ENERGY HARVESTING FROM EXERCISE MACHINES:
FORWARD CONVERTERS WITH A CENTRAL INVERTER

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by
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ABSTRACT

Energy Harvesting From Exercise Machines: Forward Converters with a Central Inverter

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This thesis presents an active clamp forward converter for use in the Energy Harvesting From Exercise Machines project. Ideally, this converter will find use as the centerpiece in a process that links elliptical trainers to the California grid. This active clamp forward converter boasts a 14V-60V input voltage range and 150W power rating, which closely match the output voltage and power levels from the elliptical trainer. The isolated topology outputs 51V, higher than previous, non-isolated attempts, which allows the elliptical trainers to interact with a central grid-tied inverter instead of many small ones. The final converter operated at greater than 86% efficiency over most of the elliptical trainer's input range, and produced very little noise, making it a solid choice for this implementation.

Keywords: DC-DC Converter, Active Clamp Forward Converter Topology, Switch Mode Transformer, Energy Harvesting From Exercise Machines, Zero Voltage Switching, Precor Elliptical Machine

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CHAPTER 1: INTRODUCTION

1.1 Introduction

This chapter provides an overview of Energy Harvesting From Exercise Machines: Forward Converters with a Central Inverter. It establishes the project's context and summarizes the literature review, then covers the motivation for the project and its innovation. Section 1.6 provides the thesis statement, and the final sections lay out the design requirements and system framework.

1.2 Context

Some Precor EFX elliptical trainers come equipped with onboard generators that convert kinetic energy into electrical energy to power the unit's CPU and display (including, but not limited to, the 546i, 556i, and 576i) [1]. These components do not require much power for operation, and typically the user generates more electrical power than the machine needs. When this happens, the trainer feeds excess electricity through large resistors where it dissipates as heat. The process wastes a significant amount of energy, especially during more intense workouts. EHFEM: Forward Converters with a Central Inverter devises a system that salvages the wasted energy and sends it back to the local grid.

This project creates a system that converts excess energy generated by Precor elliptical trainers into grid-tied electricity. Most importantly, it designs a DC-DC converter that provides a link between the elliptical trainers and a commercial, grid-tied inverter. This thesis builds on the project established by former Cal Poly students, Michelle Lum and Jonathan Yuen, "Energy Harvesting from Exercise Machines: DC Converter Troubleshooting" and coincides with "Energy Harvesting From Elliptical Machines Using Four Switch Buck-Boost Topology" by Alvin Hilario [2-3]. This project aims to reduce net energy use at Cal Poly's Rec Center and raise awareness about renewable energy and energy consumption.

Figure 1-1: System Block Diagram [3]

This project aims to outfit each elliptical trainer with a DC-DC converter, with all of the DC-DC converters feeding into a central, grid-tied inverter.

This project differs from those devised by previous groups in that it establishes a system with one, central, grid-tied inverter, rather than an individual inverter for each exercise machine. This approach should reduce overall cost, complexity, and space requirements, although it also requires more robust DC-DC converters. ReRev, an EHFEM-type company out of Clearwater, Florida, uses this sort of approach in their Precor elliptical trainer retrofits [4-6].

1.3 Literature Review

This literature review breaks down into three sections: the first examines current EHFEM companies; the second describes some EHFEM implementations; the third provides background for DC-DC converter topologies, and specifically the active clamp forward converter.

1.3.1 EHFEM Companies

Two companies – ReRev and the Green Revolution, Inc. – already provide EHFEM services to fitness centers across the nation. This section outlines their work.

ReRev creates effective EHFEM systems through retrofit. At the time of this writing, the Florida-based company had already installed custom systems at twenty-eight different gyms and recreation centers across the country [4]. ReRev typically retrofits small fleets of Precor elliptical trainers, and then feeds the DC power through one or two central inverters [5-7]. ReRev founder, Harold Harr, prefers working with Precor elliptical trainers because they come equipped with a small electrical generator. Most other brands use alternators, which work more effectively with charging batteries [7]. The total cost of a ReRev system typically ranges from \$10,000-\$40,000 depending on the number of trainers involved [4, 6-7].

The Green Revolution, Inc. designs equipment that captures energy from stationary bikes and converts it into utility grade electricity. However, their website claims “The technology used with Indoor Cycling fitness equipment is designed to work with most cardio equipment, including, ellipticals, cross-trainers, stepping machines, stationary and recumbent bikes” [8]. The Green Revolution, Inc. boasts the ability to retrofit “all major brands of indoor cycles” [8].

1.3.2 Previous Implementations

EHFEM-type projects already exist at many fitness centers across the country [4, 8, 11]. ReRev has led many of these implementations, but some gyms have taken it upon themselves to create energy-generating exercise equipment.

Figure 1-2: ReRev Implementation at Middle Tennessee State University [5]

ReRev retrofitted 5 Precor elliptical trainers at Middle Tennessee State University's rec center in February 2010. According to a news report, each trainer generates roughly 1kWh every two days [5]; an average power output of ~30W per trainer during business hours, which average 16.71 hours/day [9].

In 2009, students at the University of Oregon and Oregon State University worked with ReRev to implement grid-tied EHFEM projects [7]. ReRev retrofitted 22 Precor elliptical trainers at Oregon State University in February of 2009, and 16 Precor elliptical trainers at the University of Oregon in May of 2009 [4]. The shared sports rivalry between the two schools has since bled over into exercise competitions, where the students compete to see who can generate the most electricity [7].

ReRev has also retrofitted dozens of other facilities, including CSU San Bernadino's Student Recreation Center (20 elliptical trainers), TSU San Marcos's Student Recreation Center (30 elliptical trainers), and Portland State University's Student Recreation Center (17 elliptical trainers) [4, 10].

The Green Microgym in Portland, Oregon employs electricity-generating elliptical trainers and stationary bikes along with human-powered treadmills [11]. The gym offers an “almost free” membership with no long term commitment, and has a goal of one day generating as much energy as it consumes [11].

1.3.3 Suitable DC-DC Converter Topologies

This project requires a grid-tied inverter capable of handling roughly three kilowatts. Inverters that size typically demand higher voltage inputs – in the area of 200V-600V [12]. With the Precor elliptical trainers only outputting between 7V and 65V, a large step-up conversion must take place to ensure proper interaction with the central inverter. To achieve that large step-up, this design focused on isolated DC-DC converter topologies, which utilize a switch mode transformer that can perform large step-up or step-down conversions given the proper transformer turns-ratio. Chapter 4 provides further insight into this decision process.

The isolated topology decision process relied heavily on DC-DC converter overviews provided in Dr. Taufik’s power electronics lecture notes [13-15]. The lecture notes cover the benefits, drawbacks, and typical uses of a handful of isolated topologies, including the flyback, push-pull, bridge, and forward. After comparing these topologies using decision matrices, the single-switch forward seemed like the best fit for this EHFEM application.

Abundant research exists for the single-switch forward converter, and the topology has seen uses in telecom, energy harvesting, power supplies, and a host of other areas. Most IC manufacturers produce high power forward converter chips, including Linear Technology, Texas Instruments, and On Semiconductor [16-20]. Additionally, many research papers explore the active clamp forward converter, a variation on the single-switch forward converter that uses a resonant capacitor and extra MOSFET to increase efficiency, expand duty cycle range, and reduce noise [21-24]. This design selected the active clamp forward converter for those reasons.

1.4 Motivation

EHFEM lowers the Rec Center's net energy consumption, but more importantly it conceptualizes energy units and promotes energy efficiency. Students can monitor their energy production via a digital display developed by former Cal Poly student, Brendan Asche [25]. Once students realize the amount of energy embedded in one kilowatt-hour, they may think more about their energy consumption at home, work, and across campus [6, 10]. As put by University of Oregon's Sustainability Director, Steve Mital, "We bought it and installed it mostly because it's an educational opportunity. People will be on those things sweating away and it gets them thinking...30 minutes later you have a deep visceral understanding of what (renewable energy) means. That's what I'm after, primarily" [7].

EHFEM also brings greater excitement to workouts. Gym goers can monitor watt-hours generated in the same way that they monitor calories burned. Also, "going green" has become trendy. "Oh, wow! It's awesome! That's cool!" exclaimed University of Oregon sophomore Eileen Donnerberg after finding out that her exercise generated electricity. "I never thought of that. It's a good thing" [7].

In summary, EHFEM raises awareness about renewable energy while conceptualizing common energy units, and it makes students more mindful of their energy consumption [5-6, 10]. It results in energy saved at the Rec Center and abroad through clever engineering and behavioral change. The process completely embodies Cal Poly's "learn by doing" mantra.

1.5 State-of-the-Art

Outsourcing this project to an EHFEM-based company like ReRev may provide a quick and simple solution to the energy harvesting problem, but keeping the project local opens the door for an economically desirable system, and enhances interest among the student body, faculty, and staff. According to Dr. Braun's economic assessment, limiting the cost of a ten

machine system to \$3,600 would ensure complete return on investment after ten years [26].

Over a dozen students have already completed EHFEM projects, laying the foundation for future students to build on. Staff members, like Sustainability Director Dennis Elliot, have joined the EHFEM ranks, providing valuable insight regarding the installation of such a project. In the end, Cal Poly's EHFEM projects hope to provide exercise equipment that Cal Poly's constituents can take pride in, not only because they save the university money and produce clean energy, but because they came from fellow alumni like themselves.

1.6 Thesis Statement

This thesis defends the development, design, and testing of an active clamp forward converter designed to feed a central, CEC-approved inverter. Together, these units transform waste energy from an elliptical trainer into marketable utility-grade electricity. Chapter 2 covers the Precor elliptical trainers chosen for this EHFEM implementation, and includes resistive load tests and output characterizations for each model. Chapter 3 selects an inverter from the California Energy Commission's approved list, provides the reasoning behind the selection, and addresses potential problems associated with the chosen inverter [12]. Chapter 4 delves into the converter design process, while Chapter 5 covers the physical construction and testing of the device. Finally, Chapter 6 concludes the thesis with a project summary and future implications.

1.7 Design Requirements

The success of this project hinges on several crucial requirements. First, the system must produce utility grade AC electricity under most workout conditions. The user has the option of twenty different workout intensity levels (referred to as "intensity levels" throughout most of this report), where higher levels indicate more strenuous workouts. Energy gets dumped through the resistive coils at intensity levels 2-20, so the design should capture and convert electricity at each of those levels. Second, the modifications to the elliptical trainer must not affect the user's

typical workout experience. This implies preservation of the normal motion of the elliptical trainer, no additional noise, and similar (preferably identical) appearance compared to the original model. Third, this completed project must come at no cost to Cal Poly or Associated Students Incorporated. Most importantly, this project must not pose any safety risks to users, ASI staff, maintenance workers, or the like.

This EHFEM system must also conform to each of the following codes: Underwriters Laboratories 1741, IEEE 1547, National Electric Code, PG&E's Electric Rule 21 [2]. In addition, the grid-tied inverter must come from the list of California Energy Commission-approved inverters [12].

1.8 System Framework

This EHFEM system comprises several key elements: the elliptical trainer, a custom DC-DC converter, a grid-tied inverter, and the building panel/grid connections. The subsections below briefly describe each element.

1.8.1 Elliptical Trainer

Each elliptical trainer comes with an onboard electrical generator, which converts the user's kinetic movement into electrical energy. Some of that electricity charges a 12V battery and powers the CPU and display, although most of it runs through resistive coils and dissipates as heat. This project replaces the resistive coils in each machine with an active clamp forward converter, which converts the otherwise wasted energy into steady DC power.

EHFEM must accommodate several different Precor elliptical trainer models. While each Precor model operates in a very similar fashion, the output voltage range differs from machine to machine. On top of that, some models output more power than others. The DC-DC converter design must account for these inconsistencies to ensure durability and reliability following implementation.

Figure 1-3: Precor EFX 546i Elliptical Trainer [27]

The Precor EFX 546i elliptical trainer serves as the main test module for this, and other, EHFEM elliptical trainer projects.

1.8.2 DC-DC Converter

The DC-DC converter transforms volatile electricity from the elliptical trainer into a constant DC voltage signal. This conversion provides steady input power to the inverter, which promotes both continuous operation and high efficiency. The DC-DC converter must effectively handle a 7V-65V input range, and should output a steady 51.2V DC signal (the nominal input value for chosen inverter) [28]. Resistive load tests indicate that the DC-DC converter must effectively handle 6.5A of input current and 350W of input power (maximum) [2]. The converter must exhibit excellent line and load regulation, high efficiency, and durability.

1.8.3 Grid-Tied Inverter

The grid-tied inverter transforms DC electricity from the DC-DC converter into utility-grade 60Hz electricity. The chosen inverter must conform to the California Energy

Commission's SB1 Guidelines, and it must possess a high enough power rating to accommodate ten elliptical trainers [12]. Chapter 3 explores the specifications of the grid-tied inverter in detail.

1.8.4 Building Panel / Grid Connection

The 60Hz AC electricity from the inverter ties in with the local grid through the Rec Center's building panel. This thesis project relies on Cal Poly's Facilities Department to install the necessary conduit and make the proper connections so that electrical energy can flow from elliptical trainers to the grid.

1.9 Chapter Summary

EHFEM: Forward Converters with a Central Inverter attempts to create a system that links elliptical trainers to the local grid. The design aims to limit costs while promoting a safe, quiet, and normal exercise experience. The following chapters track the entire design process chronologically, beginning with elliptical trainer characterizations, following with inverter selection, then DC-DC converter design and testing, and finally the conclusion.

CHAPTER 2: PRECOR ELLIPTICAL TRAINERS

2.1 Introduction

This chapter details the Precor EFX elliptical trainer models involved in this EHFEM implementation. Each trainer operates similarly; however, the slight variations in size, output voltage, output power, etc. required investigation before design of the DC-DC converter could commence. Sections 2.2 and 2.3 provide brief characterizations of two popular Precor EFX elliptical trainers, each of which see use in Cal Poly's Rec Center. Section 2.4 discusses the importance of maintaining constant load impedance for the elliptical trainer, a design constraint that requires much more attention in future elliptical EHFEM projects.

2.2 EFX 546i Elliptical Trainer

Paul Grierson of Advanced Fitness Products donated a Precor EFX 546i to Cal Poly several years ago, and since then the trainer has served as the primary test module for EHFEM elliptical trainer projects [2]. As seen in Figure 2-1, the Precor EFX 546i does not incorporate arm movement into the workout. Most of the elliptical trainers in Cal Poly's Rec Center feature sliding poles and skis for arm and leg movement.

Figure 2-1: Precor EFX 546i Test Module

This Precor EFX 546i elliptical trainer serves as the primary test module for elliptical EHFEM projects.

As with other Precor EFX models, the 546i comes with an onboard AC electrical generator that produces electricity for the 12V battery, display, CPU, and load resistors. A schematic posted on the inside of the EFX 546i's rear enclosure provides a block diagram of the internal circuitry.

Figure 2-2: Block Diagram of Precor Self-Powered Elliptical Trainers [27]

This diagram shows the internal electrical connections of the EFX 546i, 556i, and 576i. Under normal operation, kinetic motion spins a six phase AC generator, which feeds into the “lower PCA.” The lower PCA connects to the 12V battery, display, and load resistors in radial fashion.

In order to size the DC-DC converter, Martin Kou and I connected a power meter across the resistor terminals and measured the power output during a 120 stride/minute workout. We recorded voltage drop across the resistors at each of the twenty intensity levels.

Table 2-1 shows the resistive load test results for the EFX 546i. The table displays each test result separately. We each struggled to maintain a constant 120 stride/minute speed under higher intensity levels, which explains the inconsistent power outputs.

Table 2-1: Precor EFX 546i Resistive Load Test Results

Martin Kou and I connected a multimeter across the EFX 546i load resistor terminals and recorded the voltage drop at each intensity level. We strove to maintain 120 strides per minute, but that became quite difficult at higher intensity levels. Calculated power comes from the squared voltage measurement divided by load resistance (10Ω).

