximum pulsed current), available in SMB (surface mount package).
- All input buck stage stage V_{CC} input diodes changed to 1N914 silicon diodes (75V reverse voltage capability, 200mA average forward current, 1-4A maximum pulsed current).
- All primary input buck stage inductors reduced to 39 μ H, as no commercial 60 μ H inductor exists (as of writing) that meets the converter's 15A saturation current requirement. As later simulation tests show, the converter still maintains CCM operation at full load.
- The physical converter design will use Fairchild FDB14N30s (100V V_{DS} capability, 0.29 Ω R_{DS-ON}) for the primary transistors in the input buck stage changed to for offsetting the costs from the higher-priced 800V STM transistors. The drain-source on resistance is nearly identical to that of the 800V STM transistors and the FDB14N30s can handle 14A continuous drain current, as well as 56A maximum pulsed current. No simulation model exists for the Fairchild FDB14N30 transistors, thus further simulations continue using the 800V STM transistors.

Figure B-79 shows the converter schematic with all the mentioned changes.

Figure B-79: Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

Figure B-80 is a plot showing converter output voltage in green, input buck stage voltage in blue, and power dissipated across Q_1 (in red), the primary buck input stage transistor, in steady-state operation with a 60V input and 8A load.

Figure B-80: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

The power transients increase into the 120-160W range; however, these power levels are still acceptable for the selected transistor. The average power dissipated across each of these primary transistors in steady-state operation is 2.79W, which is still acceptable, however these transistors would still require adequate heatsinking.

Figure B-81 is the same plot as figure B-80 above, except it now denotes in red the power dissipation across M_1 , the primary output boost-stage transistor, in steady-state operation.

Figure B-81: Output Voltage, Buck Stage Output Voltage, and M_1 Power Dissipation of Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

The power transients across M_1 still lie in the 40-45W range (maximum) in steady-state operation. Average power dissipation across M_1 in steady state is roughly 4.94W (thus, still near 5W), which means that the converter does not need any further changes to its boost output stage; however, 5W average dissipation means that these boost output stage transistors will require good heatsinking.

The converter still has one lingering problem, however - the input buck stage snubber diodes' electrical capabilities. Initially, before using the 1N914 silicon diodes, this project selected MBRS1100 Schottky diodes as the converter's buck stage snubber diodes. Unfortunately those diodes drastically increased the power transient magnitudes across Q_1 under full load. The root of that problem is that the MBRS1100 Schottky diode has a much higher junction capacitance (roughly 10 times larger) than that of the MBR20100CT Schottky diodes used in earlier simulations, according to LTSpice. The 1N914 silicon diodes have reduced junction capacitance; however, its maximum pulsed

current rating is only 4A and the converter requires a diode that tolerates at least 6A pulsed current through the snubber circuits. This project then changed the 1N914 diodes to ES1Ds, allowing for the snubber to tolerate the simulated maximum pulsed current requirements. The ES1D is a silicon diode with very low junction capacitance (roughly 10.5 pF, about the same as that of the MBR20100CT), and according to LTSpice it can tolerate up to 200V reverse voltage, 1A average (continuous) forward current, as well as 30A pulsed current (provided that the current pulse is 8.3ms or less). Figure B-82 shows the converter circuit with the new ES1D snubber diodes.

Figure B-82: Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with ES1D Snubber Diodes

Figure B-83 shows this modified converter's operation in steady-state, with a 60V input and 8A load. Converter output voltage is in green and buck stage output voltage is in blue. The red waveform denotes the power dissipation across Q_1 .

Figure B-83: Steady-State Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Optimized 100 kHz Two-Phase boost, Multi-Stage Buck-Boost Converter Design with ES1D Snubber Diodes

With the ES1D diodes in place, the power transient magnitudes across Q_1 lie between 140-160W, which is still acceptable for this converter. Overall average power dissipation across Q_1 in steady state at full load is 2.77W, which is nearly unchanged from the last circuit design iteration. Figure B-84 is the same plot as Figure B-83, except now the red waveform denotes the power dissipated across M_1 , the primary output boost stage transistor.

Figure B-84: Steady-State Output Voltage, Buck Stage Output Voltage, and M_1 Power Dissipation of Optimized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with ES1D Snubber Diodes

This power dissipation characteristic is very much unchanged from figure B-81, with the exception of the two stray transients in the 120-130W magnitude range. However, this does not pose a problem, as those transients are very brief (roughly 1ns) and operation with such power levels is still within the output boost stage transistors' safe operating areas. Overall average power dissipation across M_1 in steady state at full load is roughly 4.9W, again, nearly unchanged from the previous circuit design iterations. The next section describes additional circuitry required for this circuit for proper stage and phase synchronization.

B1.11 Linear LTC6908-1 External Oscillator Configuration

Figure B-85 shows the additional circuitry required for synchronizing each PWM controller (per phase) on the converter's buck input stage, using an LTC6908-1 external oscillator IC.

Figure B-85: LTC6908-1 Configuration for use with the LT3845A Controllers

The resistor tied to the SET3 pin sets the output frequency, as determined by the formula $f_{SET} = 10 \text{ MHz} \cdot \frac{10k}{(R_{SET})}$, where f_{SET} is the converter's switching frequency (100kHz). For 100kHz switching frequency, R_{SET} must be 1MΩ. The V^+ pin also requires 2.7-5.5V input for maintaining the proper oscillation frequency, and a zener diode in conjunction with an NPN BJT generates that regulated voltage. The V^+ pin also must be coupled to the LTC6908-1's signal ground using a 0.1μF capacitor. Q₁'s collector and R₁₂ are both connect to the V_{CC} pin on the LT3845A PWM controller at the buck input stage. Figure B-86 shows this converter's final design, with the LTC6908-1 external oscillator ICs included.

Figure B-86: Complete 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with LTC6908-1 External Oscillators

Each OUT1 pin on the LTC6908-1 IC connects to the SYNC pins (not shown in the schematic in figure B-86) of its respective LTC3845A controllers; with 2 phases connected per OUT1 pin. This project leaves the OUT2 unused thus it connects to ground. The next section describes setting up the LTC6908-1 external oscillator IC and why this project leaves its OUT2 pin unused.

B1.12 Test Circuit for LTC6908-1 External Oscillator

Ensuring that the LTC6908-1 external oscillator produces the proper switching signal for synchronizing the input stage's four phases first requires a test simulation circuit for the chip. Figure B-87 shows the test circuit for the LTC6908-1 external oscillator's output.

Figure B-87: Test Circuit for LTC6908-1 External Oscillator

The OUT2 pin on the LTC6908-1 produces a signal 180° out of phase relative to the signal the chip outputs on its OUT1 pin. Thus, this project only uses the OUT1 pin on the chip. This test circuit determines if a single LTC6908-1 IC can drive all four of the converter's input buck stage stage's phases without any oscillator signal degradation, or if this project's converter requires LTC6908-1 chips. If this converters requires two LTC6908-1 chips, this test will must ensure that both chips' OUT1 output signals are completely in phase. In the final converter, a wire to the converter's LT3845A's PWM controllers' V_{CC} pins (which supplies roughly the same 16V from each pin) replaces voltage source V₂ in figure B-87. The BZX84C6V2L is a 6.2V zener diode that helps step down the V_{CC} pin voltage to the 2.7-5.5V required for the LTC6908-1's V⁺ pin. After selecting the parts for this project's buck-boost converter design, this project next determines the converter's cost.

B1.13 Converter Cost Analysis

Tables B.B-1 to B.B-6 (each separated by component type) show the converter's component cost, excluding the additional required cost for PCB fabrication as well as safety components such as fuses. As a prototype, the converter also requires test points (a negligible additional cost relative to the overall converter cost) placed on its PCB for easy problem diagnosis; however, the below tables show the cost for the converter in its final incarnation.

Table B.B-1: Controller IC Components Cost for Converter

Controller IC	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
LT3845A	4	7.13	28.52	U1, U3, U4, U5	SMT
LTC3862	1	7.00	7.00	U2	SMT
LTC6908-1	2	3.92	7.84	U6, U7	SMT
Total Cost (\$)			43.36		

Table B.B-2: Transistor Components Cost for Converter

Transistor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
Fairchild FDB14N30	4	1.38	5.52	Q1, Q3, Q7, Q9	SMT
IRF510	4	1.51	6.04	Q2, Q6, Q8, Q10	Through-Hole
IPB600N25N3	2	2.50	5.00	M1, M2	SMT
2N2222	2	0.53	1.06	Q4, Q5	Through-Hole
Total Cost (\$)			17.62		

Table B.B-3: Diode components Cost for Converter

Diode	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
MBR20100CT	6	0.91	5.46	D2, D3, D4, D7, D10, D13	Through-Hole
MBRS140	4	0.55	2.20	D15, D16, D29, D30	SMT
B540C	4	0.75	3.00	D1, D6, D9, D12	SMT
ES1D	12	0.53	6.36	D17 to D28	SMT
1N4148	4	0.33	1.32	D5, D8, D11, D14	SMT
BZX84C6V2L	2	0.46	0.92	D31, D32	SMT
Total Cost (\$)			19.26		

Table B.B-4: Resistor Components Cost for Converter

Resistor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
1M, 1/8W, 1%	3	0.07	0.21	R1, R40, R42	SMT
357k, 1/8W, 1%	1	0.04	0.04	R14	SMT
191k, 1/8W, 1%	4	0.07	0.28	R4, R17, R21, R25	SMT
143k, 1/8W, 1%	1	0.07	0.07	R5	SMT
130k, 1/8W, 1%	1	0.07	0.07	R11	SMT
84.5k, 1/8W, 1%	1	0.04	0.04	R16	SMT
82.5k, 1/8W, 1%	1	0.04	0.04	R7	SMT
26.7k, 1/8W, 1%	1	0.07	0.07	R12	SMT
25.5k, 1/8W, 1%	2	0.04	0.08	R39, R41	SMT
24.9k, 1/8W, 1%	1	0.04	0.04	R15	SMT
20k, 1/8W, 1%	2	0.04	0.08	R19, R27	SMT
16.2k, 1/8W, 1%	1	0.04	0.04	R6	SMT
12.4k, 1/8W, 1%	1	0.05	0.05	R13	SMT
360, 2W, 5%	4	0.35	1.40	R22, R30, R32, R34	Through-Hole
200, 2W, 5%	2	0.54	1.08	R37, R38	Through-Hole
10, 1/2W, 1%	2	0.23	0.46	R28, R36	Through-Hole
0.05, 1/2W, 5%	4	0.50	2.00	R23, R31, R33, R35	SMT
0.006, 1W, 1%	6	0.56	3.36	R3, R18, R20, R24, R26, R29	SMT
Total Cost (\$)			9.41		

Table B.B-5: Inductor Components Cost for Converter

Inductor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
39 μ H, $I_{SAT}=17A$	6	7.06	42.36	L1, L4, L5, L6, L7, L8	Through-Hole
90nH, $I_{SAT}=17A$	2	2.45	4.90	L2, L3	SMT
75nH, $I_{SAT}=17A$	4	2.36	9.44	L9, L10, L11, L12	SMT
Total Cost (\$)			56.70		