Intensity Level	Nick			Martin		
	Strides per Minute	Output Voltage [V]	Calculated Power [W]	Strides per Minute	Output Voltage [V]	Calculated Power [W]
1	124	0.03	0.00	130	0.03	0.00
2	130	8.08	6.73	130	9.27	8.86
3	130	10.56	11.50	130	11.42	13.44
4	126	12.97	17.34	130	13.01	17.45
5	130	15.66	25.28	130	15.43	24.54
6	130	16.75	28.92	130	17.41	31.25
7	129	20.23	42.19	130	20.08	41.57
8	128	21.86	49.26	130	22.39	51.68
9	129	25.72	68.20	130	25.09	64.90
10	130	27.15	75.99	130	29.16	87.66
11	128	31.11	99.78	130	30.15	93.71
12	124	31.33	101.19	130	33.16	113.36
13	132	34.81	124.92	130	33.24	113.91
14	126	37.45	144.59	130	37.07	141.67
15	128	38.77	154.96	130	37	141.13
16	125	40.37	168.01	120	33.89	118.41
17	115	34.32	121.43	90	27	75.15
18	98	37.20	142.66	90	32	105.57
19	90	39.52	161.01	90	29	86.70
20	80	29.81	91.61	80	20	41.24

Figure 2-3: Precor EFX 546i Resistive Load Test Results

This graph plots power dissipated through the resistive coils vs. workout intensity level for the EFX 546i elliptical trainer. Output power drops sharply around intensity level 17, most likely because stride rate declined. From intensity level 1 to intensity level 8, the output power increases linearly, and stride rates remain fairly steady across those levels. Further characterization may show that output power increases linearly across the entire intensity level range, with stride rate held constant. For further EFX 546i characterization, refer to Alvin Hilario’s “Energy Harvesting From Elliptical Trainers Using Four-Switch Buck-Boost Topology” [3].

Table 2-1 and Figure 2-3 show how much power the EFX 546i dissipates through its resistive coils at roughly 120 strides per minute. At higher stride rates, the trainer dissipates even more power through the large resistors. Michelle Lum and Jonathan Yuen reported power outputs of up to 350 watts with the EFX 546i [2], although it seems that output power rarely exceeds 200W under normal conditions.

Figure 2-4: EFX 546i Output Voltage Waveform [3]

This screenshot captures the EFX 546i's output voltage waveform at 120 strides per minute and intensity level 10. The $16V_{pp}$ ripple oscillates at 39.8kHz and strays from the $27.9V_{DC}$ average by 29%.

In later tests, Alvin Hilario discovered that the EFX 546i's output exhibits a large voltage ripple, shown in Figure 2-4 above. The ripple oscillates at roughly 30 kHz under most conditions, and the ripple amplitude increases linearly with DC output voltage. Adding extra capacitance to the input of the DC-DC converter dampens these oscillations. For more information on this voltage ripple, see "Energy Harvesting From Elliptical Machines Using Four-Switch Buck-Boost Topology" [3].

2.3 EFX 576i Elliptical Trainer

The EFX 576i model incorporates both arm and leg movement, promoting a more complete body workout. It also shows the user's generated power on the trainer's LED display. However, the displayed power read much higher than the output power in tests conducted with

ASI Equipment Technician, Trevor Hoover. Table 2.2 shows the results from a 576i resistive load test Trevor Hoover and I conducted.

Table 2-2: Precor EFX 576i Resistive Load Test Results

Trevor Hoover and I connected a multimeter across the EFX 576i load resistor terminals and recorded the voltage drop at each intensity level. Maintaining the 120 strides/minute rate proved much easier, thanks in large part to the 576i's ski poles. Using the voltage measurements, we calculated power dissipation through the load resistors.

Intensity Level	Strides per Minute	Output Voltage [V]	Displayed Power [W]	Calculated Power [W]	Max Displayed Power [W]
1	120	0.00	94.00	0.00	
2	120	6.28	101.00	3.72	
3	120	9.01	113.00	7.66	
4	120	10.90	118.00	11.21	
5	120	13.60	121.00	17.45	
6	120	16.22	129.00	24.82	
7	120	19.80	135.00	36.98	
8	120	20.10	136.00	38.11	
9	120	23.26	142.00	51.04	
10	120	26.30	152.00	65.25	
11	120	28.72	160.00	77.81	274.00
12	120	31.40	163.00	93.02	
13	120	32.69	171.00	100.81	
14	120	35.72	178.00	120.37	
15	120	38.22	181.00	137.81	309.00
16	120	40.02	191.00	151.09	
17	120	40.02	195.00	151.09	236.00
18	120	40.02	202.00	151.09	
19	120		203.00	0.00	
20	120		212.00	0.00	

This resistive load test does not include voltage readings for intensity levels 19 and 20. A narrow time window prevented us from recording those values, but one should assume that the output voltage would have remained at 40.02V at those intensity levels. Trevor Hoover indicated that the voltage level held constant at 40.02V once I exceeded intensity level 16.

The “Max Displayed Power” column shows the maximum power displayed by the EFX 576i at the three different intensity levels. This maximum power measurement does not reflect

the output power (i.e. resistive coil power) at 120 strides per minute; rather, it represents the maximum power on the display during a sprint at that intensity level.

Figure 2-5 highlights the discrepancy between the power dissipated across the resistive coils of the EFX 576i and the power read from the EFX 576i's display panel.

Figure 2-5: Precor EFX 576i Resistive Load Test Results

This graph plots power dissipated through the resistive coils and the power shown on the display vs. workout intensity level for the EFX 576i elliptical trainer. Stride rate remained nearly constant (at 120 strides per minute) throughout the entire test, resulting in the linear rise shown above.

The large discrepancy between power shown on the elliptical trainer's display and power dissipated through the resistive coils seems to indicate that a large portion of the user-generated electricity feeds the battery, CPU, and LED display. The EFX 576i trainers in Cal Poly's Rec Center also boast small television screens, which may or may not rely on the elliptical trainer's internal generator for energy.

2.4 Interaction with DC-DC Converter

Preserving the normal operation of the elliptical trainer hinges on the system's ability to present the elliptical trainer constant impedance over the entire range of stride rates and intensity levels. This calls for an active speed and intensity level tracking system, which could exist in the DC-DC converter, the grid-tied inverter, or a separate circuit. Unfortunately, this project realized the importance of this constraint far too late. Chapter 6 addresses the issue in more depth.

Ideally, the stride rate and intensity level tracking system(s) would present the elliptical trainer constant 10Ω impedance at all times. However, tests have shown that constant 20Ω and 30Ω impedances may also work. This seems to imply that the consistency of the impedance outweighs the magnitude of the impedance. However, increasing the elliptical load impedance led to higher voltage drops across the load resistors (in order to maintain similar power dissipation). This effect may cause damage to the elliptical trainer's internal circuitry. The tracking system should aim to maintain 10Ω to prevent damage to the elliptical trainer's circuitry and uphold the normalcy of operation.

Figure 2-6: EFX 546i Response to 20Ω and 30Ω Loads

These resistive load tests replaced the stock 10Ω resistive load with high power 20Ω and 30Ω loads. Stride rate remained nearly constant at 100 strides per minute over the entire intensity level range in each case. This plot shows that modeling the input to the DC-DC converter as a 20Ω or 30Ω load may work just as well as a 10Ω model.

A closer look into eddy current braking, which establishes the difference between intensity levels, may also prove valuable. Even if the converter-inverter system presents a constant impedance to the elliptical trainer, that may not provide normal elliptical trainer operation. Previous EHFEM groups have not presented in-depth research regarding this topic.

I highly recommend that future EHFEM elliptical trainer projects examine each of these two topics: constant impedance modeling/interfaces and eddy current braking. These topics could prove crucial in the elliptical-to-grid conversion, and hopefully future groups will find a way to incorporate each of these processes using the DC-DC converter, grid-tied inverter, or both.

CHAPTER 3: GRID-TIED INVERTER

3.1 Introduction

Chapter 3 outlines the decision process used to pick the grid-tied inverter for EHFEM: Forward Converters with a Central Inverter, and suggests future research surrounding this inverter. Section 3.2 outlines the specifications for the desired inverter. Section 3.3 examines and compares several CEC-approved inverters that best fit the specifications laid out in Section 3.2. Section 3.4 addresses concerns related to commercial solar inverters, and Section 3.5 concludes the chapter with a brief summary of the inverter selection process.

3.2 Goals and Constraints

This thesis aims to find a grid-tied inverter that communicates well with ten isolated DC-DC converters, each with varying output voltages and currents. This poses a fairly large challenge, especially considering the constant impedance modeling problem addressed at the end of Chapter 2. In this section, however, input voltage range and power capacity acted as the primary constraints. Section 3.5 addresses concerns related to constant impedance modeling and eddy currents.

Only a few constraints dictated the inverter selected for this project. First and foremost, the search only considers CEC-approved inverters, which the California Energy Commission lists on their website: <http://www.gosolarcalifornia.org/equipment/inverter.php> [12]. Next, the inverter needs to handle the maximum power input from 10 elliptical trainers. Assuming a 350W maximum output from each trainer and a DC-DC converter efficiency of ~85% yields a total inverter input power of roughly 3kW. In this case, input voltage range also plays a substantial role. While isolated topologies can ideally reach any output voltage, higher output voltages require larger transformer turns-ratios, and high frequency, high turns-ratio transformers proved very difficult to find. With an efficient, custom-built transformer, input voltage range

would not weigh so heavily, but commercial switch mode transformers rarely offer high turns-ratios and relatively high power handling capabilities (200W+). In this project, the design settled on a commercial switch mode transformer, which made low input voltage inverters more desirable than higher input voltage ones. Finally, this project searches for an inverter that meets each of the previous three constraints at the highest possible efficiency. To summarize, this implementation desires a ~3kW, low input voltage, CEC-approved, grid-tied inverter that boasts high efficiency.

3.3 Selecting the Inverter

Many CEC-approved inverters may work well in this EHFEM implementation. To make the decision process easier, Dr. Braun suggested creating a decision matrix. The decision matrix establishes a list of important design metrics, like input voltage range, power capacity, etc., and weighs each metric according to its importance in the overall system design. The decision matrix breaks down into two tables, one laying out the component specifications, and the other establishing how well the inverters meet the desired specifications. Table 3-1 describes how well each inverter meets this project's needs, as a satisfaction percentage. Table 3-2 provides the specifications for each inverter.

Table 3-1: Grid-Tied Inverter Decision Matrix: Satisfaction Percentages

Table 3-1 establishes how well each inverter meets this project's needs. The weight of each metric establishes its importance to the overall system.

		Beacon Power M4 Plus		Connect Renewable Energy CE 4000		Gridpoint Connect C36		Direct Grid Tech. DGM S460	
Criteria	Weight	Satisfaction	Weight	Satisfaction	Weight	Satisfaction	Weight	Satisfaction	Weight
Input Voltage Range	50%	70.0%	35.0%	60.0%	30.0%	80.0%	40.0%	58.0%	29.0%
Power Capacity	20%	90.0%	18.0%	85.0%	17.0%	85.0%	17.0%	20.0%	4.0%
Efficiency	15%	80.0%	12.0%	85.0%	12.8%	78.0%	11.7%	83.0%	12.5%
Other	5%	75.0%	3.8%	70.0%	3.5%	75.0%	3.8%	50.0%	2.5%
Overall Satisfaction		68.8%		63.3%		72.5%		48.0%	

Table 3-2: Grid-Tied Inverter Decision Matrix: Specifications [12]

Table 3-2 lists the exact inverter specifications for each category established in Table 3-1. Each grid-tied inverter comes from the CEC's approved list, and the decision matrix only considered inverters that possessed a reasonably low input voltage range.

Criteria	Beacon Power M4 Plus	Connect Renewable Energy CE 4000	Gridpoint Connect C36	Direct Grid Teck. DGM S460
Input Voltage Range	46.5V-58V (54.5 nominal)	54V-77V (62V nominal)	40V-60V (51.2V nominal)	55V-65V (60V nominal)
Power Capacity	4,000W	3,000W	3,037W	479W
Efficiency	91.50%	93.00%	91.00%	92.50%
Other	compact, wall mount, typical solar inverter with meter	compact, wall mount, typical solar inverter	large size, contains 10kWh of backup batteries, automatic grid disconnect	small size, operates efficiently at low power inputs

3.3.1 Gridpoint Connect C36

The Gridpoint Connect C36 best met the design qualities for this EHFEM project. On top of the low input voltage range and adequate power capacity, the Gridpoint Connect C36 also contains 10kWh of battery storage, which might act as a buffer between the DC-DC converters and the grid, providing a more consistent, reflected load for the elliptical trainers [29].

Figure 3-1: Gridpoint Connect C36 Grid-tied Inverter [29]

The Gridpoint Connect C36 inverter occupies a fair amount of space, but also provides 10kWh of battery storage, which could act as a crucial buffer between the elliptical trainers and the grid.

Table 3-3: Gridpoint Connect C36 Efficiency vs. Output Power [28]

This table shows the relationship between inverter efficiency and power level for the Gridpoint Connect C36. The comparison examines three different input voltages: 40V, the minimum input; 51.2V, the nominal input; and 60V, the maximum input.

		Power Level (%; kW)						Wtd
		10%	20%	30%	50%	75%	100%	
Input Voltage (Vdc)		303.73	607.46	911.19	1518.65	2277.98	3037.30	
Vmin	40	88.0	92.2	93.2	94.0	93.0	91.3	92.9
Vnom	51.2	82.1	88.5	90.7	91.8	91.6	89.7	90.9
Vmax	60	80.4	86.1	87.9	90.3	90.5	88.4	89.4

Figure 3-2: Gridpoint Connect C36 Efficiency vs. Output Power [28]

This graph plots inverter efficiency vs. power level for the Gridpoint Connect C36. The comparison examines three different input voltages: 40V, the minimum input; 51.2V, the nominal input; and 60V, the maximum input.

3.4 Possible Flaws and Concerns

Solar inverters may not suit the needs of grid-tied EHFEM projects. Many solar inverters, including the Enphase Microinverter, employ maximum power point tracking algorithms that aim to provide the highest efficiency per given input. This posed a problem in Alvin Hilario's and Greg Hollister's final tests: the Enphase Microinverter constantly varied its input impedance in search of the maximum power point, which reflected through their DC-DC converters and presented very inconsistent intensity levels to the elliptical user [3, 30]. Unfortunately, this realization came after the completion of the active clamp forward converter. This project's final system intends to connect the active clamp forward converter to the Gridpoint Connect C36 solar inverter, which may possess the same maximum power point tracking problem as the Enphase Microinverter.

Each of the ten elliptical trainers needs to see a constant 10Ω impedance to preserve normal operation, and finding a grid-tied inverter that controls current draw from each DC-DC

converter, which determines impedance seen by each elliptical trainer, may prove impossible. Utilizing a constant input impedance inverter, or possibly even batteries, and relying on the DC-DC converters to maintain the 10Ω impedance for each elliptical trainer, may provide the necessary load to the elliptical trainers. ReRev builds systems that only use one or two inverters (depending on the number of elliptical trainers), so a central inverter approach certainly seems plausible [5-7].

The problem may not lie in constant impedance modeling either. At this point, Cal Poly's EHFEM project possesses no in-depth information about the internal circuitry of Precor EFX elliptical trainers. ReRev may use a different strategy that intercepts generated energy before it reaches the elliptical trainer's resistive coils. More research into the internal workings of the Precor EFX 546i test module may provide answers to the inverter interfacing problem.

3.5 Chapter Summary

Since the grid-tied inverter must come from the CEC's approved list, it acts as a major limiting factor in this system. If the inverter utilizes any sort of algorithm, the DC-DC converter must account for it, and find a way to present constant impedance to the elliptical trainer at the same time. This adds a tremendous amount of complexity to an already difficult DC-DC converter design. A future senior project may focus on in-depth research into suitable inverters for this scenario. Also, enhancing this project's DC-DC converter design to best communicate with the central inverter could also prove very important. This selection process chose an inverter based on input voltage range and power handling capability, with no focus on the inverter's power draw algorithm.

CHAPTER 4: FORWARD CONVERTER DESIGN

4.1 Introduction

This chapter describes the design process for the active clamp forward converter. Section 4.2 explicitly lists the DC-DC converter's specifications. Section 4.3 provides some background on DC-DC converter topologies, both isolated and non-isolated, and then examines three variations of the forward converter. Section 4.4 details the operation of the forward converter, with and without active clamp, and Section 4.5 follows with the active clamp forward converter design in LT Spice. Together, these sections provide a detailed and chronological look at the design process for the DC-DC converter in EHFEM: Forward Converters with a Central Inverter.

4.2 Goals and Constraints

The DC-DC converter forms a crucial path between the elliptical trainer's output and the grid-tied inverter's input. The DC-DC converter must stabilize the volatile power produced by the elliptical trainer at a constant DC voltage (in this case, 51.2V) to allow for efficient and constant DC-AC transformation by the inverter. This implies the following design metrics:

- Input voltage range: $7V < V_{IN} < 65V$
- Power capacity: $P_{OUT} > 300W$
- Full load efficiency: $\eta > 83\%$
- Efficiency at all loads: $\eta > 40\%$
- Output voltage: $40V < V_{OUT} < 60V$; $V_{OUT} \text{ (nominal)} = 51.2V$ [28]
- Output voltage ripple: $\Delta V_{OUT} < 5\%$
- Input impedance: $Z_{IN} \approx 10\Omega$
- Dimensions: must fit inside EFX rear housing

The DC-DC converter should boast high efficiency, durability, simplicity, affordability, and quiet operation. Most of these attributes tie right into sustainable design. In order for this project to make sense, it must “pay for itself” not only economically, but also environmentally, socially, and ethically.