Table B.B-6: Capacitor Components Cost for Converter

Capacitor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
560 μ , Electrolytic, 100V	1	2.13	2.13	C44	Through-Hole
100 μ , Electrolytic, 100V	4	1.16	4.64	C21, Additional input filter capacitors	Through-Hole
100 μ , Electrolytic, 50V	1	0.81	0.81	C43	SMT
40 μ , Electrolytic, 35V	1	0.55	0.55	C4	SMT
22 μ , Electrolytic, 50V	3	2.25	6.75	C32, C33, C38	SMT
10 μ , Ceramic, 50V	4	0.69	2.76	C39, C40, C41, C42	SMT
4.7 μ , Ceramic, 10V	1	0.30	0.30	C15	SMT
1 μ , Ceramic, 50V	9	0.24	2.16	C1, C5, C6, C14, C16, C19, C22, C25, C26	SMT
0.1 μ , Ceramic, 100V	3	0.40	1.20	C12, C24, C31	SMT
0.1 μ , Ceramic, 10V	2	0.05	0.10	C47, C48	SMT
10000p, Ceramic, 6.3V	3	0.08	0.24	C8, C13, C30	SMT
4700p, Ceramic, 50V	2	0.08	0.16	C45, C46	SMT
2200p, Ceramic, 10V	6	0.08	0.48	C3, C7, C17, C20, C23, C27	SMT
1000p, Ceramic, 10V	2	0.36	0.72	C9, C11	SMT
750p, Ceramic, 100V	4	0.47	1.88	C34, C35, C36, C37	Through-Hole
100p, Ceramic, 10V	1	0.40	0.40	C10	SMT
47p, Ceramic, 25V	4	0.09	0.36	C2, C10, C28, C29	SMT
Total Cost (\$)			25.64		

Total component cost for this converter is \$171.99. The converter also has 152 total base components, excluding any safety components or test points. Because the estimated size of the required PCB is fairly large, ExpressPCB typically requires an additional \$233 as a base price [76] for fabricating a custom-sized PCB such as what this converter requires (resulting in a minimum total cost of \$404.99 for a final production-level PCB). These costs are unfeasible if the EHFEM elliptical trainer system requires a 10-year payback period. Thus, this converter design must be downsized in order to reduce component and overall converter cost. The next section describes a possible converter downsizing proposal.

B1.14 Converter Downsize Proposal

One way to reduce this project's converter's component count and overall cost involves using only one LT3845A PWM controller IC to drive all four input stage phases. All the required pin connections on all four phases connect to a single LT3845A controller. This configuration also requires only one current sense resistor (located on the converter's primary phase); however the disadvantage with this design is that the LT3845A controller will not protect the other phases from any over-current conditions, as the current through each phase can become unbalanced if such an event occurs. Another advantage, however, is that this design does not require an additional LTC6908-1 external oscillator IC, as the frequency-set resistor tied to the LT3845A's Fset pin suffices. Figure B-88 shows this reduced-part converter design.

Figure B-88: Downsized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design with LTC6908-1 External Oscillators

Figure B-89 shows the converter operating at maximum stress conditions (60V input, 8A load) in steady state operation, with converter output voltage in green, input buck stage output voltage in blue, and power dissipation across Q_1 (which remains our primary switching transistor for the primary input phase) in red.

Figure B-89: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Downsized 100 kHz Two-Phase Boost, Multi-stage Buck-Boost Converter Design

Figure B-90 is same plot as figure B-89, but with the input buck stage output voltage in a separate plot pane.

Figure B-90: Output Voltage, Buck Stage Output Voltage and Q_1 Power Dissipation of Downsized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

Q_1 's average power dissipation drastically reduces to 1.85W. The other primary switching transistors in the other phases follow a similar trend as long as the current through each phase is balanced (i.e. equal through each phase). Figure B-91 shows in teal the power dissipation across Q_3 , the primary switching transistor on the input stage's second phase.

Figure B-91: Output Voltage, Buck Stage Output Voltage and Q_1 and Q_3 Power Dissipation of Downsized 100 kHz Two-Phase Boost, Multi-Stage Buck-Boost Converter Design

Q_3 's power dissipation waveform is identical to that of Q_1 's. From figure B-91, the converter's average output voltage only reaches a 31.14V, which is significantly below the 36V from the converter's original specifications. Thus this converter configuration cannot provide the Enphase micro-inverter with the optimal input voltage that requires for maximizing its efficiency. In reality, a single LT3845A controller may also have problems driving more than one parallel phase, which may have resulted in the converter's reduced output voltage in simulation. The next section describes this converter's costs.

B1.15 Downsized Converter Cost Analysis

Likewise with appendix B1.14, the component cost for this downsized converter does not include additional costs for PCB fabrication as well as safety components and

test points. Tables B.B-7 to B.B-12 (each separated by component type) show this converter's component cost.

Table B.B-7: Controller IC Components Cost for Downsized Converter

Controller IC	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
LT3845A	1	7.13	7.13	U1	SMT
LTC3862	1	7.00	7.00	U2	SMT
Total Cost (\$)			14.13		

Table B.B-8: Transistor Components Cost for Downsized Converter

Transistor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
Fairchild FDB14N30	4	1.38	5.52	Q1, Q3, Q7, Q9	SMT
IRF510	4	1.51	6.04	Q2, Q6, Q8, Q10	Through-Hole
IPB600N25N3	2	2.50	5.00	M1, M2	SMT
Total Cost (\$)			16.56		

Table B.B-9: Diode Components Cost for Downsized Converter

Diode	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
MBR20100CT	6	0.91	5.46	D2, D3, D4, D7, D10, D13	Through-Hole
MBRS140	4	0.55	2.20	D15, D16, D29, D30	SMT
B540C	1	0.75	0.75	D1	SMT
ES1D	12	0.53	6.36	D17 to D28	SMT
1N4148	1	0.33	0.33	D5	SMT
Total Cost (\$)			15.10		