4.3 Converter Topologies

DC-DC converters exist in two major categories: isolated and non-isolated. The main distinction lies in the utilization of a transformer; isolated topologies make use of a switch mode transformer, while non-isolated topologies do not. Typically, isolated topologies make a better choice for high power or large step-up/step-down situations. Non-isolated topologies usually apply better to low power, low step-up/step-down scenarios, although they can reach higher power levels with careful engineering [3].

Tables 4-1, 4-2, and 4-3 display some basic isolated and non-isolated DC-DC converter topologies, along with their respective transfer functions, switch limits, and diode limits.

Table 4-1: Non-Isolated DC-DC Converter Topologies [31]

This table shows four common non-isolated DC-DC converter topologies: buck, boost, buck-boost, and SEPIC. Two of these topologies – the buck-boost and SEPIC – possess both step-up and step-down capabilities, meaning they could see use in this EHFEM project, as long as the inverter has a lower input voltage range (like the Enphase Microinverter) [2-3, 30].

Type of Converter	BUCK	BOOST	BUCK BOOST (Inverting)	SEPIC
Circuit Configuration				
Ideal Transfer Function*	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{t_{ON}}{T_P}\right) = D$	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{T_P}{T_P - t_{ON}}\right) = \frac{1}{(1-D)}$	$\frac{V_{OUT}}{V_{IN}} = -\left(\frac{t_{ON}}{T_P - t_{ON}}\right) = -\left(\frac{D}{1-D}\right)$	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{D}{1-D}\right)$
Drain Current*	$I_{Q1} (max) = I_{OUT}$	$I_{Q1} (max) = I_{OUT} \times \left(\frac{1}{1-D}\right)$	$I_{Q1} (max) = I_{OUT} \times \left(\frac{1}{1-D}\right)$	$I_{Q1} (max) = I_{OUT} \times \left(\frac{D}{1-D}\right)$
Drain Voltage*	$V_{DS} = V_{IN}$	$V_{DS} = V_{OUT}$	$V_{DS} = V_{IN} - V_{OUT}$	$V_{DS} = V_{IN} + V_{OUT}$
Average Diode Current*	$I_{D1} = I_{OUT} \times (1-D)$	$I_{D1} = I_{OUT}$	$I_{D1} = I_{OUT}$	$I_{D1} = I_{OUT}$
Diode Reverse Voltage*	$V_{D1} = V_{IN}$	$V_{D1} = V_{OUT}$	$V_{D1} = V_{IN} - V_{OUT}$	$V_{D1} = V_{OUT} + V_{IN}$

Table 4-2: Isolated DC-DC Converter Topologies [31]

This table shows four common isolated DC-DC converter topologies: flyback, forward, 2 switch forward, and active clamp forward. Any of these topologies could see use in this EHFEM implementation.

Type of Converter	FLYBACK	FORWARD	2 SWITCH FORWARD	ACTIVE CLAMP FORWARD
Circuit Configuration	Equations and Waveforms for Discontinuous Mode 			
Ideal Transfer Function*	$\frac{V_{OUT}}{V_{IN}} = D \times \sqrt{\frac{T_P \times V_{OUT}}{2 \times I_{OUT} \times L_P}}$	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = \left(\frac{N_S}{N_P}\right) \times D$	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = \left(\frac{N_S}{N_P}\right) \times D$	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = \left(\frac{N_S}{N_P}\right) \times D$
Drain Current*	$I_{Q1}(\max) = \left(\frac{V_{IN} \times t_{ON}}{L_P}\right)$	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$
Drain Voltage*	$V_{DS} = V_{IN} + V_{OUT} \times \left(\frac{N_P}{N_S}\right)$	$V_{DS} = 2 \times V_{IN}$	$V_{DS} = V_{IN}$	$V_{DS} = V_{IN} \times \left(\frac{1}{1-D}\right)$
Average Diode Current*	$I_{D1} = I_{OUT}$	$I_{D1} = I_{OUT} \times D$	$I_{D1} = I_{OUT} \times D$	$I_{D1} = I_{OUT} \times D$
Diode Reverse Voltage*	$V_{D1} = V_{OUT} + V_{IN} \times \left(\frac{N_S}{N_P}\right)$	$V_{D1} = V_{OUT} + V_{IN} \times \left(\frac{N_S}{N_P}\right)$	$V_{D1} = V_{OUT} + V_{IN} \times \left(\frac{N_S}{N_P}\right)$	$V_{D1} = V_{OUT} + V_{IN} \times \left(\frac{N_S}{N_P}\right) \times \left(\frac{1}{1-D}\right)$

Table 4-3: Isolated DC-DC Converter Topologies (Continued) [31]

This table shows four common isolated DC-DC converter topologies: half bridge, push pull, full bridge, and phase shift ZVT. Any of these topologies could see use in this EHFEM implementation.

Type of Converter	HALF BRIDGE	PUSH PULL	FULL BRIDGE	PHASE SHIFT ZVT
Circuit Configuration				
Ideal Transfer Function*	$\frac{V_{OUT}}{V_{IN}} = \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = \left(\frac{N_S}{N_P}\right) \times D$	$\frac{V_{OUT}}{V_{IN}} = 2 \times \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = 2 \times \left(\frac{N_S}{N_P}\right) \times D$	$\frac{V_{OUT}}{V_{IN}} = 2 \times \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = 2 \times \left(\frac{N_S}{N_P}\right) \times D$	$\frac{V_{OUT}}{V_{IN}} = 2 \times \left(\frac{N_S}{N_P}\right) \times \left(\frac{t_{ON}}{T_P}\right) = 2 \times \left(\frac{N_S}{N_P}\right) \times D$
Drain Current*	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$	$I_{Q1}(\max) = \left(\frac{N_S}{N_P}\right) \times I_{OUT}$
Drain Voltage*	$V_{DS} = V_{IN}$	$V_{DS} = 2 \times V_{IN}$	$V_{DS} = V_{IN}$	$V_{DS} = V_{IN}$
Average Diode Current*	$I_{D1} = (I_{OUT} \times D) + \frac{I_{OUT}}{2} \times (1-2D)$	$I_{D1} = (I_{OUT} \times D) + \frac{I_{OUT}}{2} \times (1-2D)$	$I_{D1} = (I_{OUT} \times D) + \frac{I_{OUT}}{2} \times (1-2D)$	$I_{D1} = \frac{1}{2} \times I_{OUT}$
Diode Reverse Voltage*	$V_{D1} = V_{IN} \times \left(\frac{N_S}{N_P}\right)$	$V_{D1} = V_{IN} \times \left(\frac{N_S}{N_P}\right) \times 2$	$V_{D1} = V_{IN} \times \left(\frac{N_S}{N_P}\right) \times 2$	$V_{D1} = V_{IN} \times \left(\frac{N_S}{N_P}\right)$

After careful consideration, this project chose an isolated approach, mainly due to the high input voltage requirements of the central inverter [28]. Isolated topologies should make a 7V-to-51.2V transformation much simpler, given a transformer with the proper turns-ratio.

4.3.1 Isolated Topologies

In order to generate grid-tied AC power over the elliptical trainer's entire operating range, this project requires a DC-DC converter with high step-up capabilities. Non-isolated topologies cannot safely deliver 51.2V at the output given only 7V in. For that reason, this situation calls for an isolated topology with a step-up transformer. This section examines a handful of isolated topologies, all of which could find use in this application. It compiles the isolated converters into a decision matrix, and eventually chooses the most appropriate design.

All isolated topologies derive their design from one of two non-isolated converters: buck and buck-boost. Buck-derived topologies include the forward, push-pull, and bridge. Their step-up capabilities rely entirely on the transformer turns-ratio. The flyback converter takes a buck-boost approach, storing energy in the transformer coils during on-time, and releasing the energy to the output during off-time. The flyback converter can easily step up input voltage using a 1:1 turns-ratio transformer. The flyback may seem like the most logical choice at first glance, due to its natural step-up ability, but further investigation reveals better options.

Table 4-4: Isolated Topology Decision Matrix: Satisfaction Percentage

Table 4-4 establishes how well each DC-DC converter topology meets this project's needs. The weight of each metric establishes its importance to the overall system.

		Push-Pull		Single-Ended Forward		Double-Ended Forward		Flyback		Half-Bridge		Full-Bridge	
Criteria	Weight	Satisfaction	Weight	Satisfaction	Weight	Satisfaction	Weight	Satisfaction	Weight	Satisfaction	Weight	Satisfaction	Weight
Input Voltage Range	25%	40.0%	10.0%	100.0%	25.0%	100.0%	25.0%	100.0%	25.0%	40.0%	10.0%	40.0%	10.0%
Power Range	20%	40.0%	8.0%	95.0%	19.0%	100.0%	20.0%	40.0%	8.0%	30.0%	6.0%	10.0%	2.0%
Switching Scheme	15%	50.0%	7.5%	100.0%	15.0%	95.0%	14.3%	100.0%	15.0%	50.0%	7.5%	30.0%	4.5%
Simplicity	15%	20.0%	3.0%	70.0%	10.5%	60.0%	9.0%	90.0%	13.5%	40.0%	6.0%	25.0%	3.8%
Typical Efficiency	12%	60.0%	7.2%	80.0%	9.6%	80.0%	9.6%	80.0%	9.6%	60.0%	7.2%	50.0%	6.0%
Size	8%	20.0%	1.6%	90.0%	7.2%	60.0%	4.8%	90.0%	7.2%	35.0%	2.8%	25.0%	2.0%
Other	5%	10.0%	0.5%	90.0%	4.5%	90.0%	4.5%	40.0%	2.0%	60.0%	3.0%	50.0%	2.5%
Overall Satisfaction		37.8%		90.8%		87.2%		80.3%		42.5%		30.8%	

Table 4-5: Isolated Topology Decision Matrix: Specifications

Table 4-5 lists the general topology specifications for each category established in Table 4-4.

Criteria	Push-Pull	Single-Ended Forward	Double-Ended Forward	Flyback	Half-Bridge	Full-Bridge
Input Voltage Range	50V-1000V	5V-500V	5V-500V	5V-500V	50V-1000V	50V-1000V
Power Range	100W-1000W	0-500W	0-500W	0-150W	100W-500W	400W-2000W+
Switching Scheme	two switches on alternate half-cycles	single switch	two synchronous switches	single switch	two switches on alternate half-cycles	two pairs of switches on alternate half-cycles
Simplicity	flux imbalance a big problem, 4 transformer windings, lots of parts	transformer reset can be difficult, common design choice	more parts than single-ended forward converter	few parts, common design choice, need proper transformer	flux imbalance can be a problem, 3 transformer windings	larger, more complex than half-bridge converter
Typical Efficiency	75%	78%	78%	80%	75%	73%
Size	large	medium	medium	small	large	large
Other	lots of transformer problems, all around poor fit for this application	meets most criteria, lots of documentation about this converter	meets most criteria, but includes more design work than single-ended forward converter	transformer causes difficulties, high output ripple, low power quality	not a good choice for low power situations	not a good choice for low power situations

Most of the data compiled in the decision matrices above came from Dr. Taufik's power electronics lecture notes [13, 15]. The data does not hold true for all converters under each topology; rather, it represents a general range. Tables 4-4 and 4-5 provide a comparative view of the trade-offs between isolated DC-DC converters, the weight of each design metric, and the relative satisfaction percentage for each topology.

After weighing the pertinent data, the single-ended forward converter fit the desired criteria best. The single-ended forward converter (a.k.a. single-switch forward converter) sees common use in both high and low power situations, and works very well in both step-up and step-down scenarios. On top of that, abundant research exists detailing effective ways to tackle the main concern with the single-ended forward topology: transformer reset [21-24]. Indeed, the single-ended forward converter seems like an excellent fit for this central inverter scenario.

4.4 Single-Switch Forward Converter

The following sections discuss the single-switch forward converter's background and operation, followed by important design considerations and common problems associated with the topology. It also presents an in-depth look at transformer reset techniques, with special focus on the active clamp.

4.4.1 Converter Operation

The single-switch forward converter acts essentially the same as a buck converter, except it utilizes a transformer in series with the main switch. While the buck converter relies solely on duty cycle to step down the input voltage, the forward design uses both duty cycle and transformer turns-ratio to allow both step-up and step-down functions. Ideally, the forward can reach a wide range of output voltages; however, one should place practical limits on the transformer's turns-ratio to ensure that cost, size, weight, and current and voltage magnitudes do not get out of hand.

Figure 4-1: Simplified Buck Converter Schematic [32]

This figure depicts a standard buck topology, which only sees use in step-down situations.

Figure 4-2: Simplified Single-Switch Forward Converter Schematic [32]

This figure depicts a traditional forward converter topology, which derives its architecture from the buck converter.

As noted above, the single-switch forward converter functions similarly to a buck converter. Whereas the buck converter relies solely on duty cycle (D) to control the output voltage, the single-switch forward utilizes both duty cycle and turns-ratio (N_S/N_P) to determine its output voltage, giving the converter both step-up and step-down capabilities. Equations 4-1 and 4-2, below, describe the ideal transfer functions for each of these converters in constant current mode (CCM).

Equation 4-1: Ideal Buck Converter Transfer Function

$$V_{OUT} = V_{IN} \times D$$

Equation 4-2: Ideal Single-Switch Forward Converter Transfer Function

$$V_{OUT} = (N_S / N_P) \times V_{IN} \times D$$

As mentioned above, these equations represent the ideal transfer functions for each converter; they do not account for conduction and switching losses through switches, diodes, or other components.

The buck and single-switch forward converters operate in similar fashions. When the switch turns on, current passes from input to output, traveling through an LC filter (which smoothes the choppy voltage and current waveforms) along the way. When the switch turns off, the output inductor reverses polarity and tries to keep current flowing to the output. The diode (labeled D₃ in Figure 4-2; unlabeled in Figure 4-1) becomes forward-biased and creates a current loop between inductor and capacitor. That LC tank discharges its stored energy to the output until the switch turns on once again. The amount of switch “on-time”, represented by duty cycle (D), determines the output voltage in each circuit.

The forward converter’s transformer adds a few wrinkles to the standard buck operation. First off, input current flows through the primary winding during on-time. Also, the primary winding sees the full voltage input, labeled V_g in Figure 4-2. The primary voltage and current waveforms get reflected to the secondary winding, such that V_{SECONDARY} = (N_S / N_P) × V_{IN} and I_{SECONDARY} = (N_P / N_S) × I_{IN}. (Referring to Figure 4-2, this translates to V_{SECONDARY} = (n₃ / n₁) × V_g and I_{SECONDARY} = (n₁ / n₃) × I_{IN}.) Diode D₂, often referred to as the rectifying diode, conducts during on-time, allowing current to flow from the secondary winding through the LC filter to output. When the switch turns off, the LC filter supplies energy to the output, just as it does in

the buck converter. However, the transformer stored energy during on-time, and its magnetizing inductance requires a current path during off-time. In traditional single-switch forward designs, like the one pictured in Figure 4-2, a third winding in series with a diode allows the stored energy to discharge back to the input. Newer single-switch forward topologies offer more practical solutions to the magnetizing energy problem. This design opts for an active clamp to relieve the stress caused by disrupted magnetizing energy.

4.4.2 Considerations and Common Problems

This section addresses several common issues that may crop up during the single-switch forward converter design process. First, the forward converter typically sees use in step-down applications, but this situation calls for a step-up conversion. To that point, step-down forward converters receive relatively high input voltages and low input currents, but this application demands the opposite. This scenario also presents a large input voltage range (7V-65V), which requires a wide duty cycle range. Volt-second balance must ensure that the converter can accommodate those input voltages. Also, the transformer's turns-ratio must allow V_{OUT} to reach at least 40V (ideally 51.2V) given the minimum input voltage. Finally, the design must incorporate safe storage and relief of magnetizing energy to prevent large voltage and current spikes.

The forward converter acts as an isolated buck converter, so it makes sense that, conceptually, designers would choose it for step-down applications. In order for step-up to take place, the converter must utilize a transformer with more secondary turns than primary turns ($N_S > N_P$). In this scenario, input current exceeds output current by a factor of N_S/N_P . Under full load conditions, this can lead to uncomfortably large input currents and high conduction losses. One can account for the high primary-side currents through proper component selection and PCB design.