Table B.B-10: Resistor Components Cost for Downsized Converter

Resistor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
1M, 1/8W, 1%	1	0.07	0.07	R1	SMT
357k, 1/8W, 1%	1	0.04	0.04	R14	SMT
191k, 1/8W, 1%	1	0.07	0.07	R4	SMT
143k, 1/8W, 1%	1	0.07	0.07	R5	SMT
130k, 1/8W, 1%	1	0.07	0.07	R11	SMT
84.5k, 1/8W, 1%	1	0.04	0.04	R16	SMT
82.5k, 1/8W, 1%	1	0.04	0.04	R7	SMT
26.7k, 1/8W, 1%	1	0.07	0.07	R12	SMT
24.9k, 1/8W, 1%	1	0.04	0.04	R15	SMT
16.2k, 1/8W, 1%	1	0.04	0.04	R6	SMT
12.4k, 1/8W, 1%	1	0.05	0.05	R13	SMT
360, 2W, 5%	4	0.35	1.40	R22, R30, R32, R34	Through-Hole
200, 2W, 5%	2	0.54	1.08	R37, R38	Through-Hole
10, 1/2W, 1%	2	0.23	0.46	R28, R36	Through-Hole
0.05, 1/2W, 5%	4	0.50	2.00	R23, R31, R33, R35	SMT
0.006, 1W, 1%	3	0.56	1.68	R3, R24, R29	SMT
Total Cost (\$)			7.22		

Table B.B-11: Inductor Components Cost for Downsized Converter

Inductors	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
39 μ H, $I_{SAT}=17A$	6	7.06	42.36	L1, L4, L5, L6, L7, L8	Through-Hole
90nH, $I_{SAT}=17A$	2	2.45	4.90	L2, L3	SMT
75nH, $I_{SAT}=17A$	4	2.36	9.44	L9, L10, L11, L12	SMT
Total Cost (\$)			56.70		

Table B.B-12: Capacitor Components Cost for Downsized Converter

Capacitor	Quantity	Unit Price (\$)	Total Price (\$)	Components	Type
560 μ , Electrolytic, 100V	1	2.13	2.13	C44	Through-Hole
100 μ , Electrolytic, 100V	4	1.16	4.64	C21, Additional input filter capacitors	Through-Hole
100 μ , Electrolytic, 50V	1	0.81	0.81	C43	SMT
40 μ , Electrolytic, 35V	1	0.55	0.55	C4	SMT
22 μ , Electrolytic, 50V	3	2.25	6.75	C32, C33, C38	SMT
10 μ , Ceramic, 50V	4	0.69	2.76	C39, C40, C41, C42	SMT
4.7 μ , Ceramic, 10V	1	0.30	0.30	C15	SMT
1 μ , Ceramic, 50V	3	0.24	0.72	C1, C5, C14	SMT
10000p, Ceramic, 6.3V	3	0.08	0.24	C8, C13, C30	SMT
4700p, Ceramic, 50V	2	0.08	0.16	C45, C46	SMT
2200p, Ceramic, 10V	2	0.08	0.16	C3, C27	SMT
1000p, Ceramic, 10V	2	0.36	0.72	C9, C11	SMT
750p, Ceramic, 100V	4	0.47	1.88	C34, C35, C36, C37	Through-Hole
100p, Ceramic, 10V	1	0.40	0.40	C10	SMT
47p, Ceramic, 25V	2	0.09	0.18	C2, C29	SMT
Total Cost (\$)			22.40		

Using only one LT3845A controller for driving all four input stage phases, reduces the total converter component cost to \$132.11. The converter's component count also reduces to a 108 base component total, excluding any safety components or test points. However, this downsized converter still requires a custom-sized PCB because of converter's inductors' physical sizes. The PCB size that this downsized converter requires is still larger than what PCBExpress's mini-board option offers (a 2.5" × 3.8" PCB). Again, likewise with the previous converter design, because the required PCB's estimated size is fairly large, ExpressPCB typically requires an additional \$233 as a base price [76] for fabricating a custom-sized PCB such as what this downsized converter requires. This fabrication cost results in a \$365.11 minimum total cost for a final production-level PCB. These costs are still unfeasible if the EHFEM elliptical trainer system requires a 10-year payback period. Even then, a 31.14V output from this converter design is unacceptable, as the Enphase micro-inverter requires 36V input for operating at maximum efficiency. Even with the previous converter design, its overall simulated converter efficiency was roughly 87.1%. If that design was physically implemented, parasitic characteristics of various components, such as ESR on capacitors, would still exist between PCB traces. Thus, the actual converter efficiency will be significantly lower than the results from simulations. The overall converter cost is not worth the converter's efficiency loss (relative to the rest of the EHFEM elliptical trainer system). Furthermore, if designs and functionality problems occur during the physical testing phase, troubleshooting the converter would be very time-consuming and difficult because of the large number of components and intricate circuit complexity.

Although the interleaved, multi-stage buck-boost converter described in Appendices A and B initially seemed like a feasible design for the EHFEM project after running simulations on it and mitigating its switching transistor power transient issues, its overall cost was too high for justifying its maximum efficiency level. Its exorbitantly high cost also results in an unreasonable payback period for the EHFEM system's life cycle. Furthermore, the converter's intricate complexity would make troubleshooting very difficult if it were ever physically fabricated onto a PCB. For such a complex design, anything could go wrong during testing. After discovering that the Precor EFX 546i elliptical trainer's onboard 12V battery could be used to supply a bias voltage for a converter's PWM controllers, this project decided on the SEPIC topology described in chapter 3 for its DC-DC converter design. The basic SEPIC design from chapter 3 yields a fairly low component count, which results in a smaller PCB required for the converter, as well as a low overall converter cost.

Therefore, this project scrapped entire downsized interleaved buck-boost converter design and its previous design iterations in favor of a new converter design that minimizes component count and overall cost. Chapter 3 details selecting the new converter design used for defending this report's thesis statement.

APPENDIX C: FINAL SEPIC NETLIST

```
M$Q_MAIN SW N014 N017 N017 IPP110N20N3
RFREQ N018 0 220k tol=1 pwr=0.125
RFB2 FB 0 3.92k tol=1 pwr=0.125
RFB1 OUT FB 115k tol=1 pwr=0.125
L1 CLAMP_INPUT SW 60μ Ipk=28 Rser=0.0085
RTH N013 N012 33.2K tol=1 pwr=0.125
CTH2 N012 0 6800p V=10 Rser=0.02
CTH1 N013 0 56p V=16 Rser=0.02
XU1 RUN N013 FB N018 VCC 0 N014 VCC VBAT SENSE LTC1871
VIN IN 0 65
L2 0 N005 60μ Ipk=28 Rser=0.0085
COUT OUT 0 470μ V=400 Irms=6 Rser=0.19
RRUN2_MAIN RUN 0 133K tol=1 pwr=0.125
RRUN1_MAIN VBAT RUN 274K tol=1 pwr=0.125
COUT2 0 OUT 10μ V=50 Rser=0.02
VBAT VBAT 0 12
RSNS N017 0 5m tol=1 pwr=5
LSNUB N005 N010 360n Ipk=31.5 Rser=1.4m
DSNUB1 N005 N015 APT30S20BG
DSNUB2 N015 N016 MBR20100CT
DSNUB3 N016 P001 MBR20100CT
CSNUB1 N010 N016 560n V=16 Irms=8.771 Rser=0.02 Lser=0
CSNUB2 OUT N015 10n V=250 Rser=0.02
D_MAIN3 N010 N011 MBR20100CT
CFB1 FB 0 47p V=16 Rser=0.02
CSNS2 SENSE 0 0.03μ V=16 Rser=0.02
RSNS_OVP IN N001 2m tol=1 pwr=5
RGATE_OVP N002 N004 10 tol=1 pwr=0.125
RRUN1_OVP CLAMP_INPUT N008 255k tol=1 pwr=0.125
RRUN2_OVP N008 0 4.99K tol=1 pwr=0.125
CTMR N009 0 47p V=16 Rser=0.02
XU2 N009 N008 CLAMP_INPUT N004 N001 IN IN N006 RUN 0 0 0 LT4356-1
CGATE_OVP N004 0 6800p V=100 Rser=0.02
M$Q_OVP N001 N002 CLAMP_INPUT CLAMP_INPUT IPP110N20N3
D$TSD 0 IN 1N5378B
RFLT N007 N006 3.6k tol=1 pwr=0.125
D$FLT_LED VBAT N007 NSCW100
CIN IN N003 1000μ V=100 Irms=1.52 Rser=0.047
CCLAMP CLAMP_INPUT 0 47μ V=100 Irms=0.5 Rser=0.32
CBAT VBAT 0 33μ V=35 Irms=0.042 Rser=0.3
CVCC VCC 0 4.7μ V=16 Rser=0.02
COUT3 0 OUT 10μ V=50 Rser=0.02
CBAT2 0 VBAT 10μ V=25 Rser=0.02
CBAT3 0 VBAT 10μ V=25 Rser=0.02
COUT4 0 OUT 10μ V=50 Rser=0.02
CIN2 0 IN 0.1μ V=100 Rser=0.02
CIN3 0 IN 0.1μ V=100 Rser=0.02
CCLAMP2 0 CLAMP_INPUT 0.1μ V=100 Rser=0.02
CCLAMP3 0 CLAMP_INPUT 0.1μ V=100 Rser=0.02
CINT2 N005 SW 4.7μ V=250 Rser=0.02
CINT3 N005 SW 4.7μ V=250 Rser=0.02
```

```

CCLAMP4 0 CLAMP_INPUT 0.1μ V=100 Rser=0.02
CINT N005 SW 15μ V=250 Irms=4 Rser=0.02
CINT4 N005 SW 4.7μ V=250 Rser=0.02
CINT5 N005 SW 4.7μ V=250 Rser=0.02
CINT6 N005 SW 4.7μ V=250 Rser=0.02
D_MAIN1 N010 N011 MBR20100CT
D_MAIN2 N011 OUT MBR20100CT
D_MAIN4 N011 OUT MBR20100CT
DSNUB4 P001 OUT MBR20100CT
CRUN_OVP N008 0 47p
CRUN_MAIN RUN 0 47p
CBAT4 0 VBAT 100n V=25 Rser=0.02
CBAT5 0 VBAT 100n V=25 Rser=0.02
COUT_CPH1 0 OUT 0.1μ V=50 Rser=0.02
COUT_CPH2 0 OUT 0.1μ V=50 Rser=0.02
CINT7 N005 SW 0.1μ
CINT8 N005 SW 0.1μ
ILOAD OUT 0 8
RVCC1 VBAT VCC 887
RVCC2 VCC 0 680
CSNS1 SENSE 0 47p V=16 Rser=0.02
RSNS_FILT N017 SENSE 10
RGATE_MAIN N014 N019 45
CGATE_MAIN N019 0 0.02μ Rser=0.02
DGATE_MAIN N014 N019 1N4148
R_ELLIPTICAL N003 0 10
CFB2 FB 0 0.01μ V=16 Rser=0.02
CINT9 N005 SW 0.1μ
CINT10 N005 SW 0.1μ
COUT_CPH3 0 OUT 0.1μ V=50 Rser=0.02
COUT_CPH4 0 OUT 0.1μ V=50 Rser=0.02
.model D D
.lib C:\PROGRA~2\LTC\LTSPIC~1\lib\cmp\standard.dio
.model NMOS NMOS
.model PMOS PMOS
.lib C:\PROGRA~2\LTC\LTSPIC~1\lib\cmp\standard.mos
.tran 0 30m 18m 30m steady startup
.temp 40
.model APT30S20BG D(Is=15m Rs=0.005 N=1.5 Isr=15m Cjo=150p Vj=0.86
Iave=45 Vpk=200 mfg=OnSemi type=Schottky)
.lib LT4356-1.sub
.lib LTC1871.sub
.backanno
.end