Ensuring acceptable voltage regulation requires proper duty cycle assignments for minimum and maximum input voltages. This holds true for all DC-DC converters, and a quick look at Equations 4-1 and 4-2 shows an inversely proportional relationship between D (duty cycle) and V_{IN} . Volt-second balance refers to the calibration of duty cycle range given input voltage range. Rearranging the transfer function from Equation 4-2 provides the minimum and maximum duty cycle equations that follow.

Equation 4-3: General Duty Cycle Calculation for Single-Switch Forward Converter

$$D = \frac{V_{OUT}}{(N_S/N_P) \times V_{IN}}$$

Equation 4-4: Minimum Duty Cycle Calculation for Single-Switch Forward Converter

$$D_{(MIN)} = \frac{V_{OUT}}{(N_S/N_P) \times V_{IN(MAX)}}$$

Equation 4-5: Maximum Duty Cycle Calculation for Single-Switch Forward Converter

$$D_{(MAX)} = \frac{V_{OUT}}{(N_S/N_P) \times V_{IN(MIN)}}$$

$D_{(MIN)}$, the minimum duty cycle, occurs at the maximum input voltage. Likewise, $D_{(MAX)}$ occurs at the minimum input voltage. Together, they define the duty cycle thresholds. To complete these calculations, one must first choose a transformer turns-ratio that allows for the minimum input voltage to reach the correct output voltage. Assuming a maximum duty cycle of 80% ($D_{(MAX)} = 0.80$) makes this first step much simpler.

Equation 4-6: Transformer Turns-Ratio Calculation for Single-Switch Forward Converter

$$N_S/N_P = \frac{V_{OUT}}{D_{(MAX)} \times V_{IN(MIN)}} = \frac{40V}{0.80 \times 7V} \cong 7.14$$

Equation 4-6 shows that the converter requires a greater than 7:1 turns-ratio transformer in order to produce an output voltage that can communicate with the Gridpoint Connect C36 inverter across a 7V-65V input range [28].

Voltage and current spikes occur when switching interrupts the transformer's leakage current. These spikes can grow dangerously large and cause significant damage to the main switch, rectifying diode, input stage, and other circuit components. Several methods exist for mitigating these dangerous spikes and resetting the transformer, and the following section addresses three of them: reset winding, RCD clamp/snubber, active clamp. Figure 4-3 shows the sort of voltage spikes (across the switch) that can occur when no path exists for transformer magnetizing current during off-time.

Figure 4-3: Magnetizing Voltage Spikes [33]

Voltage spikes like these can occur when the transformer's magnetizing current gets interrupted. These waveforms show the voltage drop across the forward converter's main switch when no reset path exists.

4.4.3 Clamp and Reset Techniques

Three common methods exist for establishing a reset path and limiting turn-off spikes: reset winding, RCD clamp and reset, and active clamp. This section discusses each method

separately, with extra attention paid to the active clamp, which ultimately sees use in this forward converter design.

Figure 4-4: Single-Switch Forward Converter with Reset Winding [21]

Traditional forward converters use a reset winding in series with a diode to allow magnetizing current to flow during off-time. This design comes with several drawbacks, including limited duty cycle range, increased size, and reduced efficiency compared to more modern techniques.

The traditional clamp and reset method utilizes a third transformer winding, which discharges magnetizing current to the converter's input capacitor during off-time. A diode placed in series with the third winding prevents current flow during on-time. At switch turn-off, the voltage across each winding reverses, which causes the reset diode to conduct. Hence, a path forms between reset winding and input capacitor, allowing continuous current flow across ON and OFF states [15, 21].

The reset winding has several drawbacks. First, employing a third transformer winding increases the converter's size, weight, and cost. The third winding also limits maximum duty cycle to 50%. Above that value, the transformer cannot completely reset [21]. With $D_{(MAX)}$ set to 50%, the converter struggles to accommodate wide input voltage ranges; very narrow duty

cycles occur under high input voltages. These problems make the reset winding a poor choice for interaction with an elliptical trainer.

Figure 4-5: Single-Switch Forward Converter with RCD Clamp [21]

This dissipative clamp and reset design allows for larger duty cycle range and often times better efficiency than the reset winding technique. Magnetizing current flows into the clamp capacitor during off-time, then dissipates through the parallel resistor.

The RCD clamp and reset method improves upon the reset winding method by reducing transformer size, reducing overall cost and weight, and increasing maximum duty cycle. However, these improvements come at the cost of added complexity, increased voltage stress, and occasionally lower efficiency [21].

The RCD approach stores magnetizing energy in a capacitor during off-time, which then discharges through a parallel resistor during on-time. The design must ensure resistor and capacitor sizes that allow for complete energy dissipation during on-time, but also complete energy storage during off-time. To accommodate large duty cycles and high frequencies, small capacitor and resistor values make the most sense. However, reducing the size of the capacitor

increases the peak voltage at turn-off, and reducing the size of the resistor increases power losses [21]. Thus, the designer typically must choose between efficiency and input voltage range.

Figure 4-6: Single-Switch Forward Converter with Active Clamp [21]

The clamp and reset technique operates much like the RCD clamp, but offers improved efficiency by eliminating the dissipative resistor. The high-side MOSFET introduces extra complexity, but the benefits of this technique typically outweigh the extra design time.

The active clamp approach offers effective voltage spike reduction, improved efficiency, and high duty cycle limit. Instead of discharging the capacitor through a resistor during the main switch's on-time, a clamp switch (Q_C) allows the magnetizing energy to flow into the reset capacitor (C_R) and then back to the primary coil during off-time. If implemented properly, the active clamp allows for zero voltage switching (ZVS) at the main switch (Q_A), which greatly reduces switching component stress and power losses [21-23].

Figure 4-7: Active Clamp Forward Converter with Parasitic MOSFET Elements [21]

The active clamp utilizes each MOSFET's parasitic body diodes and stray capacitances to produce continuous current flow and improve transformer operation [22].

Before explaining the operation of the active clamp forward converter, one must see a more accurate depiction of the circuit. Figure 4-7 shows an active clamp forward converter with parasitic MOSFET elements. The reoriented schematic may appear entirely different than the active clamp forward converter shown in Figure 4-6, but closer investigation reveals a capacitor and diode across each of the FET drain-source junctions as the only discrepancies. These parasitic elements prove crucial in explaining the operation of this transformer reset technique.

Figure 4-8: Active Clamp Forward Converter: Power Stage [21]

During this stage, current flows from source to load directly through the transformer. The main FET (Q_A) and rectifying diode ($D1$) both conduct during this interval.

Figure 4-9: Active Clamp Forward Converter: Linear Transition Stage (1) [21]

The power stage concludes with Q_A turning off. Magnetizing current continues flowing through $D1$, and also through the two parasitic capacitances, C_A and C_C . C_A rapidly charges during this stage, while C_C discharges.

Figures 4-8 and 4-9 show the first two stages of operation. During the power stage, the input delivers voltage and current to the primary winding of the transformer. The main switch (Q_A) conducts during this period, providing a low resistance path for the input current. When Q_A turns off, the converter enters the linear transition stage. Magnetizing current charges C_A and discharges C_C , raising the voltage potential across the drain-source of Q_A and dropping it across Q_C . This transition typically lasts only a few nanoseconds, and occasionally gets ignored in active clamp design papers [22].

Figure 4-10: Active Clamp Forward Converter: Linear Transition Stage (2) [21]

In this second segment of the linear transition stage, C_A approaches its peak voltage while C_C nears zero volts. Current begins to enter the clamp MOSFET's body diode (D_C), prepping Q_C for zero voltage turn-on.

Figure 4-11: Active Clamp Forward Converter: Passive Reset Stage [21]

Q_C turns on with nearly zero volts across its drain-source junction, thanks to the conduction of D_C . Magnetizing current now flows into the reset capacitor (C_R), charging it.

As the converter moves from the linear transition to passive reset stage, C_A reaches its peak voltage and D_C begins to conduct and feed current to the reset capacitor, C_R . With D_C conducting, the voltage across the drain-source of Q_C dives close to zero, prepping the switch for zero voltage turn-on. Once Q_C turns on, the magnetizing current obtains a very low impedance path from the primary winding to C_R . During the passive reset stage, C_R charges to its peak voltage as current into the capacitor slowly diminishes. C_R 's capacitance determines the magnitude of the peak voltage across the main switch during off-time, since $V_{DS\ Q_A} = V_{IN} + V_{CR}$.

Figure 4-12: Active Clamp Forward Converter: Active Reset Stage [21]

Once the reset capacitor reaches its peak voltage, the current reverses direction and begins charging the primary winding.

Once C_R reaches its peak voltage, it begins discharging back into the primary winding. The voltage across Q_A 's drain-source descends back toward V_{IN} as the voltage across C_R declines in a resonant fashion.

Figure 4-13: Active Clamp Forward Converter: Resonant Transition Stage (1) [21]

Once the magnetizing current establishes momentum back through the primary winding, the clamp MOSFET (Q_C) turns off. This causes the magnetizing inductance to create a path through the parasitic capacitances, discharging C_A and C_R and charging C_C and C_{IN} .

Figure 4-14: Active Clamp Forward Converter: Resonant Transition Stage (2) [21]

Eventually, the magnetizing current finds a path through the main MOSFET's body diode (D_A). Once D_A turns on, it sets Q_A 's drain-source voltage to nearly zero volts ($-V_{D(ON)}$), prepping the switch for zero-voltage turn-on.

Soon, the clamp switch (Q_C) turns off once again and forces the resonant current through C_C . The voltage across Q_C rises as the voltage across the transformer's primary winding begins to collapse. Magnetizing current continues flowing around the circuit, charging C_{IN} while discharging C_A . As the voltage across the primary winding approaches zero, the voltage at the drain of the main switch (Q_A) descends toward V_{IN} . This trend continues, causing negative voltage build-up across the primary winding, which forces $V_{DS\ Q_A}$ to fall below V_{IN} . Ideally, the primary winding voltage descends all the way to $-V_{IN}$, such that the voltage at the drain of Q_A dips just below 0V. Immediately after this happens, D_A turns on, clamping the voltage across the drain-source of Q_A at $-V_{D\ ON}$ and allowing for zero voltage switching at turn-on [21-22].

Figure 4-15: Active Clamp Forward Converter: Circulation Interval [21]

In the circulation interval, D_A fully conducts and clamps Q_A to essentially zero volts. Following this stage, Q_A turns on and the process repeats.

All three of these reset techniques provide an adequate path for magnetizing energy during off-time, but the active clamp offers the highest efficiency, largest duty cycle range, and

greatest voltage stress reduction. These benefits make it the best choice for EHFEM: Forward Converters with a Central Inverter.

4.5 Designing the Active Clamp Forward Converter in LT Spice

4.5.1 Linear Technology

Linear Technology may produce the best power electronics simulation software, and they offer it to the general public free of charge. LT Spice allows users to simulate any Linear Technology IC in personalized power electronic schematics, providing an in-depth look at the chip's predicted operation for any number of conditions [33]. Users can tweak operating temperatures, implement near-real component models, perform math functions on simulated waveforms, and execute a wide array of other specialized functions. LT Spice made Linear Technology chips the first choice in this implementation.

Linear Technology manufactures at least four different forward converter control chips, two for single-switch topologies: LT1950, LT1952; and two for double-switch topologies: LT1681, LT3781 [16-17, 34-35]. Both single-switch chips boast power ranges up to 500W, more than enough to handle the output power of the elliptical trainer. The following subsections describe the active clamp forward converter design process in LT Spice. This design utilizes the LT1952-1: a variation of the LT1952 that offers a lower input voltage turn-on threshold (7.75V vs. 14.25V) [17].

4.5.2 Early Calibrations

Each LT integrated circuit chip comes with an LT Spice macromodel [33]. Figure 4-16 shows the LT1952-1's macromodel schematic. This circuit acted as the starting point for the active clamp forward converter design. The LT1952-1 macromodel converts a 20V DC input into a 12V DC output via a 1:1 transformer. Obviously, these numbers differ greatly from the

7V-65V input, 51.2V output that the project calls for. As a result, the first adjustments aimed to correct the input and output voltages.

Choosing large resistor values, like those found in the macromodel, minimizes power loss at the output, thus boosting efficiency. To keep things simple, R_8 remained at $4.99k\Omega$ and R_7 increased to $205k\Omega$ to meet the desired voltage divider ratio.

Before moving forward, we must take a closer look at Equation 4-6 from section 4.4.3 (page 36):

Equation 4-6: Transformer Turns-Ratio Calculation for a Single-Switch Forward Converter

$$N_S/N_P = \frac{V_{OUT}}{D_{(MAX)} \times V_{IN(MIN)}} = \frac{40V}{0.80 \times 7V} \cong 7.14$$

The $D_{(MAX)}$ and $V_{IN(MIN)}$ values require a bit more explanation, as does the calculated turns-ratio (N_S/N_P). Many design handbooks suggest limiting maximum duty cycle to 80% to allow the transformer enough time to reset and prevent saturation [15, 21]. Ideally, this project's DC-DC converter would output power over the entire 7V-65V range; however, several factors impede lower voltage conversion. First, the LT1952-1 has a 7.75V turn-on limit, meaning that the chip will not turn on for inputs below 7.75V (assuming the input voltage directly drives the chip's V_{in} pin) [17]. Second, a 7V output from the elliptical trainer only occurs under very light exercise conditions, and generates very little power (~5W). Accommodating the 7V input might require an entire extra turn on the transformer's secondary, increasing size and weight and possibly reducing switching frequency. With these considerations in mind, a 7:1 turns-ratio seemed most sensible. (Note that in order to adjust the turns-ratio in LT Spice, one must compare square roots of the winding inductances: $(L_{PRI} / L_{SEC})^{0.5}$.)

With the new transformer and resistor values, the converter delivers close to 51V for a wide range of inputs: $17V < V_{IN} < 70V$. Unfortunately, the converter delivers 0V when $V_{IN} < 17V$, and the circuit shows very poor load regulation. Figures 4-17 and 4-18 show the output voltage and output current waveforms at $V_{IN} = 17V$ and $V_{IN} = 70V$.

Figure 4-17: Input and Output Voltages for Early 1952-1 Forward Converter Design (17V In) [33]

The green waveform represents V_{OUT} (52V_{DC}), and the blue waveform represents V_{IN} (17V_{DC}). The x-axis represents time in milliseconds.

Figure 4-18: Input and Output Voltages for Early 1952-1 Forward Converter Design (70V In) [33]

The green waveform represents V_{OUT} (52V_{DC}), and the blue waveform represents V_{IN} (70V_{DC}). The x-axis represents time in milliseconds.

The LT1952-1 datasheet clearly states that the V_{in} pin cannot exceed 25V. Thus, a direct tie between the V_{in} pin and the system input voltage (V_{IN}) would not work. Example circuits in

the datasheet suggest placing a resistor in series with a zener diode (where $7.75V < V_Z < 25V$) to set a voltage limit for the V_{in} pin. A capacitor connected in parallel with the zener diode helps regulate voltage at the V_{in} pin [17].

The LT1952-1 requires 4.5mA of continuous current into V_{in} during operation. While the resistor-zener network (Formed by R_{START} , D_{IN} , and C_{START} in Figure 4-19) provided enough current to the chip during start-up, it could not source 4.5mA once the transformer began to conduct. A current path from output back to V_{IN} solves this problem. A series diode-resistor network (formed by D_{SUPPLY} and R_{SUPPLY} in Figure 4-19) provides a unidirectional path from the output to the LT1952-1 V_{in} pin. Figure 4-19 shows the forward converter schematic with the new feedback supply.

Figure 4-19: LT1952-1 Forward Converter with V_{in} Pin Protection and Feedback Supply [33]

A 15V zener diode and series resistor create a 15V clamp at the LT1952-1's V_{in} pin. The parallel capacitor (C_{START}) helps regulate voltage at the node below 15V. The feedback path from OUT ensures that the V_{in} pin receives enough power to keep the chip on.

The feedback path and V_{in} protection method shown in Figure 4-19 allow the converter to supply rail voltage to the LT1952-1 chip and output DC power using a single source: the 7V-

65V input from the elliptical trainer. However, this method also comes with some drawbacks. The output voltage from the elliptical trainer must exceed $\sim 8.5\text{V}$ for the chip to turn on. The added resistors reduce converter efficiency and may get pushed above their power limits under high input voltages. Also, driving the chip's rail supply with a variant source doesn't ensure consistent chip operation. For these reasons, the design opted to use the elliptical trainer's 12V battery to drive the LT1952-1 chip, instead of the feedback path and start-up network shown in Figure 4-19. Still, this alternate method could see use in future implementations, especially if the trainer's 12V battery proves inaccessible. The V_{in} feedback path and start-up network do not appear in upcoming schematics; a 12V source representing the elliptical trainer's onboard battery appears instead.