```

REFERENCES

- [1] J. Yuen; M. Lum; C. Cinkornpumin; J. Chan. “Energy Harvesting From Exercise Machines (EHFEM) Self-generator Elliptical Machine,” *Cal Poly Digital Commons*. 2008. [Online]. <http://digitalcommons.calpoly.edu/eesp/12>
- [2] Precor, “EFX® 546i Elliptical Fitness Crosstrainer™ Specifications” Precor EFX 546i product page, 2010. [Online]. <http://www.precor.com/products/en/efx-546i-elliptical-fitness-crosstrainer-1#>
- [3] Enphase Energy, “Enphase Micro-Inverter,” M175-24-240 datasheet. [Online]. http://enphase.com/downloads/8261_Datasheet_24_32.pdf
- [4] Vicor Power, “28V Wide Input Maxi Family DC-DC Converter Module Data Sheet” Vicor Maxi, Mini and Micro DC-DC Converters product page, 2010. [Online]. http://cdn.vicorpower.com/documents/datasheets/ds_28vin-maxi-family.pdf
- [5] Underwriters Laboratory code 1741 (2007). *Inverters, converters, controllers, and interconnection system equipment for use with distributed energy resources* [Online]. Available: <http://ulstandardsinfonet.ul.com/scopes/scopes.asp?fn=1741.html>
- [6] IEEE Standards Association (2010, Dec. 28). *IEEE 1547 Standard for Interconnecting Distributed Resources with Electric Power Systems* [Online]. Available: http://grouper.ieee.org/groups/scc21/1547/1547_index.html
- [7] Pacific Gas and Electric Company (2004, Aug. 11). *Summary of General Technical Requirements for the Interconnection of Distributed Generation (DG) to PG&E’s Distribution System* [Online]. Available: http://www.pge.com/includes/docs/pdfs/shared/newgenerator/ebio/dg_technical_white_paper_0419_attachment.pdf
- [8] National Fire Protection Association (n.d.). *NFPA 70® National Electric Code® 2011 Edition* [Online]. Available: http://www.nfpa.org/onlinepreview/online_preview_document.asp?id=7011SB
- [9] D. Braun. “Soft Switching DC-DC Conversion to Harvest Exercise Machine Electricity EISG Proposal,” 2010.
- [10] A.J. Hilario, “Energy Harvesting from Elliptical Machines Using Four-Switch Buck Boost Topology,” *Cal Poly Digital Commons*. 2011. [Online]. <http://digitalcommons.calpoly.edu/theses/511/>

- [11] N. Mohan, T. Undeland, et al., "Switching DC Power Supplies," in *Power Electronics: Converters, Applications and Design*, 2nd Ed. New York: Wiley, 1995, pp. 310-319.
- [12] A.J. Prabhakar; J.D. Bollinger; H.T. Ma; M. Ferdowsi; K. Corzine. "Efficiency analysis and comparative study of hard and soft switching DC-DC converters in a wind farm," *Industrial Electronics, 2008. IECON 2008. 34th Annual Conference of IEEE*, pp.2156-2160, 2008.
- [13] NFN Taufik, *Advanced Power Electronics Design*, Winter 2010, San Luis Obispo, CA: California Polytechnic State University, 2010.
- [14] N. Mohan, T. Undeland, et al., "DC-DC Switch-Mode Converters," in *Power Electronics: Converters, Applications and Design*, 2nd Ed. New York: Wiley, 1995, pp. 179-188.
- [15] M. Gaboriault; A. Notman. "A high efficiency, noninverting, buck-boost DC-DC converter," *Applied Power Electronics Conference and Exposition, 2004. APEC '04. Nineteenth Annual IEEE*, vol.3, pp. 1411- 1415, 2004.
- [16] W. Huang; K. Yen; G. Roig; E. Lee. "Voltage divided noninverting Cuk converter with large conversion ratios," *Southeastcon '91., IEEE Proceedings of*, vol. 2, pp.1005-1007, 1991.
- [17] M. Orellana; S. Petibon; B. Estibals; C. Alonso. "Four Switch Buck-Boost Converter for Photovoltaic DC-DC power applications," *IECON 2010 - 36th Annual Conference on IEEE Industrial Electronics Society*, pp.469-474, 2010.
- [18] Wikipedia, "Single-ended primary-inductor converter", 2012. [Online]. <http://en.wikipedia.org/SEPIC>
- [19] E. Niculescu; M.C. Niculescu; D.M. Purcaru. "Modelling the PWM Zeta converter in discontinuous conduction mode," *Electrotechnical Conference, 2008. MELECON 2008. The 14th IEEE Mediterranean*, pp.651-657.
- [20] K.B. Park; H.W. Seong; H.S. Kim; G.W. Moon; M. J. Youn. "Integrated boost-sepic converter for high step-up applications," *Power Electronics Specialists Conference, 2008. PESC 2008. IEEE*, pp.944-950, 2008.
- [21] Maxim IC, "SEPIC Equations and Component Ratings," *Maxim Design Support*, 2011. [Online]. <http://www.maxim-ic.com/app-notes/index.mvp/id/1051>

- [22] T.F. Wu; Y.K. Chen. "Modeling PWM DC/DC converters out of basic converter units," *Power Electronics, IEEE Transactions on*, vol.13, no.5, pp.870-881, 1998.
- [23] J.P.M. Figueiredo; F.L. Tofoli; R.L. Alves. "Comparison of nonisolated dc-dc converters from the efficiency point of view," *Power Electronics Conference (COBEP), 2011 Brazilian*, pp.14-19, 2011.
- [24] Texas Instruments, "Designing DC/DC converters based on ZETA topology." *Texas Instruments*. 2012. [Online]. <http://www.ti.com/lit/an/slyt372/slyt372.pdf>
- [25] NFN Taufik, *Introduction to Power Electronics*, Fall 2009, San Luis Obispo, CA: California Polytechnic State University, 2009.
- [26] A.C.-C. Hua; B.C.-y Tsai. "Design of a wide input range DC/DC converter based on SEPIC topology for fuel cell power conversion," *Power Electronics Conference (IPEC), 2010 International*, pp.311-316, 2010.
- [27] Texas Instruments, "LM3478, LM3488 Application Note 1484 Designing A SEPIC Converter." *DC/DC Controller (External Switch) - Step-Up Controller - LM3478 – TI.com*. 2011. [Online]. <http://www.ti.com/lit/an/slyt372/slyt372.pdf>
- [28] NFN Taufik, *Switching Mode Power Supply: Components and Design*, Spring 2010, San Luis Obispo, CA: California Polytechnic State University, 2010.
- [29] J. Chen; R. Erickson; D. Maksimovic. "Averaged switch modeling of boundary conduction mode DC-to-DC converters," *Industrial Electronics Society, 2001. IECON '01. The 27th Annual Conference of the IEEE*, vol.2, pp.844-849, 2001.
- [30] C.H. Chia; P.S. Lei; R.C.-H. Chang. "A high-efficiency PWM DC-DC buck converter with a novel DCM control under light-load," *Circuits and Systems (ISCAS), 2011 IEEE International Symposium on*, pp.237-240, 2011.
- [31] Y.T. Chen; C.H. Chen. "A DC-DC buck converter chip with integrated PWM/PFM hybrid mode control circuit," *Power Electronics and Drive Systems, 2009. PEDS 2009. International Conference on*, pp.181-186, 2009.
- [32] G. Rizzoni, "Energy-Storage (Dynamic) Circuit Elements," in *Principles and Applications of Electrical Engineering*, 3rd Ed. Boston: McGraw-Hill, 2000, p. 126.
- [33] M. Sadiku and C.K. Alexander, "Inductors," in *Fundamentals of Electric Circuits*, 3rd Ed. Boston: McGraw-Hill, 2005, pp. 212-213.

- [34] A. Pressman, K. Billings, et al., "MOSFET Basics," in *Switching Power Supply Design*, 3rd Ed. New York: McGraw-Hill, 2009, pp. 459-464.
- [35] Linear Technology, "Wide-Input Range, No Rsense Current Mode Boost, Flyback, and SEPIC Controller," LTC1871 product page, 2010. [Online].
<http://cds.linear.com/docs/Datasheet/1871fe.pdf>
- [36] Vishay Siliconix, "Filter Inductors, High Current, Radial Leaded," Vishay IHV28BZ60 datasheet, 2012. [Online].
<http://www.vishay.com/docs/34022/ihv.pdf>
- [37] S. Franco, "Stability," in *Design with Operational Amplifiers and Analog Integrated Circuits*, 3rd Ed. Boston: McGraw-Hill, 2002, pp. 365-372.
- [38] A.J. Yiin; R.D. Schrimpf; K.F. Galloway. "Gate-charge measurements for irradiated n channel DMOS power transistors," *Nuclear Science, IEEE Transactions on*, vol.38, no.6, pp.1352-1358, 1991.
- [39] Infineon Technologies, "OptiMOS™ 3 Power-Transistor," IPP110N20N3 datasheet, 2011. [Online].
http://www.infineon.com/dgdl/IPP_I_110N20N3+G_IPB107N20N3+G+Rev2.3.pdf?folderId=db3a3043163797a6011637d4bae7003b&fileId=db3a3043243b5f170124968e7d118e7
- [40] Micro Commercial Components, "MBR2020CT Thru MBR20100CT," MBR20100CT product page, 2012. [Online].
http://61.222.192.61/mccsemi/up_pdf/MBR2020CT-MBR20100CT%28TO-220AB%29.pdf
- [41] C.J. Tseng and C.L. Chen. "A passive lossless snubber cell for nonisolated PWM DC/DC converters," *Industrial Electronics, IEEE Transactions on*, vol.45, no.4, pp.593-601, 1998.
- [42] A. Elasser and D. A. Torry, "Soft switching active snubbers for dc/dc converters," *IEEE Trans. Power Electron.*, vol. 11, pp. 710–722, 1996.
- [43] G. Hua and F. C. Lee, "Soft-switching techniques in PWM converters," *IEEE Trans. Ind. Electron.*, vol. 42, pp. 595–603, 1995.
- [44] R. L. Lin and F. C. Lee, "Novel zero-current-switching-zero-voltage switching converters," in *Proc. PESC'96*, pp. 438–442, 1996.