Another factor also limits chip turn-on: the SD_V_{sec} pin. This pin requires at least 1.32V in conjunction with 7.75V at the V_{in} pin in order for the chip to turn on [17]. In previous tests, the default 178k Ω -17.8k Ω voltage divider ensured that SD_V_{sec} would not surpass 1.32V until the input voltage exceeded 14.52V. Adjusting these resistor values allows the LT1952-1 forward converter to accommodate lower input voltages; unfortunately, it comes at the expense of limiting the maximum input voltage. While the LT1952-1 datasheet suggests using a simple resistive voltage divider to drive SD_V_{sec} , this project's wide input voltage range prohibits that simple option because it would exceed SD_V_{sec} 's voltage range: $1.23\text{V} < SD\_V_{sec} < 6\text{V}$ [17]. The following equations show that a simple resistive network cannot provide for the elliptical trainer's full input range. Let R_{SD1} represent the top resistor in the voltage divider and R_{SD2} represent the bottom resistor in the network that drives SD_V_{sec} .

Equation 4-8: LT1952-1 SD_Vsec Voltage Reference Calculations – Purely Resistive

$$SD\_Vsec = V_{IN} \times \frac{R_{SD2}}{R_{SD1} + R_{SD2}}$$

$$\frac{R_{SD1}}{R_{SD2}} = \frac{1 - SD\_Vsec/V_{IN}}{SD\_Vsec/V_{IN}} = \frac{V_{IN} - SD\_Vsec}{SD\_Vsec} = \frac{V_{IN}}{SD\_Vsec} - 1$$

$$\frac{R_{SD1}}{R_{SD2}} = \frac{V_{IN(MIN)}}{SD\_Vsec(MIN)} - 1 = \frac{7V}{1.32V} - 1 \cong 4.30$$

$$SD\_Vsec(MAX) = V_{IN(MAX)} \times \frac{1}{R_{SD1}/R_{SD2} + 1} = 65V \times \frac{1}{4.30 + 1} \cong 12.26V > 6V$$

To ensure that SD_Vsec remains within its limits for the full 7V-65V input range, this design called upon a diode to provide a small “DC offset.” The diode’s forward voltage drop acts as a near constant DC source, meaning that R_{SD2} only requires a 1.32V – V_{D(ON)} voltage drop at the minimum input instead of the entire 1.32V. This diode boosts the ceiling for the input voltage range such that the chip can remain on with 65V at the input. The following equations use 0.9V as the forward voltage drop of the series diode (henceforth referred to as D_{SD}), which closely resembles the average voltage drop of the diode selected for the actual design [36].

Figure 4-20: Adjusted SD_Vsec Network

The forward voltage drop of D_{SD} actually increases the converter's operable input voltage range by ensuring that the voltage at SD_Vsec always remains within its limits ($1.32V < SD\_Vsec < 6V$) [17].

Equation 4-9: LT1952-1 SD_Vsec Voltage Reference Calculations – Resistors Plus Diode

$$SD\_Vsec = (V_{IN} - V_{D(ON)}) \times \frac{R_{SD2}}{R_{SD1} + R_{SD2}} + V_{D(ON)}$$

$$\frac{R_{SD1}}{R_{SD2}} = \frac{1 - (SD\_Vsec - V_{D(ON)}) / (V_{IN} - V_{D(ON)})}{(SD\_Vsec - V_{D(ON)}) / (V_{IN} - V_{D(ON)})} = \frac{V_{IN} - SD\_Vsec}{SD\_Vsec - V_{D(ON)}}$$

$$\frac{R_{SD1}}{R_{SD2}} = \frac{V_{IN(MIN)} - SD\_Vsec_{(MIN)}}{SD\_Vsec_{(MIN)} - V_{D(ON)}} = \frac{7V - 1.32V}{1.32V - 0.9V} \cong \mathbf{13.53}$$

$$SD\_Vsec_{(MAX)} = (65V - 0.9V) \times \frac{1}{13.53 - 1} + 0.9V \cong 6.00V$$

Hence, the series diode maintains $1.32V < SD\_Vsec < 6.00V$ for $7V < V_{IN} < 65V$.

However, this only ensures that the chip stays on for the entire input voltage range; it doesn't ensure that the chip will function properly. The LT1952-1 datasheet describes how the SD_Vsec and SS_MAXDC pins determine maximum duty cycle, and provides design equations for

calibrating the duty cycle limit across the entire input voltage spectrum [17]. The equations and graphs below compare the ideal duty cycle (given by the single-switch forward converter's transfer function) with the duty cycle limit dictated by the voltages at SD_Vsec and SS_MAXDC.

Equation 4-10: LT1952-1 Maximum Theoretical Duty Cycle Calculation (Full Form) [17]

$$Max\ Duty\ Cycle = k \times 0.522 \times \left(\frac{SS\_MAXDC}{SD\_Vsec} \right) - t_{Delay} \times f_{osc}$$

where

$$k = 1.11 - 5.5e^{-7} \times f_{osc}$$

$$t_{Delay} = R_{Delay} \times 10^{-12} \frac{seconds}{\Omega}$$

$$f_{osc} = \frac{4100}{1 + R_{osc}/9125}$$

$$SS\_MAXDC = 2.5V$$

Equation 4-10 deserves more explanation. First off, the converter's switching frequency (f_{osc}) can range from roughly 100kHz to 500kHz, but in this design it retained the default value of 200kHz (corresponds to $R_{osc} = 178k\Omega$). The design considered other frequencies, but in the end 200kHz best met the needs of the switch mode transformer and active clamp. The time delay between SOUT and OUT (t_{Delay}) will receive more explanation later, but for now let it equal 237ns (corresponding to $R_{Delay} = 237k\Omega$). Designers can reduce the value of SS_MAXDC

to limit maximum duty cycle, but in this design SS_MAXDC retained its default (and maximum) value of 2.5V. With these settings, the max duty cycle equation reduces to:

Equation 4-11: LT1952-1 Maximum Theoretical Duty Cycle Calculation (Reduced Form) [17]

$$\text{Max Duty Cycle} = \frac{1.305V}{SD\_Vsec} - 0.0474$$

The LT1952-1 datasheet also states that “the maximum duty cycle clamp calculated in (Equation 4-10/4-11) should be programmed to be 10% greater than the maximum operational duty cycle calculated” [17]. Thus,

Equation 4-12: LT1952-1 Maximum Operational Duty Cycle Calculation (Reduced Form) [17]

$$\text{Max Operational Duty Cycle} = \frac{1.305V}{SD\_Vsec} - 0.1474$$

Maintaining $R_{SD1} / R_{SD2} = 13.53$, the ratio calculated in Equation 4-9, and assuming D_{SD} has a constant 0.9V forward voltage drop eliminates the remaining unknowns from the maximum duty cycle equation. Plotting the maximum operational duty cycle and desired duty cycle (determined using Equation 4-3) with a 7:1 step-up transformer produced the graph in Figure 4-21.

Figure 4-21: Max Duty Cycle and Ideal Duty Cycle vs. Input Voltage

The ideal duty cycle, calculated from the forward converter transfer function, must remain below the LT1952-1's duty cycle limit to provide safe and correct operation. At 8V (in) and below, the converter cannot achieve the ideal duty cycle calculated by the transfer function. As a result, the converter may not produce the desired output voltage (51.2V) for the bottom edge of the input voltage range.

Leaving R_{SD1} at 178k and changing R_{SD2} to 13.3k Ω in the schematic provides the proper 13.53 ratio using typical resistor values.

With proper turn-on and duty cycle settings secured, the next step in modifying the macromodel dealt with increasing the power limit. The LT1952-1's OC (over current) pin monitors current through the primary winding and MOSFET using a current sense resistor. The LT1952-1 compares the voltage at OC (hence referred to as " V_{OC} ") to an internal 107mV reference. When V_{OC} exceeds 107mV, the internal comparator commands the OUT pin to temporarily halt switching. In this way, the chip prevents damage to the external MOSFET and other components. In this case, however, the default 10m Ω resistor greatly limits the converter's power capacity. A quick Ohm's Law calculation reveals that 10m Ω sets a 10.7A-peak current limit, which doesn't provide nearly enough headroom to reach 300W at the converter's output.

The sense resistor needs to allow for maximum operating power under realistic conditions, while still providing protection above that level. For simulation purposes, however, the converter only needed a resistor that stayed out of the way. Thus, many of the simulations utilized a $1\text{m}\Omega$ sense resistor (referred to as “ R_{OC} ” from now on), which allowed plenty of headroom for high primary-side currents and high power outputs. When it came time to test the actual forward converter, parasitic inductance caused relatively large voltage spikes at the OC pin, resulting in unnecessary tripping. A low pass filter could have corrected this problem, but instead this project opted to eliminate the overcurrent network entirely.

Figure 4-22: LT1952-1 Forward Converter Following Early Calibrations [33]

This schematic depicts the original LT1952-1 macromodel following several key changes: R_{fb1} boosted to $205\text{k}\Omega$ to allow for proper feedback monitoring; transformer turns ratio changed from 1:1 ($10\mu\text{H}:10\mu\text{H}$) to 1:7 ($10\mu\text{H}:490\mu\text{H}$) to provide full step-up for 7V-65V input range; SD_Vsec branch adjusted to provide proper duty cycle for full input voltage range; Vin start-up/feedback network replaced with 12V_{DC} source (representing elliptical trainer’s battery).

The output capacitor and inductor also increased in size to reduce output voltage and current ripples.

Figure 4-23: Full Load Output Waveforms for LT1952-1 Forward Converter [33]

This plot shows the output voltage (green) and current (blue) waveforms for the LT1952-1 forward converter in Figure 4-22, where $V_{IN} = 65V$ and $I_{LOAD} = 5.5A$. $P_{OUT} = 275W$.

4.5.3 The Active Clamp

Before delving into the active clamp design process, one should fully understand the problem it aims to correct. When the forward converter's main switch turns off, the transformer's magnetizing and leakage inductances resist the sudden change in current. With no reset path present, the magnetizing current forces its way through high impedance routes, causing large voltage spikes. Figure 4-3 shows the type of voltage spikes that tend to occur across the main FET. At ~750V peak, those spikes can cause significant damage to primary side circuitry, especially the main FET. Power spikes associated with hard-switching can also wreak havoc on DC-DC converters, leading to reduced efficiency, added heat, and added stress on the switches. Figure 4-24 shows the voltage, current, and power waveforms associated with the hard-switching forward converter in Figure 4-22.

Figure 4-24: Hard Switching Waveforms [33]

Hard switching often causes large power losses at FET turn-on/off. In this situation, switch voltage (blue) and current (red) overlap at turn-off, creating 4.5kW power spikes (green).

The active clamp provides a reset path for the magnetizing energy and typically produces softer switching (ideally zero voltage switching), thus reducing voltage and power spikes and improving efficiency and durability. If the active clamp does not provide adequate soft-switching, the designer can add dissipative snubber circuits to further reduce switching stress and losses.

William Andreyak's *Active Clamp and Reset Technique Enhances Forward Converter Performance*, and Dhaval Dalal's *Design Considerations for Active Clamp and Reset Technique* both provide excellent NMOS active clamp forward converter design guides [21-22]. Those two papers provided much of the information in section 4.4.3. For PMOS active clamp design, refer to Juan Carlos Pastrana's *Design of a 100W Active Clamp Forward DC-DC Converter for Telecom Systems Using the NCP1562* [23]. Pastrana's paper also proved helpful in this active clamp design.

Before implementing the active clamp in the LT1952-1 schematic, it makes sense to create a simplified circuit model to establish proof of concept. Using a simplified model saves on simulation time and makes trouble-shooting much easier. Figure 4-25 shows the first simplified active clamp schematic, which includes components and switching signals that mimic those in Figure 4-22. The Q_{MAIN} gate signal derives its amplitude, frequency, and duty cycle from simulations of the LT1952-1 forward converter with a 40V input. The Q_{CLAMP} gate signal ($V_{\text{GS } Q_{\text{CLAMP}}}$) nearly represents the inverse of the Q_{MAIN} gate signal ($V_{\text{GS } Q_{\text{MAIN}}}$), with only a small dead-time appearing between the falling edge of $V_{\text{GS } Q_{\text{CLAMP}}}$ and the rising edge of $V_{\text{GS } Q_{\text{MAIN}}}$. The $V_{\text{GS } Q_{\text{CLAMP}}}$ signal meets the requirements detailed in the active clamp design papers: it goes high almost immediately after $V_{\text{GS } Q_{\text{MAIN}}}$ goes low; it remains high while the magnetizing energy flows into the reset capacitor and back out again; it goes low for a short period before $V_{\text{GS } Q_{\text{MAIN}}}$ turns the Q_{MAIN} back on, facilitating voltage rise across the primary winding and promoting zero voltage turn-on [21-23].

Figure 4-25: Simplified Active Clamp Forward Converter Model [33]

Creating a simplified version of LT1952-1 forward converter allowed simulations to run much faster, and made trouble-shooting much easier. This design provided a convenient test environment for the active clamp circuitry.

On page 9 of *Design Considerations for an Active Clamp and Reset Technique*, Dhaval Dalal states, “For analysis of the steady state operation of active clamp circuits, it is assumed that the clamp capacitor, C_{cl} , is large enough to approximate V_{cl} as a voltage source for the particular operating point” [22]. Other papers reiterate this design metric [23-24]. According to Bill Andreyak, zero voltage turn-on (of the main FET) will occur if “enough inductive energy is stored to overcome the opposing capacitive energy requirements of the circuit and power MOSFET(s)” [21]. Indeed, the reset capacitor and primary inductance had to reach a delicate balance before ZVS occurred. On top of that, the reset capacitor had to sufficiently limit voltage stress across Q_{MAIN} , and the primary inductance had to provide an impedance path low enough to allow full power transfer at 200kHz. The ZVS realization came after countless simulations with various reset capacitor and primary inductor sizes, along with tweaks to the MOSFETs’ parasitic

elements. At 40V and ~250W in, with $C_{\text{RESET}} = 90\text{nF}$ and $L_P = 20\mu\text{H}$, zero voltage turn-on finally appeared.

Figure 4-26: First Active Clamp Forward Model that Utilized ZVS [33]

After tweaking the reset capacitance, primary inductance, and parasitic elements, zero-voltage turn-on occurred at Q_{MAIN} , resulting in lowered voltage stress, improved efficiency, smoother waveforms, and drastically reduced switching losses.

Figure 4-27: Simplified Active Clamp Forward Converter Key Waveforms (250W Load) [33]

This plot shows some of the most important waveforms in the zero voltage switching realization. Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = I_{IN} ; Pink = I_{Lp} ; Green = V_{Lp} . Compare the V_{Lp} signal to the voltage waveform in Figure 4-3, and note the reset capacitor's effect during off-time: instead of 700V+ spikes, the active clamp limits voltage stress to 24V. Also note the gentle rise of V_{Lp} during the dead-time between $V_{GS\ QCLAMP}$ going low and $V_{GS\ MAIN}$ going high. This phenomenon depicts the zero voltage transition, and sees more attention in Figure 4-28.

Figure 4-28: Zero Voltage Turn-On at Q_{MAIN} in Simplified Active Clamp Forward Converter (250W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = V_{CRESET} ; Green = V_{Lp} . When Q_{CLAMP} turns off, it blocks the current path from C_{RESET} to L_p . The primary inductance finds a new current path through C_{IN} and the parasitic capacitance in Q_{MAIN} , charging C_{IN} and discharging C_{MAIN} . Eventually, V_{Lp} surpasses V_{IN} , but D_{MAIN} clamps the maximum voltage at $V_{IN} + V_{D(ON)}$. With only $-V_{D(ON)}$ across the drain-source of Q_{MAIN} , zero voltage turn-on commences. Note that the voltage across C_{RESET} (blue) never fully discharges, and acts like a ~14V DC source.

Figure 4-29: Simplified Active Clamp Forward Converter MOSFET Power Dissipation (250W Load) [33]

With zero voltage switching in place, the two MOSFETs experience very low power dissipation at switching intervals. Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = P_{QCLAMP} ; Green = P_{QMAIN} . While the hard switching signals in Figure 4-24 created 4.5kW spikes at Q_{MAIN} turn-off, the active clamp converter only sees brief 240W spikes at turn-off and 60W spikes at turn-on.

Establishing the active clamp circuitry in the LT1952-1 forward converter posed two major obstacles. First, the LT1952-1 only generates two synchronous switching signals: OUT and SOUT [17]. The active clamp requires two nearly opposite signals to drive Q_{MAIN} and Q_{CLAMP} , and those signals require a period of dead-time between the falling edge of $V_{GS\ QCLAMP}$ and the rising edge of $V_{GS\ QMAIN}$. Second, realizing the active clamp requires a high-side driver to control Q_{CLAMP} . This necessitates an additional IC; one that can sync with the LT1952-1 switching signal(s). As with the previous active clamp tests, each of these tasks saw implementation in the simplified active clamp circuit first.