- [45] R. Streit and D. Tollik, "High efficiency telecom rectifier using a novel soft-switched boost based input current shaper," in *Proc. INTELEC'91*, pp. 720–726, 1991.
- [46] S. J. Finney, B. W. Williams, and T. C. Green, "The RCD snubber revisited," in *Conf. Rec. IEEE-IAS Annu. Meeting*, pp. 1267–1273, 1993.
- [47] M. Ferranti, P. Ferraris, A. Fratta, and F. Profumo, "Solar energy supply system for induction motors and various loads," presented at INTELEC'89, Paper 15.7, 1989.
- [48] N. Machin and T. Vescovi, "Very high efficiency techniques and their selective application to the design of a 70 A rectifier," in *Proc. INTELEC'93*, pp. 126–133, 1993.
- [49] J. Lambert, J. Vieira, L. C. Freitas M. Vilela, and V. Farias, "A boost PWM soft-single switched converter without stresses of voltage and current," in *Proc. APEC'96*, pp. 469–474, 1996.
- [50] X. He, B. W. Williams, S. J. Finney, Z. Qian, and T. C. Green, "New snubber circuit with passive energy recovery for power inverters," *Proc. IEE—Elect. Power Applicat.*, vol. 143, no. 5, pp. 403–408, 1996.
- [51] G. Carli, "Harmonic distortion reduction schemes for a new 100 A–48V power supply," in *Proc. INTELEC'92*, pp. 524–531, 1992.
- [52] M. Sadiku and C.K. Alexander, "Capacitors," in *Fundamentals of Electric Circuits*, 3rd Ed. Boston: McGraw-Hill, 2005, pp. 204.
- [53] G. Hollister, "Elliptical Machine DC-DC Converter for the Energy Harvesting From Exercise Machines Project," *Cal Poly Digital Commons*. 2011. [Online]. <http://www.digitalcommons.calpoly.edu/eesp/70/>
- [54] N.K. Lovgren, "Energy Harvesting From Exercise Machines: Forward Converters with a Central Inverter," *Cal Poly Digital Commons*. 2011. [Online]. <http://www.digitalcommons.calpoly.edu/theses/520/>
- [55] Linear Technology, "LT4356-1/LT4356-2 Surge Stopper," LT4356-1/LT4356-2 product page, 2010. [Online]. <http://cds.linear.com/docs/Datasheet/4356fa.pdf>
- [56] Lumex, "SML-LX1206SRC-TR datasheet," Lumex: LED & LCD Solutions: Product, 2012. [Online]. <http://www.lumex.com/specs/SML-LX1206SRC-TR.pdf>

- [57] IXYS Corporation, "GigaMOS™ Power MOSFET," IXFK230N20T product page, 2012. [Online]. <http://ixdev.ixys.com/DataSheet/DS100133%28IXFK-FX230N20T%29.pdf>
- [58] D. Gerke; W. Kimmel, "EMI and circuit components: Where the rubber meets the road." *Electronic Design News*. 2000. [Online]. http://www.edn.com/article/491125-EMI_and_circuit_components_Where_the_rubber_meets_the_road.php
- [59] Nippon Chemi-Con, "KHD251E156M99C0B00 datasheet," KHD251E156M99C0B00 product page, 2012. [Online]. <http://www.chemi-con.co.jp/e/catalog/pdf/ce-e/ce-sepa-e/ce-thd-e-110701.pdf>
- [60] CapSite 2009, "ESR," Introduction to Capacitors, 2012. [Online]. <http://my.execpc.com/~endlr/esr.html>
- [61] Microsemi, "High Voltage Schottky Diode," APT30S20BG product page, 2012. [Online]. http://www2.microsemi.com/datasheets/30S20B_S.pdf
- [62] Daycounter, Inc. Engineering Services, "Heatsink Temperature Calculator," Electronics Calculators, 2012. [Online]. <http://www.daycounter.com/Calculators/Heat-Sink-Temperature-Calculator.phtml>
- [63] Ohmite, "M Series Heatsinks: Heatsinks with clips for TO-264 and TO-247," MV-102-55E product page, 2012. [Online]. http://www.ohmite.com/cat/sink_m.pdf
- [64] Aavid Thermalloy, "TO-220 & TO-202 Heat Sinks," 531202B02500G - Board level heatsinks, 2012. [Online]. <http://www.aavid.com/sites/default/files/literature/Aavid-Board-Level-Heatsinks-Catalog.pdf#page=59>
- [65] Ohmite, "W Series Heatsinks For TO-220, TO-247, and TO-264 devices" WA-T247-101E product page, 2012. [Online]. http://www.ohmite.com/cat/sink_w.pdf
- [66] Aavid Thermalloy, "TO-220 & TO-202 Heat Sinks," 574502B00000G - Board level heatsinks, 2012. [Online]. <http://www.aavid.com/sites/default/files/literature/Aavid-Board-Level-Heatsinks-Catalog.pdf#page=45>
- [67] D. Rhodes; M. Davis-Marsh. "*Proper PCB Layout Techniques Maximize Power Module Performance*," 2010. [Online]. <http://i.cmpnet.com/powermanagementdesignline/2010/08/C0610posted.pdf>

- [68] S. Maniktala. "Proper Board Layout Makes Switchers Work Right the First Time." *EE Times*. 2002. [Online]. <http://www.eetimes.com/electronicsnews/4164323/Proper-board-layout-makes-switchers-work-right-the-first-time>
- [69] Maxim Integrated Products. "Correct Board Layout Lowers EMI of Switchmode Converters." *Maxim IC*. 2005. [Online]. <http://www.maxim-ic.com/appnotes/index.mvp/id/3645>
- [70] ExpressPCB, "Tips for Designing PCBs." *ExpressPCB - Free PCB Layout Software – Low Cost Circuit Boards - Top Quality PCB Manufacturing*. 2010. [Online]. <http://www.expresspcb.com/ExpressPCBHtm/Tips.htm>
- [71] Curious Inventor, "Tap & Drill Sizes." *Tools, Parts, Kits for DIY'ers - Curious Inventor*. 2012. [Online]. http://store.curiousinventor.com/guides/tap_drill
- [72] Axotron, "Crowbar circuit." *Articles about electronics*. 2012. [Online]. http://www.axotron.se/index_en.php?page=26
- [73] Circuitmaniac.com, "Electronic crowbar protection." *Power Plant*. 2012. [Online]. <http://www.circuitmaniac.com/2009/04/17/electronic-crowbar-protection/>
- [74] National Semiconductor (2007, May). *Application Note AN-1628: Minimizing FET Losses for a High-Input Rail Buck Converter* [Online]. Available: <http://web.archive.org/web/20070821192228/http://www.national.com/national/ed/may07/>
- [75] Linear Technology. *Datasheet – LTC3862 - Multi-Phase Current Mode Step-Up DC/DC Controller* [Online]. <http://cds.linear.com/docs/Datasheet/3862fb.pdf>
- [76] ExpressPCB. *Manufacturing Specs – Production Service* [Online]. Available: <http://www.expresspcb.com/ExpressPCBHtm/SpecsProduction.htm>