The LT1952-1 only provides two synchronous switching signals, but they're not identical. Designers can program SOUT to rise before OUT, with the delay between the two rising edges controlled by the external resistor R_{DELAY} [17]. Inverting SOUT (with a CMOS inverter) creates a perfect gate drive signal for Q_{CLAMP} , where R_{DELAY} actually controls the dead

time between the falling edge of $V_{GS\ QCLAMP}$ and the rising edge of $V_{GS\ QMAIN}$. Figures 4-30 and 4-31 illustrate the differences between SOUT and OUT, and the effect of R_{DELAY} .

Figure 4-30: OUT and SOUT Switching Signals [17]

The LT1952-1 generates two synchronous switching signals: OUT and SOUT. An external resistor (R_{DELAY}) provides the designer a programmable delay between the rising edge of SOUT and rising edge of OUT. While SOUT typically sees use in driving synchronous rectifier MOSFETs, this design uses SOUT to control an active clamp MOSFET [17].

Figure 4-31: t_{DELAY} vs. R_{DELAY} [17]

This plot depicts the linear relationship between t_{DELAY} (the time between the rising edge of SOUT and the rising edge of OUT) and the external resistor R_{DELAY} . In this design, setting R_{DELAY} to 237kΩ provides 237ns of dead time (after inverting the SOUT signal), during which the primary winding voltage builds up to V_{IN} and preps Q_{MAIN} for zero voltage turn-on.

An inverted SOUT provides a nice switching signal for Q_{CLAMP} , but it requires a high-side driver in order to communicate with the high-side MOSFET. Luckily, Linear Technology produces a synchronous high-side driver, the LTC4444 [37]. Using this IC in LT Spice creates a convenient and accurate simulation. The LTC4444 also comes with a simple, nearly universal macromodel schematic, shown in Figure 4-32. After replacing D_{BOOST} with a higher breakdown voltage diode, the macromodel interfaced directly with the simplified active clamp forward schematic.

Figure 4-32: LTC4444 Macromodel Test Fixture [33, 37]

This design opted for the LTC4444 high-side driver due to its high voltage range (max bootstrap supply voltage = 114V) and availability in LT Spice.

Figure 4-33: Simplified Active Clamp Forward Converter with LTC4444 [33]

After replacing the D_{BOOST} diode, the LTC4444 drove the high-side Q_{CLAMP} MOSFET with ease.

Figure 4-34: Simplified Active Clamp Forward Converter with LTC4444 and CMOS Inverter [33]

Running the SOUT signal through a high-speed CMOS inverter and then the high-side driver created a perfect switching signal for Q_{CLAMP}.

After these additional components worked in the simplified active clamp forward converter, they joined the LT1952-1 forward converter to create the complete circuit.

Figure 4-35: First Revision LT1952-1 Active Clamp Forward Converter [33]

The first revision LT1952-1 Active Clamp Forward Converter employs a 7:1 step-up transformer and 90nF reset capacitor.

Figure 4-36: First Revision LT1952-1 Active Clamp Forward Converter Output Waveforms (205W Load) [33]

This figure shows the output voltage and current waveforms for the first revision active clamp forward converter under a 205W load. Green = V_{OUT} ; Blue = I_{LOAD} . The output voltage waveform displays very low ripple, even with a 4A load connected to the output. Programming the load current to rise slowly and linearly prevented the overcurrent network from tripping unnecessarily during start-up.

Figure 4-37: First Revision LT1952-1 Active Clamp Forward Converter Key Waveforms (205W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = I_{Lp} ; Green = V_{Lp} . The current through the primary winding (I_{Lp} , in blue) displays the typical sawtooth-like waveform, but comes with a good deal of ringing and an inexplicable spike at turn-on. The voltage across the primary winding (V_{Lp} , in green) does not rise during the dead-time between the falling edge of $V_{GS\ QCLAMP}$ and rising edge of $V_{GS\ QMAIN}$, which likely causes the larger power spikes seen in Figure 4-40. However, other input conditions resulted in more familiar ZVS waveforms (see Figure 4-39).

Figure 4-38: First Revision LT1952-1 Active Clamp Forward Converter MOSFET Power Dissipation (205W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = P_{QCLAMP} ; Green = P_{QMAIN} . The turn-on and turn-off power spikes in the LT1952-1 forward converter exceed those seen in the simplified active clamp converter, depicted in Figures 4-26 through 4-29. However, these power spikes show dramatic improvement over the 4.5kW spikes resulting from the hard switching waveforms in Figure 4-24. Also, other load conditions presented much smaller switching power spikes, as well as zero-voltage switching (see Figures 4-39 and 4-40).

Figure 4-39: First Revision LT1952-1 Active Clamp Forward Converter Key Waveforms (25W Load) [33]

Red = $V_{GS\ Q_{MAIN}}$; Teal = $V_{GS\ Q_{CLAMP}}$; Blue = I_{Lp} ; Green = $V_{DS\ Q_{MAIN}}$. With a 16V, 25.6W input to the first revision LT1952-1 active clamp forward converter, zero voltage switching appeared once again. Note how the voltage across Q_{MAIN} 's drain-source (green) descends to zero before the gate signal (red) turns the FET back on. This led to substantially smaller switching losses, shown in Figure 4-40.

Figure 4-40: First Revision LT1952-1 Active Clamp Forward Converter MOSFET Power Dissipation (25W Load) [33]

Red = $V_{GS\ Q_{MAIN}}$; Teal = $V_{GS\ Q_{CLAMP}}$; Blue = $P_{Q_{CLAMP}}$; Green = $P_{Q_{MAIN}}$. With a 16V, 25.6W input to the active clamp forward converter, zero voltage switching appeared once again. The waveforms shown in Figure 4-39 coincide with very small switching power spikes: only ~50W peak across Q_{MAIN} and ~5W peak across Q_{CLAMP} . Thus, ZVS still occurs in the first revision LT1952-1 active clamp forward converter, but not under all load conditions.

Figure 4-41a: First Revision LT1952-1 Active Clamp Forward Converter Efficiency Plot

LT Spice offers a simulation option that calculates steady state efficiency [33]. This plot shows the first revision LT1952-1 active clamp forward converter's efficiency over a range input voltages. Each data point maintains a 10Ω impedance, such that $V_{IN}/I_{IN} \approx 10\Omega$ (the ideal load impedance for the elliptical trainer). Maintaining the 10Ω impedance dictates a proportional relationship to input power. For better clarity, Figure 4-42b shows the same plot, but with input power replacing input voltage.

Figure 4-41b: First Revision LT1952-1 Active Clamp Forward Converter Efficiency Plot

LT Spice offers a simulation option that calculates steady state efficiency [33]. This plot shows the first revision LT1952-1 active clamp forward converter's efficiency over a range power inputs. Each data point maintains a 10Ω impedance, such that $V_{IN}/I_{IN} \approx 10\Omega$ (the ideal load impedance for the elliptical trainer). Maintaining the 10Ω impedance dictates a proportional relationship to input voltage. For better clarity, Figure 4-42a shows the same plot, but with input voltage replacing input power.

Figure 4-42: First Revision LT1952-1 Active Clamp Forward Converter Voltage Regulation

This plot shows the first revision LT1952-1 active clamp forward converter's voltage output over a range input voltages. These data points come from LT Spice simulations, where input voltage and output current varied in order to maintain $V_{IN}/I_{IN} \approx 10\Omega$ (the ideal impedance for the elliptical trainer [33]). In the simulations, the converter failed to turn on below 12.5V in. For $12.5V < V_{IN} < 52V$, the active clamp forward converter showed excellent load regulation, never straying far from the 51.2V target [28].

CHAPTER 5: FORWARD CONVERTER REALIZATION

5.1 Introduction

Chapter 5 follows the construction of the active clamp forward converter. Section 5.2 covers component selection, with extra focus on the switch mode transformer. Section 5.3 goes through the printed circuit board design. Section 5.4 follows physical testing, and includes results and trouble-shooting maneuvers. Finally, Section 5.5 reviews the active clamp forward converter's performance and its potential use in future EHFEM implementations.

5.2 Component Selection

Once the LT1952-1 active clamp forward converter performed well in LT Spice, the component selection process began. To ensure the highest quality operation, this design opted for a printed circuit board over a perf board, and selected mostly surface mount components. LT Spice simulations essentially determined component sizing, and in areas that posed any doubt, this project chose the more robust element. To promote brevity, this section only touches on the sizing of a few components, particularly the switch mode transformer.

Selecting MOSFETs for this converter focused on the relationship between gate capacitance and $R_{DS(ON)}$. Gate capacitance determines a MOSFET's ability to respond to switching signals; higher gate capacitance leads to a slower response, while lower gate capacitance leads to a quicker response. $R_{DS(ON)}$ dictates the conduction losses through the MOSFET's drain-source channel. These two metrics possess an inversely proportional relationship, and designers must find a harmonious balance between the two, especially in high frequency, high power applications. In choosing the MOSFET for Q_{MAIN} , $R_{DS(ON)}$ outweighed gate capacitance, predominantly due to the high primary-side currents. With Q_{CLAMP} , gate capacitance played a more pivotal role, since the MOSFET had to turn on almost immediately after Q_{MAIN} turned off. Finally, the two MOSFETs that form the CMOS inverter each boasted

very low gate capacitances, ensuring a clean switching signal for Q_{CLAMP} . Each of these MOSFETs demanded different trade-offs for optimal performance.

Like gate capacitance and $R_{DS(ON)}$, inductance and saturation current also share an inversely proportional relationship. This design requires an output inductor that can maintain at least 5A of continuous DC current. The chosen inductor must possess a saturation rating equal to 5A plus the peak current ripple. However, inductance plays a large role in determining current ripple, and while lowering inductance may increase saturation current rating, it also increases current ripple. Finding the proper inductor proved quite difficult, and like the MOSFETs, it took many simulations in LT Spice to find one that would fit this converter.

Resistors and capacitors did not receive much attention in the component selection process. To achieve enough capacitance at the input and output to the converter, the design placed low-ESR ceramic capacitors in parallel with a large, electrolytic capacitor. This provided both high capacitance and low ESR. Since all of the resistors dissipate very little power, they did not require much consideration when it came time to order.

For the complete list of components (aside from the transformer and Linear Technology chips), including stress ratings and cost, refer to Appendix C.

No component proved more difficult to find than the switch mode transformer. The next section follows the search for an adequate transformer, and the design trade-offs incurred with the model received.

5.2.1 The Switch Mode Transformer

This project's active clamp forward converter calls for a highly specialized switch mode transformer. While many websites sell custom-built switch mode transformers, none of the vendors contacted would handle this transformer request [39-46]. Some only provide custom

builds for 10,000+ unit orders, while others simply refused such a complex design. Ideally, this converter's switch mode transformer would meet the following specifications:

- Primary winding voltage range: $7V < V_{PRI} < 65V$
- Power capacity: $P_{OUT(MAX)} \approx 350W$
- Operating frequency: $f_s \approx 200kHz$
- Turns-ratio: 1:7 (step-up)
- Primary Inductance: $L_P \approx 20\mu H$
- Voltage waveform: square
- Package: through-hole or surface mount

With no company willing to build a handful of these custom models, the project had to settle on one of two options: (1) build a custom transformer that meets the specifications; (2) find a commercial switch mode transformer that fits as many of the specifications as possible. With little to no background in inductor and transformer design, this project searched out a commercial transformer that met most of the converter's needs. A long and difficult search ended with Pulse Engineering's PA0934 switch mode transformer, which offers the following specifications [47]:

- Primary winding voltage range: $18V < V_{PRI} < 76V$
- Power capacity: $P_{(MAX)} \approx 250W$
- Operating frequency: $200kHz < f_s < 700kHz$
- Turns-ratio: 1:3.5 (step-up)
- Primary Inductance: $L_P = 54\mu H$
- Voltage waveform: square
- Package: surface mount

Rita Jendro, a representative at Pulse Engineering, offered to send three PA0934 samples free of charge. While the turns-ratio, input voltage range, and primary inductance all strayed from the desired specifications, the transformer still met most of the active clamp forward

converter's needs, and preliminary runs in LT Spice showed that it functioned fairly well in the LT1952-1 active clamp forward converter.

Figure 5-1: Pulse PA0934 Switch Mode Transformer [47]

While the Pulse PA0934 switch mode transformer doesn't meet all the needs of the active clamp forward converter, it boasts high power capacity, high frequency, and reasonable turns-ratio and primary inductance.

Several components required adjusting to meet the needs of the new transformer. The 3.5:1 turns-ratio dictated a new SD_V_{sec} branch (formed by R_{SD1} , R_{SD2} , and D_{SD}). The PA0934's lowered turns-ratio hinders the converter's step-up ability, preventing inputs below 12V from generating output voltage in the Gridpoint Connect C36's input range: $40V < V_{INVERTER} < 60V$ [47]. Also, the input voltage range for the PA0934 transformer bottoms out at 18V, meaning inputs below that range may not work either. Ordering various surface mount resistors provided options with setting the new SD_V_{sec} voltage divider. With the new primary inductance more than double the old value, the reset capacitor also demanded tweaking to ensure proper active clamp operation, and ideally, zero voltage switching. Ordering extra capacitors and creating large C_{RESET} pads in the PCB design provided room to adjust the reset capacitance value physically.

The second revision LT1952-1 active clamp forward converter includes a mock PA0934 transformer, a reworked SD_V_{sec} branch, and a new reset capacitor. The schematic shown in Figure 5-2 contains all of the components used in the built converter.

Figure 5-2: Second Revision LT1952-1 Active Clamp Forward Converter [33]

The second revision active clamp forward converter differs from the first revision in that it uses the Pulse PA0934 3.5:1 step-up transformer instead of the ideal 7:1 step-up transformer, and a 100nF reset capacitor instead of a 90nF one. Some other components (like R_{SD1} and R_{SD2}) required tweaking to work with the new transformer. The lower turns-ratio prevents input voltages below 12V from reaching the 40V output threshold, meaning that milder exercise routines would not generate grid-tied electricity.

Figure 5-3: Second Revision LT1952-1 Active Clamp Forward Converter Key Waveforms (160W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = I_{LP} ; Green = $V_{DS\ QMAIN}$. As with Figure 4-37, the current through the primary winding (I_{LP} , in blue) displays the typical sawtooth-like waveform, but comes with a good deal of ringing and a large spike at turn-on. The voltage across Q_{MAIN} 's drain-source ($V_{DS\ QMAIN}$, in green) does not descend to zero during the dead-time between the falling edge of $V_{GS\ QCLAMP}$ and rising edge of $V_{GS\ QMAIN}$, which likely causes the larger power spikes seen in Figure 5-4.

Figure 5-4: Second Revision LT1952-1 Active Clamp Forward Converter MOSFET Power Dissipation (160W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = P_{QCLAMP} ; Green = P_{QMAIN} . While Q_{MAIN} sees very little power dissipation at turn-on, it experiences 1kW spikes at turn-off. The waveforms in Figure 5-3 indicate that the new transformer does not maintain zero voltage switching with the 100nF reset capacitor. This problem grew worse in lower power simulations (see Figure 5-6), and ultimately earned extra attention after the first two converters failed during physical testing.

Figure 5-5: Second Revision LT1952-1 Active Clamp Forward Converter Key Waveforms (25W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = I_{Lp} ; Green = $V_{DS\ QMAIN}$. As with the 160W load waveforms in Figure 5-3, $V_{DS\ QMAIN}$ (in green) does not descend to 0V during off-time and I_{Lp} (in blue) shows excessive ringing. These odd waveforms led to tremendous switching power spikes, shown in Figure 5-6 below.

Figure 5-6: Second Revision LT1952-1 Active Clamp Forward Converter MOSFET Power Dissipation [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = P_{QCLAMP} ; Green = P_{QMAIN} . The odd switching waveforms shown in Figure 5-5 coincide with very high power spikes across the both MOSFETs when Q_{MAIN} shuts off. This problem saw more attention following the first round of physical testing.

Figures 5-3 through 5-6 indicate an imbalance between C_{RESET} and L_P (the primary magnetizing inductance) in the second revision 1952-1 active clamp forward converter. Many active clamp design papers note problems like this when the capacitive energy exceeds the magnetizing inductive energy [21-22, 48]. Reducing the reset capacitor's size allows for quicker discharge and aids ZVS, although it also yields higher peak voltage across the main MOSFET. This report revisits the magnetizing energy balance problem in section 5.4.3.

Increased switching losses aside, the PA0934 transformer showed that it can deliver the proper output voltage and power for most of the specified input conditions. A custom-built transformer would likely work better with the converter, but the PA0934 allows for adequate testing of the active clamp forward converter topology in this EHFEM scenario. The second revision active clamp forward converter achieved very similar system efficiencies to the first revision converter, but showed reduced load regulation due to the lowered turns-ratio. Figure 5-

7 depicts the relationship between input voltage and output voltage for the second revision active clamp forward converter.

Figure 5-7: Second Revision LT1952-1 Active Clamp Forward Converter Voltage Regulation

This plot shows the second revision LT1952-1 active clamp forward converter’s voltage output over a range input voltages. These data points come from LT Spice simulations, where input voltage and output current varied in order to maintain $V_{IN}/I_{IN} \approx 10\Omega$ (the ideal impedance for the elliptical trainer [33]). In the simulations, the converter failed to turn on below 13V in. Due to PA0934’s lowered turns-ratio, the converter failed to output greater than 40V (the minimum input to the Gridpoint Connect C36 inverter) for $V_{IN} < 15V$.

5.3 PCB Design

Poor PCB layouts can ruin well-designed circuits. To ensure that didn’t happen here, the layout relied heavily on PCB layout guides [49-52]. These papers promoted proper trace length and width, component spacing, suitable ground planes, and other techniques. The layout took place in Express PCB’s patented software programs: ExpressSCH and ExpressPCB [52].

Figures 5-8, 5-9, and 5-10 show the end product: the first revision printed circuit board design.

Figure 5-8: PCB Layout – Top and Bottom Layers [52]

This figure shows all three layers of the PCB design: the top traces in red; the bottom traces in green; the painted text in blue. All of the components retain the same names they had in previous LT Spice schematics.

Figure 5-9: PCB Layout – Top Layer [52]

This figure shows only the top layer of the PCB design, which includes the main traces (in red) and painted text (in blue). The wider traces carry high currents, while the thinner traces carry switching and reference signals.

Figure 5-10: PCB Layout – Bottom Layer [52]

This figure shows only the bottom side of the board, which includes all of the bottom traces (in green). Solid ground planes occupy most of the space on the underside of the board: the left plane represents the secondary-side power ground; the right plane represents the primary-side power ground; the top plane represents the analog ground. The two squares in the middle of the primary-side power ground connect to the drain of each MOSFET and serve as heat sinks; thermal vias establish a connection between the top and bottom layers.

5.4 Physical Testing

5.4.1 Construction

Once all of the components arrived, soldering commenced. Most of the assembly only required a soldering iron, but a few components – the LTC4444, R_{OC} , Q_{MAIN} , and Q_{CLAMP} – required a hot air gun to completely attach their hidden pads to the board. Each converter took several hours to assemble, but the builds encountered few obstacles.

5.4.2 First Round Testing

Figure 5-11: First Active Clamp Forward Converter

This figure shows the first completed active clamp forward converter. The scattered, empty through holes represent nodes for test leads.

The first circuit experienced severe loading problems early on. While the converter delivered roughly 51V to the output under no load conditions, it could not operate continuously under loads greater than 25W. Eventually, the solution emerged: leakage inductance caused the LT1952-1's OC pin to see high voltage spikes at switching transitions, causing the overcurrent circuitry to trip unnecessarily. When the overcurrent circuitry trips, the LT1952-1 temporarily halts switching, which causes the output voltage to sag and ultimately leads to complete output voltage deregulation [17]. Figure 5-12 shows the voltage spikes at the converter's OC pin.

Figure 5-12: Leakage Inductance Trips Overcurrent Protection

In this oscilloscope capture, the top waveform represents $V_{GS_{QMAIN}}$, the gate signal to the main MOSFET; the bottom waveform represents V_{OC} , the reference signal to the LT1952-1's overcurrent (OC) pin. The zoom on the V_{OC} waveform measures 500mV per division, so the OC pin sees close to 1.5V at Q_{MAIN} turn-off. It only takes 107mV to trip the overcurrent sensor, and these 1.5V magnitude spikes greatly surpass that threshold.

A low pass filter connected between the OC pin and the R_{OC} node might successfully block the high frequency spikes and prevent the overcurrent tripping. However, the PCB lacked space for that sort of correction. Instead, this situation called for a simpler fix. Severing the connection between the OC pin and sense resistor (R_{OC}) and routing the OC pin to analog ground disabled the overcurrent protection entirely. This allowed for higher primary side currents and higher power conversion.

Figure 5-13: OC Pin Routed to Analog Ground

Severing the connection between OC and the R_{OC} (the white component labeled “R 002”) and routing OC to analog ground (using the jumper connected to the near side of “Ris”) disabled the overcurrent protection network. This allowed the converter to reach much higher power levels.

After eliminating the overcurrent sensor, the converter performed much better. It quickly reached 100W at the output, and boasted nearly 90% efficiency. Before moving to higher power levels though, the converter required extra heat-sinking. The printed circuit board design includes pads and thermal vias that link the drains of Q_{MAIN} and Q_{CLAMP} to the bottom side of the board. This architecture allows for heat sink attachment to the drain of each FET, which lowers the chances of overheating at high loads. Thermal paste formed a direct, non-electrical connection between the MOSFETs and the heat sinks. The switching diodes on the secondary side of the converter received a vertical heat sink of their own (connected across each of their cathodes). These heat sinks help dissipate heat incurred during high load conditions.

Figure 5-14: Heat-Sinking Q_{MAIN} and Q_{CLAMP}

Ceramic thermal paste formed a bond between the power MOSFETs and heat sinks, allowing for safer power dissipation at high loads. Insulating tape covered nearby through-holes and prevented shorting.

With the heat sinks secure, high power testing began. A Hewlett Packard 6574A DC power supply drove the input to the active clamp forward converter, playing the role of the elliptical trainer, while a GW GPR-6060D power supply represented the Precor EFX's 12V battery. The converter's output fed a BK Precision 5410 electronic load, which dictated output (and input) current. Finally, the Tektronix TDS 1012 oscilloscope provided views of the high frequency waveforms. Figure 5-15 depicts the test setup.

Each data point strove to maintain a 10Ω input to the active clamp forward converter, to mimic the ideal load for the elliptical trainer. For example, when the input voltage rose to 24V, the electronic load current also increased until input current equaled 2.4A ($V_{IN}/I_{IN} = 10\Omega$).

Figure 5-15: First Round Test Setup

The first tests utilized the equipment shown in Figure 5-15: a 600W HP 6574A DC power supply (bottom left), a Tektronix TDS 1012 oscilloscope (top center), a BK Precision 5410 electronic load (bottom center), and a GW GPR-6060D DC power supply (top right).

Table 5-1: First Round Test Results

This table lists the first round test results for the active clamp forward converter. The reworked SD_Vsec branch set the undervoltage lockout to just below 13V in, and the PA0934 transformer likely impeded efficiency at input voltages below 18V [47]. After a few seconds at 40V and 158W in, the active clamp portion of the converter failed.

Vin [V]	Iin [A]	Zin [Ω]	Pin [W]	Vcc [V]	Icc [A]	Pcc [W]
14	1.45	9.66	20.3	12	0.08	0.96
16	1.67	9.58	26.7	12	0.08	0.96
20	1.94	10.31	38.8	12	0.09	1.08
24	2.46	9.76	59.0	12	0.10	1.20
28	2.81	9.96	78.7	12	0.10	1.20
32	3.19	10.03	102.1	12	0.09	1.08
36	3.61	9.97	130.0	12	0.09	1.08
40	3.96	10.10	158.4	12	0.09	1.08
Vin [V]	Vo [V]	Io [A]	Po [W]	η [%]	Vopp [V]	Vopp [%]
14	41.73	0.10	4.17	19.63%	-	-
16	47.39	0.10	4.74	17.12%	-	-
20	51.66	0.55	28.41	71.25%	-	-
24	51.55	1.00	51.55	85.57%	-	-
28	51.44	1.35	69.44	86.94%	-	-
32	51.32	1.76	90.32	87.56%	-	-
36	51.2	2.22	113.66	86.74%	-	-
40	51.11	2.70	138.00	86.53%	-	-

Figure 5-16: First Round Test Results: Efficiency vs. Input Voltage

Since input voltage shares a proportional relationship with input power, the efficiency vs. input power plot would like identical to this one, with input power range extending from 20W (at 14V) to 158W (at 40V).

Figure 5-17a: First Round Test Results: Output Voltage vs. Input Voltage

This plot shows the converter's efficiency over a range of input voltages from the DC power supply. Each data point maintains a $\sim 10\Omega$ impedance, such that $V_{IN}/I_{IN} \approx 10\Omega$. Maintaining the 10Ω impedance dictates a proportional relationship between input voltage and power. Figure 5-17b replaces input voltage with input power.

Figure 5-17b: First Round Test Results: Output Voltage vs. Input Power

This plot shows the converter's efficiency over a range of input powers from the DC power supply. Each data point maintains a $\sim 10\Omega$ impedance, such that $V_{IN}/I_{IN} \approx 10\Omega$. Maintaining the 10Ω impedance dictates a proportional relationship between input power and voltage. Figure 5-17a shows the same plot, but with input voltage replacing input power.

The first two active clamp forward converters failed after several seconds running at 40V and 158W in. Something – possibly the reset capacitor – caused the LTC4444, Q_{MAIN} , and D_{MAIN} to all fail. The second round of testing looked to correct this failure.

5.4.3 Second Round Testing

The second round of testing examined the possible causes of failure in the active clamp portion of the converter. Previous tests showed that the converter operated normally with 40V in and small input currents, so the initial hypothesis figured that high power, rather than high voltage, caused Q_{MAIN} , D_{MAIN} , and the LTC4444 driver to fail. LT Spice simulations depicted high power spikes at Q_{MAIN} turn-off (see Figures 5-3 through 5-6), and Q_{MAIN} became quite hot under the 40V, 158W input during the previous tests. Drastically reducing those power spikes, and the overall power dissipated in Q_{MAIN} , seemed like a reasonable solution.

The active clamp forward converter relies on smooth resonance between the reset capacitor and magnetizing inductance to promote zero voltage switching. When the resonant frequency becomes too large, the magnetizing energy transfer between L_P and C_{RESET} diminishes, reducing the chances for soft switching. Equation 5-1 shows the relationship between the primary inductance (L_P), reset capacitance (C_{RESET}) and resonant frequency.

Equation 5-1: Active Clamp Resonant Frequency Calculation

$$f_{resonant} = \frac{1}{2\pi\sqrt{L_P C_{RESET}}}$$

Reducing C_{RESET} 's capacitance increases the switching frequency and should allow for smoother, less dissipative switching transitions for Q_{MAIN} and Q_{CLAMP} . Subsequent LT Spice simulations reinforced this hypothesis.

Figure 5-18: Key Waveforms Using Smaller C_{RESET} (25W Load) [33]

Red = $V_{GS\ Q_{MAIN}}$; Teal = $V_{GS\ Q_{CLAMP}}$; Blue = I_{L_P} ; Green = $V_{DS\ Q_{MAIN}}$. These waveforms depict the effect a smaller reset capacitance (in this case, $C_{RESET} = 2\text{nF}$) can have on the active clamp forward converter. Compare these waveforms to those shown in Figure 5-5. The voltage across Q_{MAIN} 's drain-source decays much closer to 0V prior to Q_{MAIN} turning on, but the waveform also boasts a much higher peak voltage. This sort of reaction promotes zero-voltage switching, but also increases voltage stress across the main MOSFET.

Figure 5-19: MOSFET Power Dissipation Using Smaller C_{RESET} (25W Load) [33]

Red = $V_{GS\ QMAIN}$; Teal = $V_{GS\ QCLAMP}$; Blue = P_{QCLAMP} ; Green = P_{QMAIN} . While the turn-off power spikes still reach high levels, they show dramatic reduction over those depicted in Figure 5-6. The smaller reset capacitor ($C_{RESET} = 2nF$) promotes softer switching than the 100nF reset capacitor used in earlier simulations.

The third (and final) active clamp forward converter underwent the same construction process as the previous two, with one exception: C_{RESET} decreased in size from 100nF to 50nF (by placing two 100nF capacitors in series). The test setup established a no-load condition at the converter's output, 24V at the converter's input, and 12V as the rail supply for the two chips and CMOS inverter. The test setup also employed more robust heat sinking to help with heat management. Unfortunately, Q_{MAIN} failed immediately after turning the input supply on. With no more MOSFETs available, testing halted.

This test seemed to indicate that voltage stress destroyed the final Q_{MAIN} MOSFET ($V_{DS(MAX)} = 100V$ [38]). Halving the size of C_{RESET} may have promoted more efficient switching, but it also causes higher peak voltages across Q_{MAIN} 's drain-source. To investigate the high voltage problem further, this project found replacement MOSFETs with higher drain-source voltage tolerances and reset capacitors with higher voltage ratings [53] [54].

5.4.4 Third Round Testing

The third and final round of testing hoped to create a more robust converter and determine the previous causes of failure. Two converters received replacement MOSFETs at Q_{MAIN} . One converter kept the 50V-rated 100nF reset capacitor used in previous tests; the other saw its reset capacitor replaced by two 100V-rated 47nF capacitors connected in parallel (creating 94nF of total capacitance) [54]. Each converter utilized the same test setup as before.

The new MOSFET at Q_{MAIN} seemed to function well, although each forward converter performed very strangely. One converter showed extremely high step-up under no-load conditions, transforming a $20V_{\text{DC}}$ input into a $>80V_{\text{DC}}$ output (the BK Precision 5410 electronic load stopped measuring output voltage around 80V). However, introducing a small load current of 100mA brought the output voltage back down to roughly 47V. Touching the transformer with one's finger provided a discharging effect, causing the output voltage to momentarily droop. The other converter exhibited roughly 51V at its output under no-load condition, but the output voltage instantly fell to zero following the introduction of any load current. These strange behaviors prompted the idea that the PA0934 transformers may have failed.

Figure 5-20: Drain Voltage at Q_{MAIN}

One forward converter (the one with the new reset capacitors) showed very little voltage clamping at Q_{MAIN} 's drain node. With a 20V and 200mA input, the drain voltage peaked at 80V, before resonating back down to 20V. This response proved very puzzling; waveforms like these shouldn't occur with low input current and a 54 μ H magnetizing inductance.

The GW Instek LCR Meter allows users to measure inductive and capacitive components. Following the strange behavior from the two active clamp forward converters, each transformer deserved testing to see if saturation may have occurred. The PA0934 datasheet lists the transformer's primary inductance at 54 μ H, but following tests showed wildly different values in each transformer [47].

Figure 5-21a: PA0934 Primary Inductance Measurement

Once soldered into place, the PA0934 transformers proved extremely difficult to remove. Instead of attempting removal, the first converter got scrapped, and all copper traces connected to the transformer nodes got severed. This isolated the transformer, allowing for accurate inductance measurements.

Figure 5-21b: PA0934 Primary Inductance Measurement

Instead of reading $54\mu\text{H}$ (the listed value from the datasheet), the PA0934's primary winding read only $0.89\mu\text{H}$. The secondary winding showed a similar discrepancy: $9.54\mu\text{H}$ instead of $661.5\mu\text{H}$ [47].

Table 5-2: Damaged Switch Mode Transformer Measurements

This table lists the measured primary inductances of each PA0934 transformer, and compares the measured values to the value listed in the datasheet. Transformer 1 had no connections to other components during testing (see Figure 5-21a), while transformers 2 and 3 saw measurement while fully connected in the forward converter circuit.

	Measured L_{PRI} [μ H]	Ideal L_{PRI} [μ H]	Difference [%]
Transformer 1	0.89	54.0	98.4%
Transformer 2	154.6	54.0	186.3%
Transformer 3	90.96	54.0	68.4%

Following these measurements, it appears reasonable to assume that each PA0934 transformer failed during active clamp forward converter testing. Saturation caused by heavy load currents or inadequate reset may have led to each transformer's failure. Transformer saturation may also explain the strange behavior from the two converters that received new MOSFETs.

HIGH FREQUENCY PLANAR TRANSFORMERS

PA09XXNL Series (up to 250W)

- Power Rating: up to 250W
- Height: 9.1mm to 10.4mm Max
- Footprint: 29.5mm x 26.7mm Max
- Frequency Range: 200kHz to 700kHz
- Isolation (Primary to Secondary): 1750V_{DC}

Figure 5-22: PA0934 Power Handling Discrepancy [47]

The PA0934 datasheet lists power rating at two instances near the top of the datasheet (shown). It makes no mention of whether this value matches output power or total apparent power.

The Pulse PA0934 datasheet only mentions power rating at a few instances, and doesn't state whether the rating describes deliverable output power or total apparent power [47]. If the 250W amount actually describes total apparent power, that would establish the output power rating at less than 104W, using the equation: $P_{TOTAL} = P_{OUT}(\frac{1}{\eta} + \sqrt{2})$ [55]. The datasheet also notes, "The package is typically capable of handling between 150W-250W of power depending

on the application, ambient conditions and available cooling” [47]. These clues indicate that previous tests may have pushed the transformer above its rated power levels, resulting in permanent damage.

5.5 Active Clamp Forward Converter in Review

The active clamp forward converter provided high-efficiency conversion over most of the specified input levels, although it performed very poorly under low load conditions and failed under the higher end power levels. Each of these problems likely ties back to the switch mode transformer. The 3.5:1 turns-ratio and 18V-76V input range essentially limited minimum input voltage to 18V. On the upper end, the PA0934’s increased primary inductance led to more magnetizing energy, which resulted in higher voltage and power spikes across the primary MOSFET (Q_{MAIN}). The vague power rating may have played a huge limiting role, if the “up to 250W” listing describes total apparent power rather than output power [47] [55]. If this project could attain a custom-built transformer that meets the specifications laid out in section 5.2.1, the converter would likely perform much better.

CHAPTER 6: CONCLUSION

6.1 Overview

This chapter provides the conclusion for Energy Harvesting From Exercise Machines: Forward Converters with a Central Inverter. Section 6.2 discusses future ideas for the EHFEM elliptical trainers project, including a custom-built switch mode transformer, additional elliptical trainer modeling, and a new series DC-DC converter approach. Section 6.3 finalizes the report with a brief conclusion.

6.2 Future Works

This section describes potential future tasks for EHFEM students, some of which may constitute entire senior projects and/or theses.

6.2.1 Switch Mode Transformer

This thesis's active clamp forward converter settled for a switch mode transformer with less than ideal specifications. The transformer's voltage range and turns-ratio prevented efficient energy conversion for voltage inputs under 18V, while the increased magnetizing inductance led to higher voltage stress, which likely caused the converter's primary MOSFET to repeatedly fail.

A custom-built transformer may allow for efficient conversion at all input levels. The ideal transformer could arise as a future senior project or thesis for a Cal Poly electrical engineering student. Opting for a custom-built transformer also makes it possible for the active clamp forward converter to feed inverters with higher input voltage ranges than the Gridpoint Connect C36, if that option becomes favorable. However, the custom built transformer should see careful modeling and simulation prior to implementation. This project demonstrated the delicate balance between the switch mode transformer and active clamp circuitry, and future transformer designs must communicate well with an active clamp.

6.2.2 Better EFX 546i Modeling

EHFEM teams have not yet explored the inner workings of the Precor EFX 546i elliptical trainer. Thus far, DC-DC converter designs have focused mainly on input voltage and power ranges, with only mild interest in maintaining constant 10Ω impedance to the elliptical trainer. This oversimplified view needs reworking. Future teams should explore the EFX 546i's "Lower PCA" module, and provide more information about eddy current braking, the creation of intensity levels, and the rectification process [27].

6.2.3 Constant Impedance Modeling / A New Converter Approach

During his thesis defense, Alvin Hilario suggested a system that incorporates two DC-DC converters in series. A high-efficiency, non-isolated converter (like the one Alvin designed) would feed a constant voltage to an isolated topology, which would step the voltage up to typical inverter input ranges (200V-600V) [3,12]. On top of the higher voltage output, this series DC-DC converter approach would reduce the switch mode transformer's input voltage range requirements.

As an added wrinkle, this setup could employ a current regulator between the two DC-DC converters, as a method for maintaining 10Ω impedance to the elliptical trainer. The current regulator would use pulse width modulation to control a high-side MOSFET, such that the system draws more current under higher voltages and less current under lower ones. Figure 6-1 depicts a block diagram of this potential setup, and Figure 6-2 provides a closer look at the pulse width modulated current regulator.

Figure 6-1: Series DC-DC Converter System with Current Regulator

This figure depicts a setup that allows large voltage step-up and maintains constant load impedance to each elliptical trainer. The non-isolated DC-DC converter conditions the input voltage for interaction with the step-up, isolated unit. In between, the PWM (pulse width modulation) current regulator controls the load current to promote constant impedance to the elliptical trainer.

Figure 6-2: PWM Current Regulator

This figure presents a closer look at the current regulator, which aims to control impedance seen by the elliptical trainer. A resistive network scales down the input voltage and feeds it into a pulse width modulator, which controls a high-side MOSFET that links the two DC-DC converters. Under higher input voltages, the MOSFET would allow more current to pass; under lower voltages, less current would pass.

6.3 Conclusion

This thesis defends the use of an active clamp forward converter as the critical link between Precor elliptical trainers and a grid-tied inverter. The active clamp forward converter offers wide duty cycle range and step-up capability, allowing for wide input voltage range and high output voltage. Making elliptical trainers interact with a central inverter necessitates each of these design traits. For optimal performance, the active clamp forward converter relies on a highly specialized switch mode transformer; one, however, that a Cal Poly electrical engineering student could build. The free switch mode transformers from Pulse Engineering held total DC-DC converter cost to just \$61.61 (see Appendix A, C for details), far below the ten-year, zero net cost laid out in Dr. Braun's proposal [26]. With the proper components, the active clamp forward converter could provide safe energy conversion for the Energy Harvesting From Exercise Machines elliptical trainers project.

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APPENDIX A: ABET Senior Project Analysis

Nick Lovgren

April 16, 2010

ABET Senior Project Analysis

Energy Harvesting From Exercise Machines: Forward Converters with a Central Inverter

Economic Considerations

This project aims to produce human-powered, grid-tied electricity using elliptical trainers at Cal Poly's Rec Center. Ten elliptical trainers will connect to ten DC-DC converters, which then feed a central, grid-tied inverter. Electrical engineering students, hopefully with the help of outside donations, will front the cost of the DC-DC converters and inverter. This system will reduce net energy consumption at the Rec Center by (1) converting human kinetic energy into marketable, grid-tied electricity, and (2) reducing heat emissions from the elliptical trainers, thus reducing the Rec Center's air conditioning needs. Reducing net energy consumption lowers the Rec Center's operating costs, a boon for the California State University and Associated Students Incorporated (ASI), the two bodies responsible for paying the Rec Center's utility bills.

Keeping the ten-machine system cost under \$3,600 heightens the chance of achieving net-zero cost after ten years (meaning that, after ten years, the system offsets its initial cost through energy generation/conservation) [26]. Table 1 breaks down the cost per elliptical trainer, but multiplying each amount by ten provides the complete system cost. The net-zero cost assessment assumes no change in maintenance fees following the retrofit. It also assumes that the DC-DC converters and inverter do not require replacement over that time.

Table 1: Itemized Cost Estimate for EHFEM: Elliptical Trainers and a Centralized Inverter

<u>Item</u>	<u>Estimated Cost per Trainer</u>	<u>Notes</u>
Gridpoint Connect C36 Inverter	\$250	Estimated. Total cost divided amongst 10 elliptical trainers.
Forward Converter Components	\$31.14	Includes everything except PCB and transformer
Printed Circuit Board	\$30.47	Cost per board from ExpressPCB
Switch Mode Transformer	\$0	Pulse Engineering provided 3 free samples
Total:	<u>\$311.61</u>	

If this project does not offset its initial cost through direct energy savings at the Rec Center, one must consider its indirect impact on visitors to the Rec Center. Each person that uses an EHFEM elliptical trainer gains a better understanding of the value of electrical energy.

Becoming cognizant of electrical generation at the gym may heighten awareness of electrical consumption at home, work, and school. This could result in lowered electrical consumption in and around San Luis Obispo, especially in Cal Poly's residence halls and academic buildings (where most of the users would come from). On top of that, the completed EHFEM implementation should catch the eye of donors that happen to visit the Rec Center, which could lead to monetary donations and sponsorships for Cal Poly.

Environmental Considerations

The modified elliptical trainers reduce net energy consumption at Cal Poly's Rec Center. The decreased energy demand translates to decreased energy generation, at power plants in California and other western states. This benefits the environment by lowering emissions from current power plants, reducing the need for new power plants, and quieting the call for new fuel procurement. These three avenues promote air and water quality, and the preservation of natural habitats.

This project strives to use as many local and ROHS-compliant components as possible. The project avoids toxic batteries, and promotes minimal use of materials. Ideally, the system will find another use once its life cycle comes to an end, but if components require disposal, they shall find a safe resting place.

Sustainability Considerations

To ensure a sustainable design, the trainers must conserve and generate enough electricity over their lifetimes to offset the energy embedded in their EHFEM modules. It may take many years for a trainer to generate enough electricity to offset this cost; however, the trainers could have a strong impact on Rec Center visitors. When students use the EHFEM elliptical trainers, they become more cognizant of the value of electrical energy. That awareness may bleed over into their daily lives, leading to lowered energy use and more environmentally-conscious behavior. In this way, the project promotes sustainability through two distinct avenues: technological improvements and behavioral change.

Sustainable design implies endless life-cycle. While this system may only see use with the Rec Center's elliptical trainers for ten years or less, the DC-DC converters and central inverter could certainly find use in other EHFEM applications afterward. If any DC-DC converter ever fails, this report includes in-depth details about its functionality, allowing students or technicians to repair the device rather than completely replace it. When the EHFEM elliptical trainers become obsolete, their parts can find reuse in other configurations. This design promotes integrity by maintaining a high value over the course of its lifetime. At the same time, it promotes sustainability through the reduction of unwanted waste.

Manufacturability Considerations

This project outfits each elliptical trainer with a DC-DC converter and a connection to the central inverter. Thus, any EHFEM: Forward Converters with a Central Inverter system requires

a single inverter, at least one DC-DC converter, and a set of cables connecting each DC-DC converter to the inverter. Each additional trainer calls for its own DC-DC converter, along with an additional set of cables.

To ensure safety in the fitness center, the cables should be housed in unobtrusive conduit, and the inverter should be mounted in a safe and secure location.

Ethical Considerations

One could argue that Cal Poly should compensate EHFEM users for their work, since the university sells the electricity they generate. However, exercisers use the machines because they desire an elliptical trainer workout. This project captures otherwise wasted energy and converts it into usable, grid-tied electricity without affecting the usual elliptical trainer experience. This nonintrusive process allows exercisers to continue their normal routine while reducing fuel consumption elsewhere.

Health and Safety Considerations

The EHFEM elliptical trainers promote safety at Cal Poly's Rec Center. The modifications do not increase the chance of burn, electrical shock, or injury to the user. The design houses a DC-DC converter under each elliptical trainer's plastic body, protecting it from spills. Floor conduit protects the cables connecting the DC-DC converters to the inverter, which minimizes cable wear and reduces the chance of users tripping. Finally, each section of the design fastens securely to a surface, further enhancing durability and safety.

Social and Political Considerations

This project raises awareness about the potential for distributed generation, and educates students about the value of electrical energy. The EHFEM products could inspire students to come up with their own ways of generating on-site electricity. When students pursue these ideas in the form of senior projects, Cal Poly inches closer toward a zero net energy campus. Students can also carry these ideas off-campus in their future endeavors, ideally leading to advancements in renewable energy and distributed generation.

APPENDIX B: PROJECT TIMELINE

APPENDIX C: CIRCUIT COMPONENTS INVOICE

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Customer P.O. No.	Master Tracker No.	
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	VI 8310	10-APR-11

Buyer Name	Ship Via	Customer No.	Terms	Order Date
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Line No.	Supplier Number Customer Part No. Description	Quantity Ordered	Quantity Shipped	Quantity Pending	Unit Price USD	Extended Price USD
1	771-BUK7613-75B118 QCLAMP NXP MOSFET Power US HTS: 8541.29.0075 ECCN:EAR99	4	4	0	1.420	5.68
2	771-BSP250135 M2 NXP MOSFET Small Signal US HTS: 8541.21.0075 ECCN:EAR99	4	4	0	0.430	1.72
3	576-P6SMB51CA TVS Littelfuse TVS Diodes - Transient US HTS: 8541.10.0080 ECCN:EAR99	4	4	0	0.410	1.64
4	771-BZX100A115 DBOOST NXP Zener Diodes US HTS: 8541.10.0050 ECCN:EAR99	4	4	0	0.250	1.00
5	588-FCSL90R002GER ROC Ohmite Current Sense Resistors - S US HTS: 8533.21.0090 ECCN:EAR99	4	4	0	1.200	4.80
6	542-2316-V-RC LOUT Bourns Power Inductors US HTS: 8504.50.4000 ECCN:EAR99	3	3	0	3.990	11.97
Merchandise		Handling	Freight	Tax	Continued	

Shipping Information

* IMPORTANT: In order to process your payment correctly you must include the reference number and the invoice number when you submit your payment. These are located at the top of this invoice.
To expedite your order, we upgraded it to ship via UPS 3 DAY SELECT at a total cost of \$16.96. We paid for the upgrade and only charged you the rate of UPS GROUND SERVICE.

Tracking Number(s) and Billed Weights

1Z7759451204801386	1 lb		

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Line No.	Supplier Number Customer Part No. Description	Quantity Ordered	Quantity Shipped	Quantity Pending	Unit Price USD	Extended Price USD
7	810-C3225X7S2A335M CIN, COUT (CERAMIC) TDK Multilayer Ceramic Capacitors US HTS: 8532.24.0020 ECCN:EAR99	12 COO:JP	12	0	1.650	19.80
8	667-EEV-FK1K471M CIN, COUT (ALUMINUM) Panasonic Aluminum Electrolytic Ca US HTS: 8532.22.0020 ECCN:EAR99	8 COO:JP	8	0	2.770	22.16
9	810-C1608X7R1H474K CTS, RESET EXTRAS TDK Multilayer Ceramic Capacitors US HTS: 8542.32.0040 ECCN:EAR99	6 COO:JP	6	0	0.350	2.10
10	581-08055C502KAT2A CREF, RESET EXTRAS AVX Multilayer Ceramic Capacitors US HTS: 8532.24.0020 ECCN:EAR99	6 COO:US	6	0	0.240	1.44
11	81-GRM1885C1H202JA1D CCOMP Murata Multilayer Ceramic Capacito US HTS: 8532.24.0020 ECCN:EAR99	4 COO:JP	4	0	0.110	0.44
12	771-BYT79-600,127 DOUT2 NXP Diodes (General Purpose, Power US HTS: 8541.10.0080 ECCN:EAR99	4 COO:CN	4	0	1.490	5.96
Merchandise		Handling	Freight	Tax	Continued	
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Line No.	Supplier Number Customer Part No. Description	Quantity Ordered	Quantity Shipped	Quantity Pending	Unit Price USD	Extended Price USD
13	771-BAS116H115 DSD NXP Diodes (General Purpose, Power US HTS: 8541.10.0070 ECCN:EAR99	4	4	0	0.050	0.20
14	621-DFLS1150-7 DMAIN Diodes Inc. Schottky (Diodes & Rec US HTS: 8541.10.0080 ECCN:EAR99	4	4	0	0.660	2.64
15	625-SS29-E3 DCLAMP Vishay Schottky (Diodes & Rectifie US HTS: 8541.10.0080 ECCN:EAR99	4	4	0	0.350	1.40
16	660-RK73H1ETTP3483F ROSC KOA Speer Thick Film Resistors - S US HTS: 8533.10.0042 ECCN:EAR99	3	3	0	0.060	0.18
17	290-237K-RC RDELAY, RBLANK? Xicon Thick Film Resistors - SMD US HTS: 8533.10.0042 ECCN:EAR99	8	8	0	0.050	0.40
18	292-178K-RC RSDL, RBLANK? Xicon Thick Film Resistors - SMD US HTS: 8533.10.0042 ECCN:EAR99	8	8	0	0.040	0.32
Merchandise		Handling		Freight	Tax	Continued
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Line No.	Supplier Number Customer Part No. Description	Quantity Ordered	Quantity Shipped	Quantity Pending	Unit Price USD	Extended Price USD
25	726-IPB039N10N3G QMAIN Infineon MOSFET Power US HTS: 8541.29.0095 ECCN:EAR99	5	5	0	3.240	16.20
26	810-CGA4J2X8R1H104K CRESET TDK Multilayer Ceramic Capacitors US HTS: 8532.24.0020 ECCN:EAR99	4	4	0	0.200	0.80
27	71-CRCW25122K74FKEG RSD2 Vishay Thick Film Resistors - SMD US HTS: 8533.21.0030 ECCN:EAR99	4	4	0	0.390	1.56
28	726-IDT04S60C DOUT1 Infineon Schottky (Diodes & Rectif US HTS: 8541.10.0080 ECCN:EAR99	3	0	3	2.010	0.00
Merchandise		Handling	Freight	Tax	Paid by Credit Card	
106.97		0.00	10.33	9.36	USD \$126.66	
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